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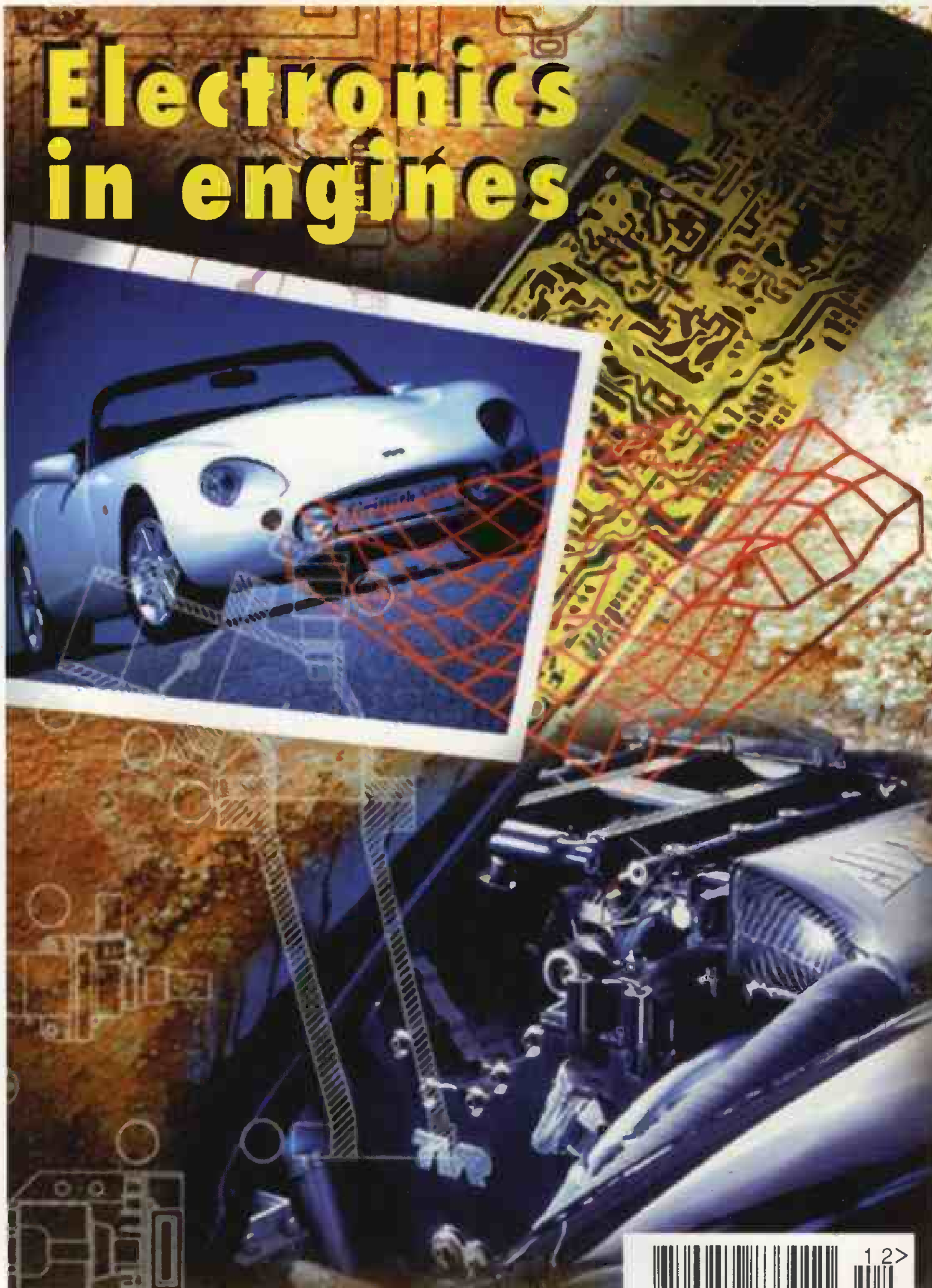

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CONTENTS

REGULARS

Next month: Orientation via magnetic sensing, Valve power, Duncan on CMR and cable modelling.

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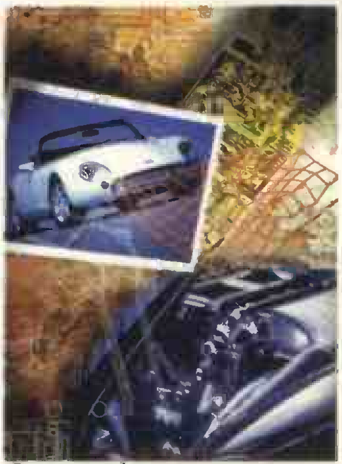

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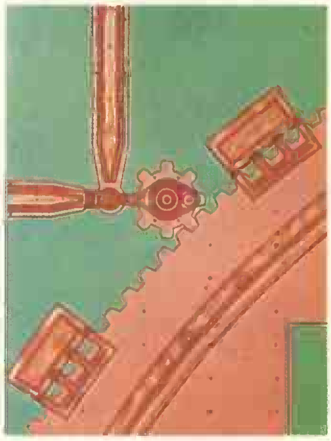

Micromotors ready for the high street? See page 1034.

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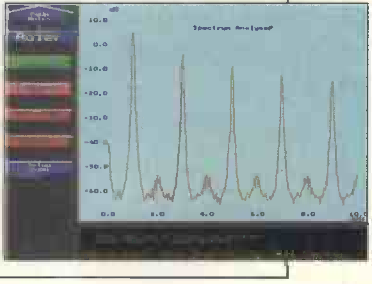

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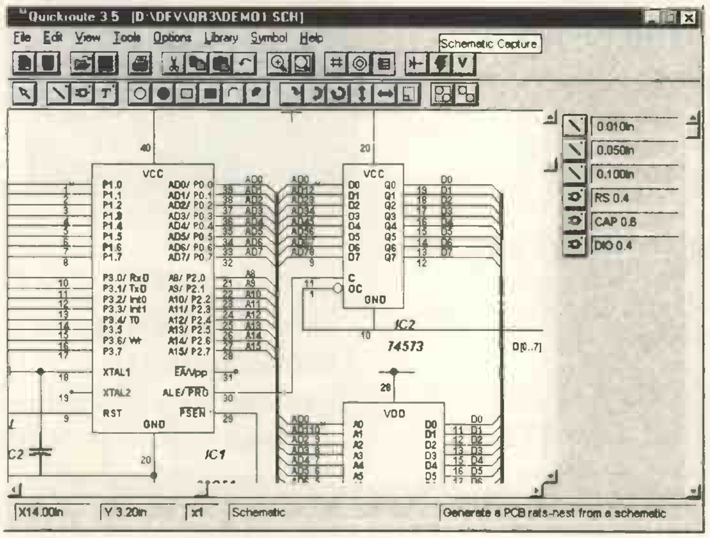

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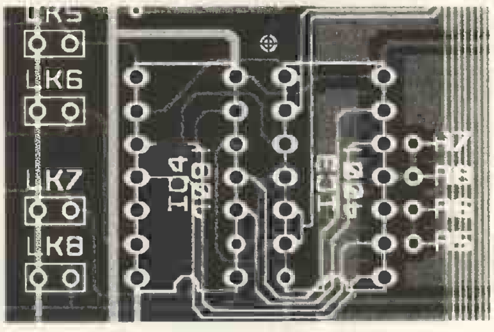

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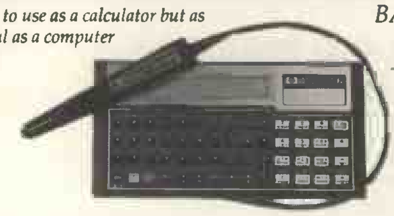

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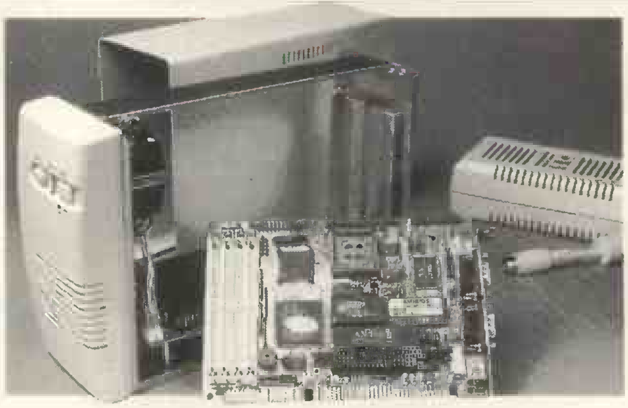

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Waiting for the mobile data train

No one would question the value of mobile communications. The mobile telephone has quickly become a commercial necessity for many business travellers and a vital security tool for the more vulnerable in society.

It is equally self-evident that the transmission of digital speech over the airwaves is little different from the transmission of data files. So why are we still waiting for the mobile phone operators to get there act together and start promoting mobile computing services?

The somewhat cynical view is that the mobile phone operators are reluctant to promote a mobile data which makes less efficient use of their networks. UK cellular phone operators Cellnet and Vodafone can already offer 9600bit/s data transmission on their digital GSM networks. But arguably they have no real incentive to heavily promote that capability while they are so preoccupied with building up their numbers of mobile phone subscribers.

The view of most mobile phone manufacturers is that voice traffic will dominate European networks for a long time yet, and in the view of one European mobile phone manufacturer, mobile data "is still seen as a Christmas tree decoration by many service providers."

One mobile phone supplier has been selling a PCMCIA data modem for its handsets since 1993. But the operators are unimpressed by its £400 price tag – to which the supplier responds "The operators are still only talking about voice."

While equipment makers blame the operators, who in turn point the finger at the developers of what they call "the necessary applications software"– whoever they might be. The growing number of laptop computer users must wait a little longer for a service which will allow them to communicate without wires.

One man who is far too astute not to recognise the absurdity of a world without mobile computing is Dr Andy Grove, who sits on top of Intel, the $13bn microprocessor giant. But without a suitable mobile communications network in the US, poor old Dr Grove had to come to Europe and its GSM digital cellular network to demonstrate the type of mobile data services he would like to see making money for Intel in the market.

"I am overjoyed about the existence of the GSM data technology," said Dr Grove. "It is a very significant development and very important for us." Point made.

The point is that the ability to transmit data at standard modem rates of 9.6, 14.4 and even 28.8kbit/s is available today. This isn't rocket science after all. So what is the problem?

...not having a suitable mobile communications network in the US, poor old Dr Grove had to come to Europe...

Part of the problem is the array of different radio protocols already used for mobile data services. As well as Cellnet and Vodafone the UK has around five radio data networks using at least three different protocols.

No longer is the transmission of digital data over radio a question of physics, but making it commercially attractive for programmers to sit down and compile the pages of computer code which will shape the services users want. From questions of interoperability between the various protocols to the design of graphical user interfaces, the mobile computing market is hamstrung by issues which were solved in the desktop market ten years ago.

It will take the commercial might of the PC industry to smash the logjam which has so paralysed the mobile phone community. Only now are there signs that the likes of Intel and Microsoft are looking for action in the mobile computing market. ■

Richard Wilson

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Will pass transistor logic make cmos obsolete?

Engineers at Hitachi have demonstrated marked savings in power, area and delay characteristics of circuits using its pass transistor logic (PTL). The firm sees PTL as an alternative to cmos for designing low power, ULSI ICs.

According to Yasuhiko Sasaki, at Hitachi's Central Research Lab in Tokyo, a PTL adder implementation is twice as fast as an equivalent cmos implementation, and yields a 40% area saving. "However, to date PTL has been limited to circuits of several hundred transistors only. For larger circuits it does not work well as it is difficult to synthesise," said Sasaki, speaking at this year's IEEE symposium on Low Power Electronics in California.

To exploit PTL for larger designs, Hitachi has extended its previous work and developed its 'multi-level pass transistor logic' (MPL). Its earlier PTL is based on single-level logic. Here source-drain inputs are connected to each other while gate inputs are driven by the circuit's primary inputs only. "This results in a lot of unshared logic making it difficult to exploit redundancy," said Sasaki.

In Hitachi's MPL approach, gate inputs are either driven by the primary inputs or from the outputs of other pass transistor circuits. This introduces a hierarchy in logic designs enabling circuitry to be shared. Delay is also reduced due to the more parallel circuit operation that results.

In experiments, Hitachi has used MPL to synthesise 27 random logic circuits selected from a microprocessor. "The effect of multi-level optimisation is clearly confirmed for circuits with a longer delay," said Sasaki.

On average total power was reduced by 23%, the area by 15%, and delay by 12%.

● The first commercial use of Hitachi's successor to CMOS technology – PASS Transistor Logic (PTL) – is due out next year according to Dr Tsugio Makimoto, executive managing director of Hitachi's chip business.

The first use of PTL in a commercial chip will be for the next generation of Hitachi's SH series of microprocessors – the SH4.

SH4 will show a dramatic five fold performance increase over the SH3 – an increase in power from the 60 MIPS SH3 to a 300 MIPS SH4. The SH4's designer, Toshimasa Kihara, said: "PTL is a kind of magic. It reduces transistor count, giving us a 20-30% improvement in die size, power consumption and speed.'

PTL is different from cmos in that, with cmos, transistors are charged by cmos and discharged by nmos, but in PTL they are both charged and discharged by nmos.

This reduces both the size and the number of transistors required, considerably saving silicon space.

Pocket video... NEC engineers have developed a pocket sized gadget, called Silicon View as a proof-of-concept for storing MPEG-compressed video and audio on solid-state storage. With a 40MB pc card, Silicon View can play back four minutes of video on its 312-by-230 resolution screen with the audio/video signal compressed 23-fold to 1.4Mbit/s. Silicon View employs NEC's MPEG-1 decoder IC and weighs 295g. However, the gadget is unlikely to be commercialised until 1Gbit drams become a reality early in the next century and will initially be expensive, requiring five such chips to provide 60 minutes of video playback. Silicon View follows on from a similar gadget called Silicon Audio which stored 26 minutes of audio on a flash pc-card.

Data-rate boost for mobile computing is imminent

Developers of digital mobile phone systems on both sides of the Atlantic are proposing new standards which will increase the data rates available for mobile computing on cellular telephone networks.

Nokia Mobile Phones has proposed to the European standards group, ETSI, a new data interface specification for the GSM mobile phone system which will support a 28.8kbit/s data rate. This is the equivalent of the V.34 wireline modem rate and three times the speed of the current 9600bit/s GSM data rate.

Qualcomm, developer of the CDMA digital radio protocol which will be used by mobile phone networks in the US and Korea, is planning to increase data throughput to 64kbit/s in 1997.

Like GSM, the CDMA system uses a 13kbit/s data .rate on the radio channel supporting voice and 144kbit/s data rates. According to Chris Simpson, Qualcomm's vice president of international sales, the same channel codec will support a 32kbit/s data rate in the 1.25MHz radio channel, which should be available sometime next year.

***Richard Wilson,** Electronics Weekly*

Chip set for CDPD

VLSI Technology has introduced its first chipset and software drivers for the cdpd data over cellular communications technology currently gaining favour in the US.

CDPD (cellular digital packet data) provides 14.4kbit/s data transmission over cellular telephone networks. In the US it is being deployed as an overlay to existing AMPS analogue cellular networks. The development of digital GSM data services has effectively stifled any European market for CDPD.

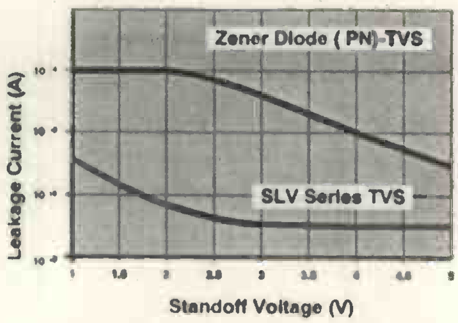
TVS breakthrough handles low voltages

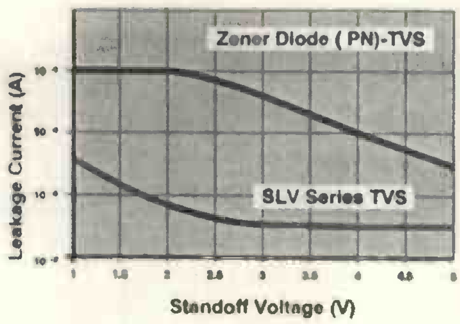
A new transient-voltage suppressor know as an enhanced punch-through diode is said to offer sharp knee voltages to well below 1V.

Developed jointly by Semtech and the University of California, the enhanced punch-through diode is fabricated using IC technology. It comprises a p+n+ zener in series with a pin diode to cut capacitance and leakage current.

The enhanced punch-through diode can be considered as a bipolar transistor with light base doping. This makes punch-through occur at a lower voltage than the conventional avalanche breakdown voltage between collector and emitter. The four layer n+p+p-n+ structure is responsible for the low clamping voltage, low capacitance and low leakage. The p+ and p layers are lightly doped to prevent the reverse-biased n+p+ junction from avalanching.

The first products to appear will have clamping levels of 4.3, 4.9 and 6.5V at 1A.





Leakage current and capacitance of the pnnp enhanced punch-through diode are orders of magnitude lower than typical pn zener diodes.

British GaAs lags

The UK is not putting enough effort into GaAs technology despite recent work of Middlesex University's Microelectronic Centre that promises low power, high speed GaAs-based static ram cells.

Dr Ebrahim Bushehri, deputy head of the centre and leader of the vlsi design group said: "Europe is putting more emphasis and spending a lot more money than the UK on high speed GaAs digital circuits and design methodologies."

Dr Bushehri stresses that GaAs is no longer an esoteric technology confined to research labs but a commercial reality. He cites Vittesse's 0.6µm mesfet GaAs technology capable of a million transistor devices as one example.

The centre's SRAM research work also involves collaborative work with the German Fraunhofer Institute for applied solid state physics, and uses high electron mobility transistors, hemts.

The research's motivation follows an idea of the group's for reducing power: "With existing SRAMs, the cell's cross-coupled inverters are used for storing data and driving the bit lines. For fast operation, the inverters need to be large to source and sink the bit-line's currents," said Dr Bushehri.

The centre's adopted design decouples the bit-line's driving inverters from the data storage. This enables the inverter size to be reduced, and the resulting standby current, saving on overall power dissipation.

Based on the centre's simulation work the hemt static ram cell has a standby current of 14µA and an active current of 0.29mA. This contrasts with the traditional six transistor direct coupled fet logic (dcft) cell currently used for GaAs that requires 570µA and 1.14mA currents.

Dr Bushehri said that at present the hemt approach is limited to 6000 gates due to yield problems. He believes however that this will soon be solved. The most natural application area for the work is for very high speed cache memory: "We are developing a design methodology that will enable very high performance circuits."

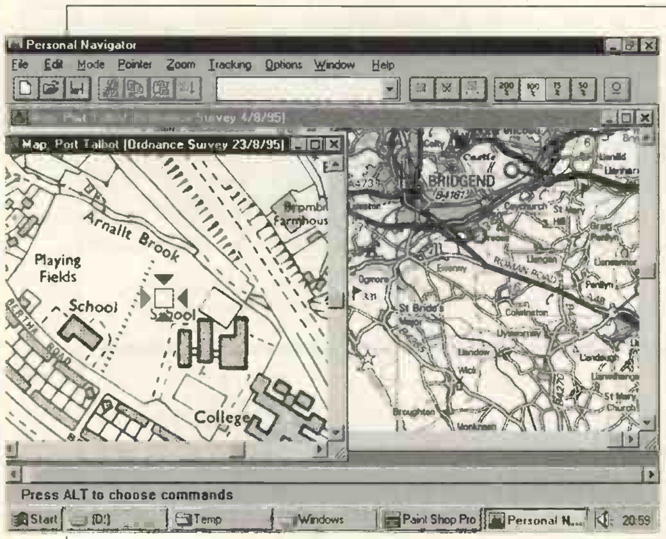
Roy Rubenstein*, EW*

Pentium power down

A 120MHz Pentium microprocessor aimed at portable computers with integrated power saving features to extend battery life has been released by Intel. The company is also set to unveil a new technology that brings together Internet and video communications.

At least ten major manufacturers of portable computers plan to make new models based on the new Pentium chip. The device features Intel's voltage reduction technology which reduces the inner core of the microprocessor to 2.9V while being used in a 3.3V environment. This can boost power savings by almost one third.

The new microprocessor costs $680 in 1000 unit quantities. Intel will also unveil a technology it calls Intercast which combines a cable tv or antenna based video link with an Internet connection. Users of pcs will need a regular Internet connection and a new PCI digital-analogue conversion card that will translate an incoming analogue signal into digital video. Intel will sell the card for about $150.



Where are you?

London-based software company Softwair unveiled its Personal Navigator device – based on global positioning systems, GPS, from Trimble – at the Motor Show in October. Suitable for use with portable computers, Navigator's software combines a cd-rom containing map data with GPS position information to pinpoint the user's location on the ordnance-survey map display. Companies such as Panasonic, Bosch and Alpine are looking into introducing similar devices within the next two years.

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RESEARCH NOTES

Jonathan Campbell

Green light for hdtv?

If the making of high-definition tv programmes is ever to become the norm rather than the exception there is a desperate need for a compact high-performance camera that is much lower in cost than anything currently available. Now, researchers at the NHK Science and Technical Research Laboratories in Tokyo think they may have the answer – a completely new image acquisition system that uses four ccds instead of the standard red green blue approach. This dramatically reduces the number of pixels required.

It is cheaper, smaller and lighter than normal hdtv cameras and a prototype was used successfully for news gathering following the Hanshin earthquake.

CCD imaging is increasingly popular for hdtv. But systems need approximately 2 million pixels to satisfy broadcast standards. Unfortunately, using a large number of pixels means the area of each pixel has to be small, narrowing the dynamic range, causing a drop in the aperture ratio, degrading highlight characteristics and lowered yield.

The NHK design uses a 16mm ccd camera – smaller than the usual 25mm format – with only 1.3million pixels. To compensate for the reduction, the researchers use a fourth ccd to boost the sensitivity of the system in the area where the eye is most sensitive.

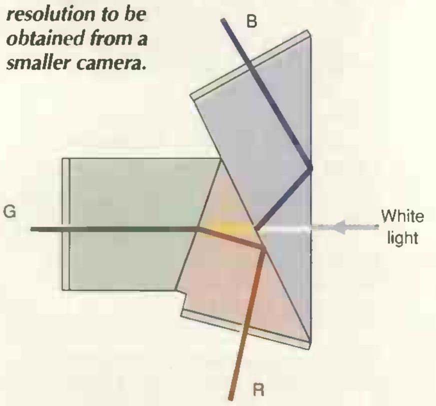
Instead of the normal rgb (Yoshihiro Fujita *et al*, *IEEE Transactions on Broadcasting*, Vol 41, No 2, pp76-82), two of the four ccds are assigned to green light, with other two being used for red and blue as normal. Putting more emphasis on the resolution of the green signal improves efficiency and makes the system compatible with the human eye which has a much greater sensitivity to luminance signals than chroma signals: 70% of hdtv luminance signals come from the green component.

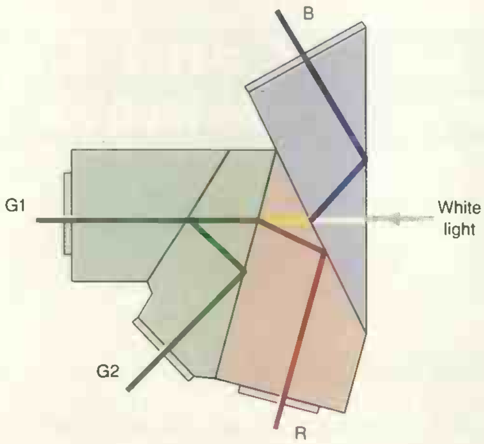
Colour separation in the four-ccd prism is essentially the same as with an rgb prism, with the addition of a half mirror to separate the green light into two portions. The resulting system meets resolution requirements as two 1.3 million pixel ccds – 2.6 million sampling points – are allocated to the green channel containing most of the luminance information, with 1.3 million pixels for each of the red and blue channels and the chroma information.

The prototype camera is reported to have fully satisfied hdtv requirements, demonstrating a resolution of 1200 tv lines, sensitivity F5 at 2000 lux and a dynamic range of 500%. Signal-to-noise ratio is 52dB. Size of the camera is 96mm by 250mm by 293mm, including the 16mm viewfinder, and it weighs in at 5.7kg.

Yoshihiro Fjuita, NHK Science and Technical Research Laboratories, 1-10-11 Kinuta, Setagaya-ku, Tokyo 157, Japan.

Emphasising the green component allows better resolution to be obtained from a smaller camera.





Cogs turn for commercial micromotors

Cheap and practical micromotors may soon be with us. This follows the news that researchers at Sandia National Laboratories in the US have used conventional microelectronic fabrication techniques to build a micro device that can drive external gearing.

Developing 0.5μW of power delivered through a gear 50μm in diameter, the motor could operate tiny micromedical pumps in drug delivery systems inside the body. It could also act as low-cost, high-performance gyroscope, having a dramatic impact on the design of future automobiles and military systems – both motor and gearing have much less mass than their macro-world counterparts and so can survive impact better.

So far, several hundred million rotations have been demonstrated by the smaller gears. But the breakthrough is in using etching processes and silicon materials already in use in the microelectronics industry to open up the possibility of mass production.

"We believe we are the first to demonstrate a really good silicon micromotor that can connect up with a variety of devices," says Jeff Sniegowski, the scientist who – with engineer Ernest Garcia and group leader Paul McWhorter – has led the effort to build the millimetre-square engine and its even tinier gearing.

Sandia's construction method actually extends a technique first developed at the University of California at Berkeley. The basic batch process – which, when perfected, should leave behind thousands of fully assembled, operational microengines – begins on a silicon substrate.

Researchers deposit a layer of electrically insulating material and then a film of polycrystalline silicon, patterned to form electrically conducting lead-ins.

On top of this, a film of sacrificial silicon dioxide serves as a support layer as the remainder of the structure is built. When it is removed, by several etching processes, openings through the oxide allow the next applied layer of polysilicon to anchor to the insulating layer on the substrate. This is how the vertical axles for gears and elastic supports for the engine are formed. Other layerings and subsequent removals of the oxide free the gears and linkages.

During the process, silicon nitride is added, functioning as a grease to let the gears turn more freely.

As a final step, hydrofluoric acid is added to remove all the sacrificial supporting layers of silicon dioxide.

The final motor consists of two tiny silicon combs with a shuttle placed between them. The edges of the shuttle also form combs with teeth that interdigitate with those of the stationary combs.

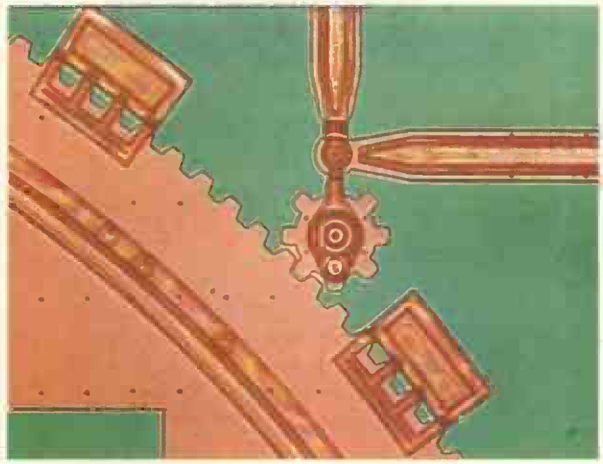
Applying an on-off voltage to energise the stationary combs alternately, pulls the shuttle by electrostatic attraction so that an attached shaft will turn a drive gear in a quarter of a circle during the shaft's power stroke.

Another comb-drive engine, at right angles to the first, is timed to turn the gear on the second quarter of its rotation. The two drives, alternating their force, turn reciprocating motion into rotary motion to drive the gear completely around.

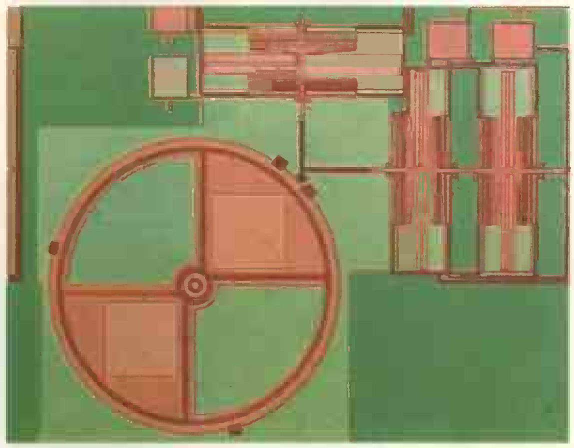
Electronic circuits not part of the micromotor chip drive the motor.

Sandia researchers are currently working to place control circuitry next to the microengine, and to develop a single chip with circuits and machines fabricated side by side.

Paul McWhorter, Sandia National Laboratories, Albuquerque, New Mexico, 87185-0167, USA.
Paul_I_McWhorter%smtplink.mdl.sandia.gov@sass165.sandia.gov



Close up of the output gear of Sandia's microengine.



Sandia has used microelectronic techniques to build a microengine that could be mass-produced. The output drive gear is 50μm in diameter.

Chemical weapon – or just a bad egg?

An easy-to-use transducer and receiver system that establishes the acoustic response characteristics of a container and its contents is being used as a non-invasive alternative to normal analysis of chemical weapons.

The (ars) system, developed at the Los Alamos National Laboratory in the US, could also prove useful for much more peaceful applications such as detecting salmonella in eggs or measuring intraocular pressure.

Acoustic resonance spectroscopy is a non-invasive system that uses a sensor head with two transducers attached magnetically to the container being tested. One transducer induces minute vibrations in the container while the other detects the resonance frequencies at which the container naturally vibrates. The pattern of the vibration frequencies is affected by the physical properties of the contents and can be used as an acoustic signature.

Traditional methods of verifying the contents of chemical munitions require a hole to be drilled into the container and a sample of the fill extracted for laboratory analysis. It is time-consuming and has the potential to contaminate the environment and expose workers to nerve gas or other chemical agents.

But, by measuring the acoustic vibrations of an object, the Los Alamos instrument quickly and safely identifies the fill content of chemical weapons or other containers holding toxic substances. To identify the chemical fill, vibration patterns can be matched with signatures in a library, and the entire procedure takes less than a minute and the operator is never exposed to the chemical contents of the container.

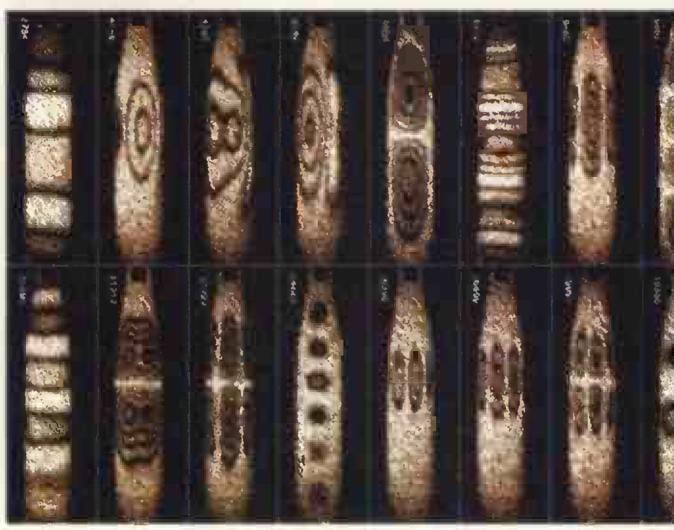
Measuring the vibrational modes of objects is a well-established technology. But using acoustic signatures to identify fill materials and the software algorithms that implement this identification put the instrument ahead of traditional technologies.

Originally developed with the Defense Nuclear Agency as a non-invasive inspection tool to verify compliance with treaties on chemical weapons destruction, the detector could prove suitable for any non-invasive identification of fill materials in sealed containers.

Los Alamos says the technique could be extended to quality control applications too, in which defective parts have a different acoustic signature than their good counterparts.

Paul Lewis, Los Alamos National Laboratory, California, USA.
lewis@lanl.gov

Video holograms showing some of the vibrational resonance modes of a 105mm munition. Each 'contour' corresponds to 0.5μm.



Picturing the sun's magnetic field

Physicists at Nasa have just released the first instantaneous view of the spiral structure of the solar system's magnetic field. The picture shows how the lines of magnetic force originate in the Sun and extend outward into the solar system.

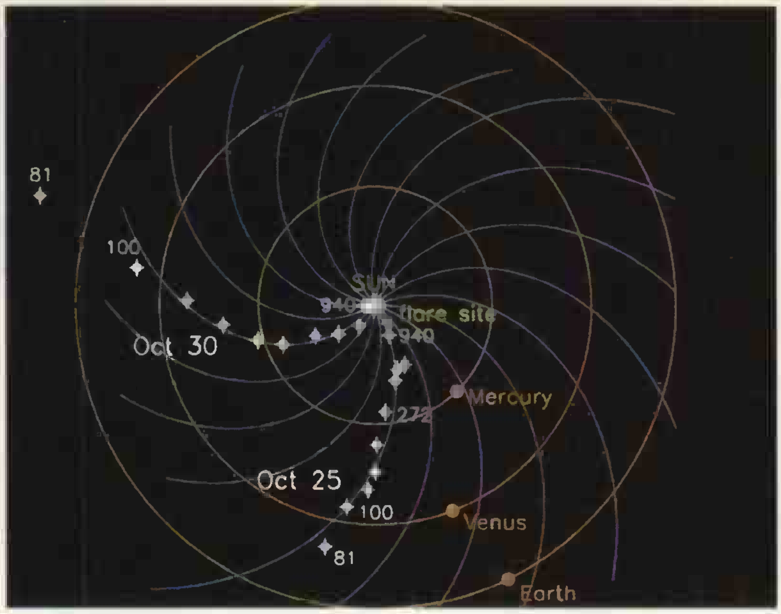
Nasa's snap-shot, assembled from observations of radio waves by a US-French radio receiver on the Ulysses spacecraft, shows the spiral magnetic field extending from the Sun past the orbit of the planet Venus toward the orbit of Earth.

From its vantage point over the south pole of the Sun in 1994, Ulysses was able to track the path of the bright spot of radio waves excited by moving electrons ejected from the Sun at speeds over 62,100mile/h.

Radio emissions – caused by the fast electrons moving through with the slower solar wind – allow the magnetic lines of force to be traced out in a way similar to deducing the course of a road at night from an airplane by tracking the headlights of individual cars.

A chart of the received radio emissions shows that they follow the expected spiral shape, even including the kinks due to variations in solar wind speed.

Previous radio observations made by space probes orbiting in or near the plane of the Earth's orbit did not provide a good vantage point for observing the spiral shape of the magnetic field. Observations in space are required because the radio frequencies of the solar wind do not get through the Earth's ionosphere.
Goddard Space Flight Center, Greenbelt, MD, USA.

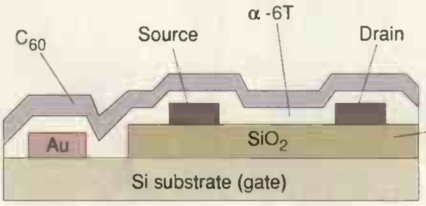


View from the north ecliptic pole is based on Ulysses radio measurements made as it was passing over the Sun's south pole. The white symbols represent the actual observations of the location of outward moving streams of electrons, ejected from the Sun on October 25 and October 30, 1994. The numbers indicate the frequency of radio emission, so that '940' represents emission at a radio frequency of 940kHz.

Mercury, Venus, and Earth are shown in their approximate true positions at the time of observation – the large orange circles illustrate their orbits around the Sun. A yellow arrow points out the location of the Sun where a solar flare explosion on October 25, 1994 ejected the electrons tracked by Ulysses on that date. The spiral blue lines illustrate the shape of the magnetic field as predicted from theory for a constant solar wind speed.

Organics breakthrough into low power devices

Structure of the heterojunction TFT with alpha-6T and C_{60} active layers that could open up production of low power organic complementary circuits.



Work on thin film organic transistors has produced several p-channel versions based on one combination of materials, and n-channel versions based on others. But researchers at AT&T Bell Laboratories have now announced development of an organic transistor structure with two active materials that permits both p-channel and n-channel operation in a single device.

For the first time, fabrication of complementary circuits – with all their well-known advantages in terms of power dissipation and device lifetime – could now be possible with organic technology.

First active layer of the device, adjacent to the gate dielectric (A Dodabalapur *et al*, 'Organic Heterostructure Field-Effect Transistors', *Science*, Vol 269, pp.1560-1562) is alpha-6T, a thiophene oligomer which has been used in p-channel devices. This layer is about 10-20nm thick. Second layer, for n-channel operation, is C_{60} and is about 20-40nm thick. A third electrically inactive layer is deposited on top of the C_{60} to protect it.

Energy levels of the highest-occupied and lowest-occupied molecular orbitals of the two materials is such that when the gate is biased negatively with respect to the source, the p-channel material is filled with holes; and when the gate is biased negatively, the n-channel material is filled with electrons.

Other experiments have since been successfully carried out with alternatives to alpha-6T and C_{60}, the only requirement being that the materials have similar highest-occupied and lowest-occupied molecular orbitals.

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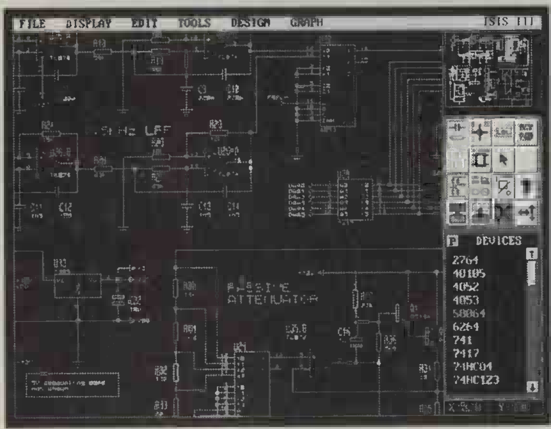

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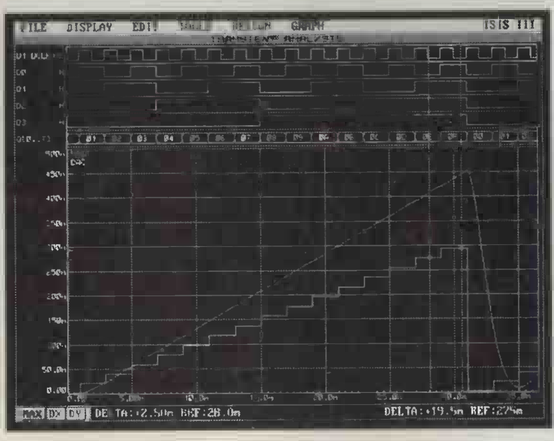

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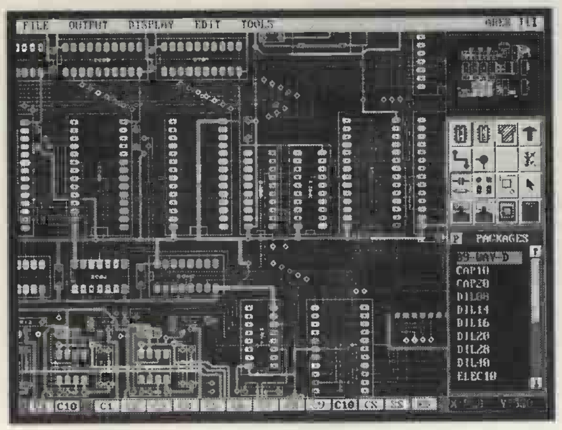
- Partitioned simulation of large designs with multiple analogue & digital sections.
- Graphs displayed directly on the schematic.



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CIRCLE NO. 111 ON REPLY CARD

Audio preamp – *simple but sound*

Reg Williamson's preamplifier is possibly the first to use a generalised impedance converter. This reactive block helps keep the design simple and transparent to the audio signal passing through.

Designed to be as simple as possible, this circuit is transparent and provides two basic functions – equalisation and gain control.

A universal interface for almost any programme source and following power amplifier is also provided. Inclusion of an equaliser was an attempt to disprove that even a properly designed equaliser would always subjectively degrade signal quality in some way.

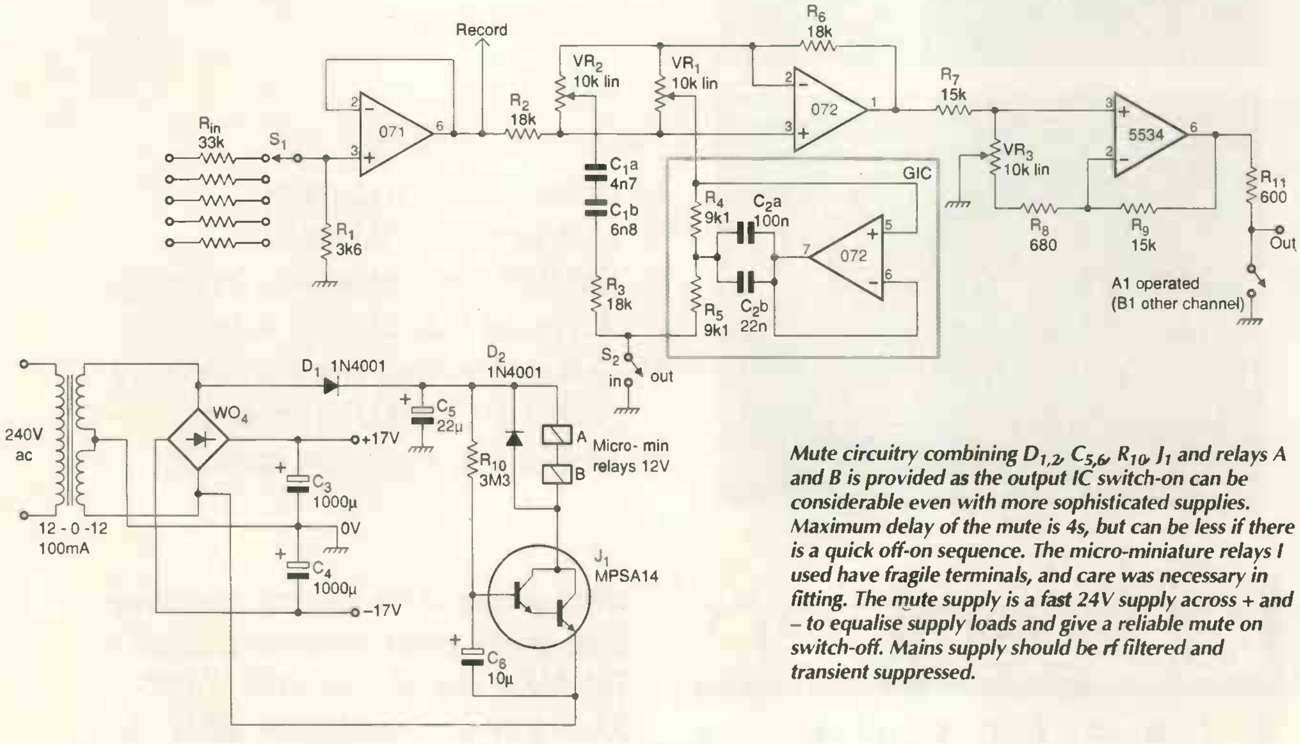
Distortion plus noise is less than 0.05% at 1kHz and 3.5V rms with the gain controls at maximum – and these results are from a source with an inherent 0.03% thd+noise figure. Maximum output swing is 10V rms.

Equalisation and gain control

Despite its apparent simplicity, the design is wholly innovative in concept. It offers flexibility of overall gain, modest equalisation of high and low ends of the audio spectrum around a central point of 1kHz, and gain control.

The whole system is transparent, with no phase inversion – the *bête noir* of the purists. When set to the electrical centres of their respective controls, the reactive elements of the equalisers are virtually out of circuit. With these settings, the circuit behaves as a unity gain amplifier with 100% negative feedback.

Even so, reactive elements can be switched out altogether with a single double pole on/off



Mute circuitry combining $D_{1,2}$, $C_{5,6}$, R_{10}, J_1 and relays A and B is provided as the output IC switch-on can be considerable even with more sophisticated supplies. Maximum delay of the mute is 4s, but can be less if there is a quick off-on sequence. The micro-miniature relays I used have fragile terminals, and care was necessary in fitting. The mute supply is a fast 24V supply across + and – to equalise supply loads and give a reliable mute on switch-off. Mains supply should be rf filtered and transient suppressed.

Does component price make a difference?

Listening tests were carried out involving 31 people with a high quality programme source of varied music material. Some of these were music enthusiasts from the Recorded Music Society, but a few were 'professional listeners'.

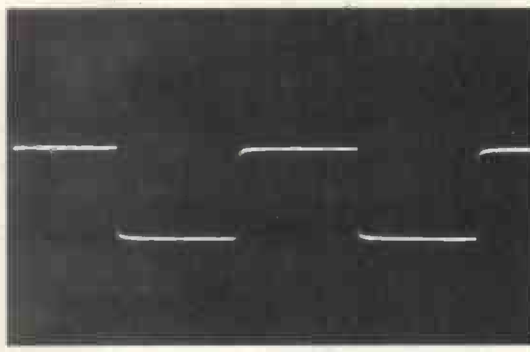
The idea was to make two identical preamplifiers. One was built using off-the-peg components and the other had highly expensive alternative components such as super-capacitors and resistors, and even gold plated sockets.

A comparative switching system was also provided to allow the listener to switch randomly between the original programme source and either of the two preamplifier outputs. High quality electrostatic headphones provided the audio output.

In all cases, individual gains were matched at 1kHz within 0.2dB and any equalisation controls set to a measured linear position. The listener did not know which output was which.

The tests demonstrated what we expected – that hardly anyone could tell any difference between all three sources. Of the 341 steps in the test, there were 248 opportunities to detect the inclusion of the preamplifier in the audio chain. Only 24 of these were noticed, and even then, the differences were described as 'slight' or 'doubtful'.

Besides confirming the amplifier's transparency, these tests also indicate that exotic components – whether passive and active – are a waste of money.



Response of the preamplifier subjected to a 5kHz squarewave at 6V pk-pk amplitude. Output loading was the quasi-standard of 50kΩ.

switch. In either the midband nominally linear setting or the switched-out position, overall gain of the pre-amplifier is as originally selected by the constructor and is totally flat. Reactive elements for the low end include for the first time, a generalised impedance converter, which simulates an inductor in series with a resistor.

The whole is flexible in design, allowing for variation. However in this instance, circuit parameters selected result in the controls having a shelving action, limiting equalisation to 6dB. Because the same reactive elements are used for both attenuate and accentuate functions, there is absolute symmetry of the two respective curves.

Negative feedback gain control

Gain control adopted is also unusual, being a negative feedback type. For high gain settings, and low signal level inputs, feedback is reduced – for low gain and high level signal inputs, it is increased. While using a linear control, the action is sensibly logarithmic and unlike some other versions, has a genuine zero gain position fully anticlockwise. It contributes a maximum gain of 20dB and has a very low output impedance of 600Ω.

It could be argued that a balance control is redundant these days, but one can be fitted. However, it is easier to fit a concentric gain control which is, fortunately, still available.

To complete the flexible interface requirements, the input is a unity gain buffer stage without phase inversion and 100% negative feedback. Input signals of differing mean levels can be accommodated by a passive L pad at the input. Distortion and noise are negligible and maximum peak-to-peak voltage swing is far in excess of normal requirements. ■

Further reading

Watling, Alan, 'Golden Ears?', *Hi-Fi News*, May 1994.

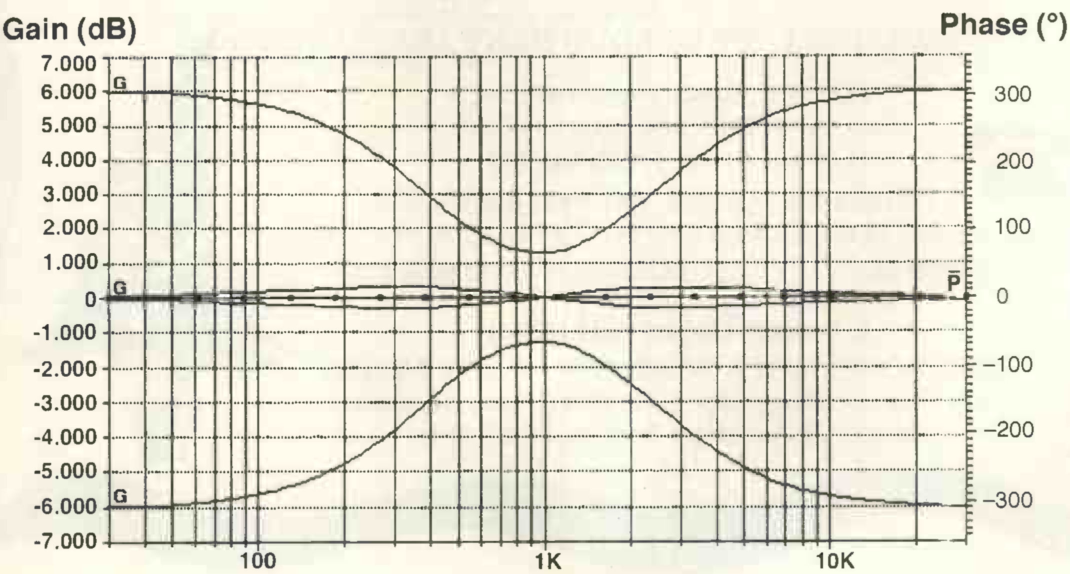
A tribute to Peter Baxandall

This article is offered in tribute to the late Peter Baxandall, my friend and technical mentor for almost 40 years. His observations and evaluation of my design ideas I appreciated highly.

This design was the last valuable service he did for me. With modest deference, I thought my equaliser an ideal replacement for his own 43 year old brain-child that rightly bears his name worldwide. It was with characteristic generosity that he gave my alternative his warm and unstinting approval – accompanied by three A4 pages of detailed analysis.

Peter never failed to find some useful facet of the design procedure that I had overlooked. For virtually half a century, his unique abilities contributed to the art and science of audio engineering.

He will be greatly missed.



Simulation of boost and cut curves with the preamplifier's tone controls set flat and at maximum and minimum. Centre curves show phase shift. Measured results tally.

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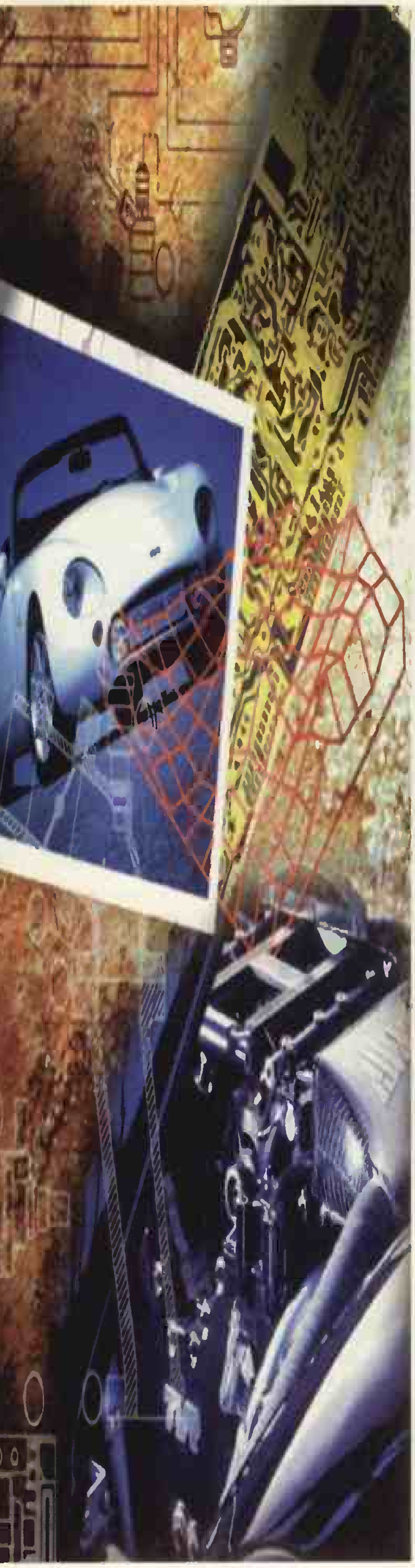


Engine *management developments*

Eric Russel looks at how electronic engine management systems developed for the race track are affecting performance in the latest cars intended for the nation's highways.

Computerised engine management systems for cars are becoming more and more popular, and as the number on the road rises so do the horror stories. There is no doubt that such systems are effective, but when problems arise, the support leaves a lot to be desired. Stories are already emerging of out-of-warranty charges of £600 for a replacement control unit which, in one case, left the car with its original problem.

Components are comparatively inexpensive and there could be an opportunity for electronics entrepreneurs to look at repairs while garages remain mechanically orientated and equipment manufacturers do not want to repair units.

Increased engine efficiency

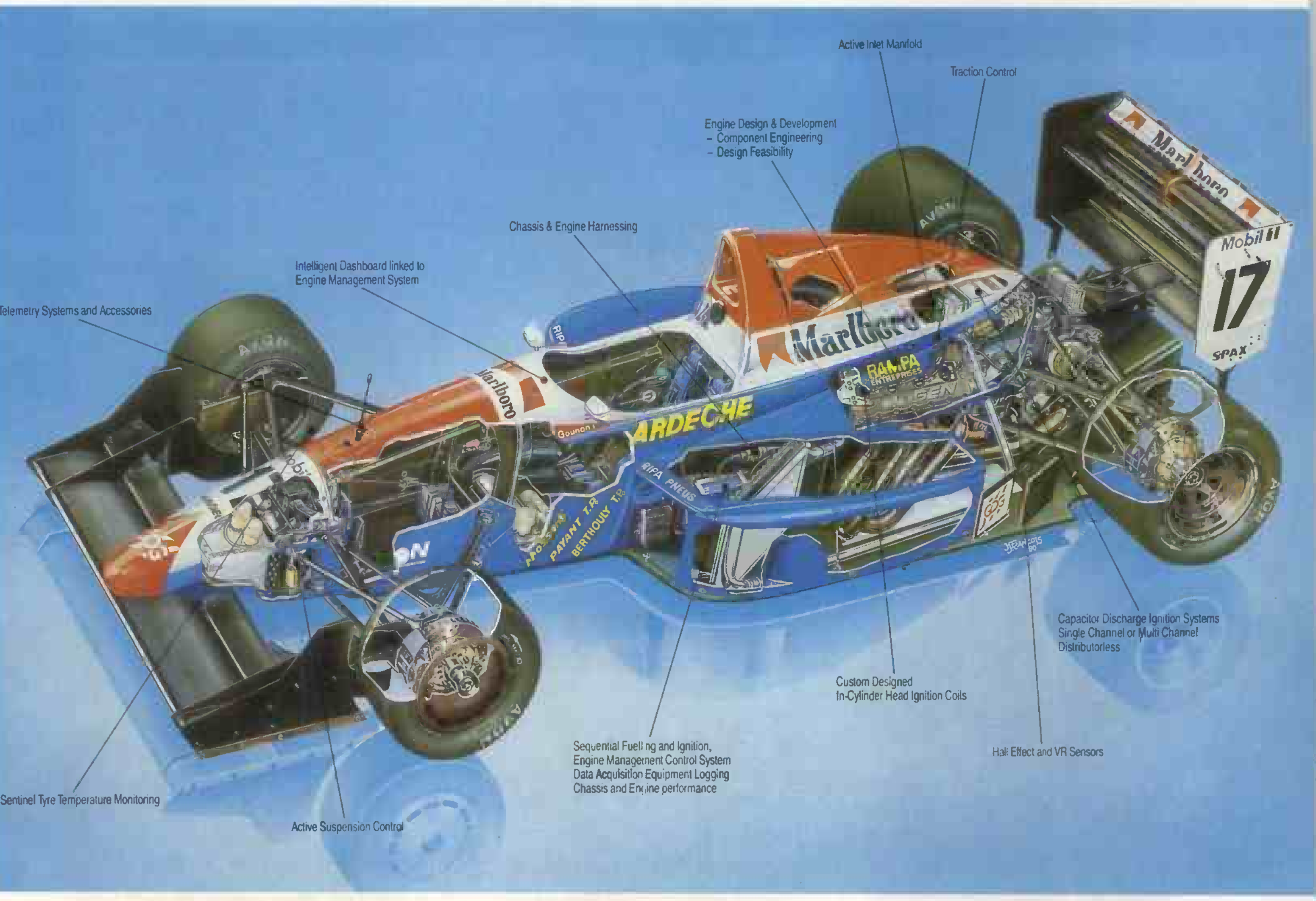
Although the principal driving force behind the increased popularity of EMS is pollution legislation, there are other reasons for its use. While noxious emissions can be cut by 10% and fuel consumption improved, driveability is also improved by minimising engine speed oscillations during acceleration, for example.

Idle quality is also enhanced because tick-over is more tightly controlled. The Lucas Epic engine management system, for example, has a special software routine called adaptive cylinder balancing which compensates for mechanical tolerance variations and wear to ensure smooth running.

Three dimensional profiling

Computerisation has added an extra dimension to engine control. Whereas mechanical linkages can be regarded as two dimensional, with a direct relationship between throttle position and fuel demand, electronic systems have a three dimensional mapping ability which gives a greater range of options. Instead of following points on a flat graph, engine control follows a contoured 3D profile, illustrations of which shown are on page 1019.

Because the 3D map is more detailed it has to be created individually to suit driver, engine and the driving environment. This takes several hours while engine settings are noted, on a rolling road, for example, for a set number of engine speeds. These points are stored in



As this racing car cutout from Zytek shows, above, there are many electronic links between engine and the rest of the car. This is also the trend in road cars, which will soon have more electronics than mechanical systems. Ride, road holding and performance are now all coming under electronic control and it is the current breed of low cost processors which makes this possible. Microprocessor speeds far exceed mechanical response times so it is easy to control suspension, braking and engine in less time than it takes a wheel partially to rotate or for an engine to complete one revolution.

memory and accessed during a journey, with modifications to the engine controls from sensors around the engine.

With the Bosch Motronic EMS, engine load determines how the electronic control unit, ECU, reacts and is computed from air mass flow and throttle position. Air flow sensors are fitted in the air inlet and consist of a flap attached to a potentiometer. As air flow increases the flap rotates and the changing resistance in the potentiometer is measured by the ECU. An air temperature sensor is also connected to the ECU which can then calculate the volume of air entering the engine.

Hot wire or hot film meters can be used to provide an index for mass air flow. A heated element in the air stream is cooled by the air and the heating current needed to return the element temperature to a pre-set value, as measured by an integral sensor, provides the data. The system automatically compensates for variations in air density, a factor that determines the amount of warmth the air absorbs from the heated element.

A 'knock' sensor can also be fitted to an engine to cut pre-ignition problems. Knocking occurs when fresh mixture in the cylinder

Aided by comprehensive and interactive electronic systems, this 4.2l V8 engine – the TVR AJP 8 – can develop 350bhp at 6500 rev/min.

ignites spontaneously before the controlled flame front arrives from the spark plug. This creates a flame velocity some seventy times faster than normal and the resulting high pressure shock wave is picked up by a knock sensor which again feeds back to the ECU. Pre-ignition causes thermal stresses which can result in mechanical damage.

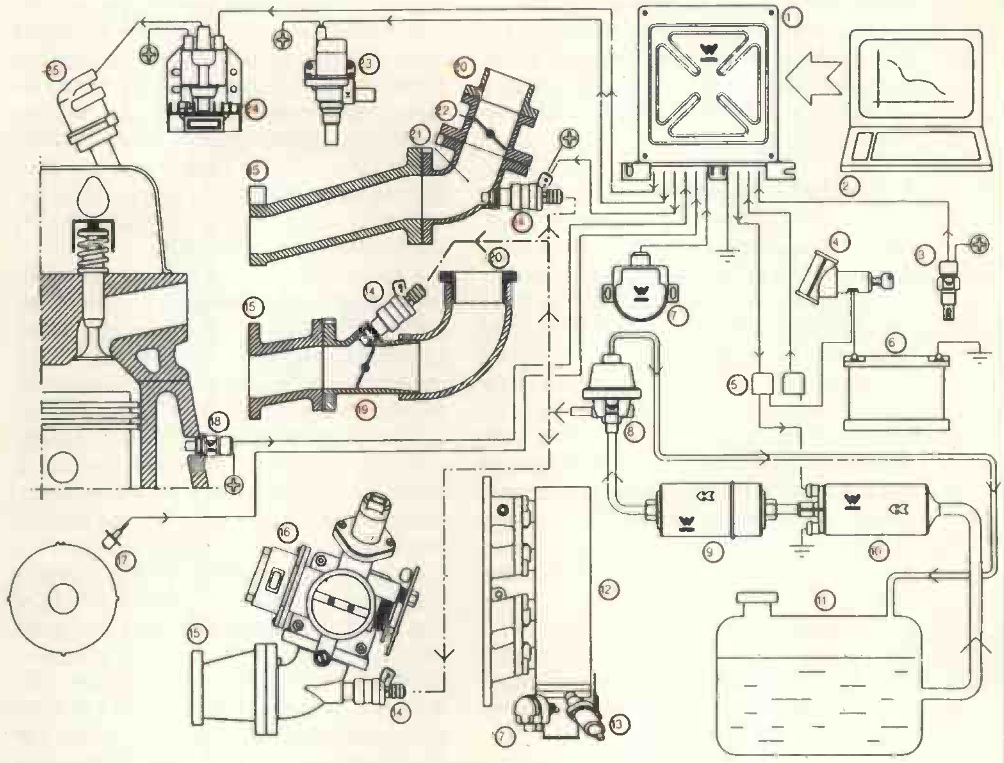
In the exhaust system a lambda oxygen sensor monitors excess air in the exhaust and feeds back to the ECU which then controls the air to fuel ratio. To complete the monitoring, battery voltage has to be sensed as it affects the operating times of devices such as electromagnetic fuel injectors.

Integrating injection

After a century or so of mechanically operated carburettors, electronically controlled fuel injection seems set to take over as part of a complete concept – computerised engine management. The accent is on software. The engine department is comprehensively monitored and the computerised control of air, fuel flow and ignition timing means optimum combustion.

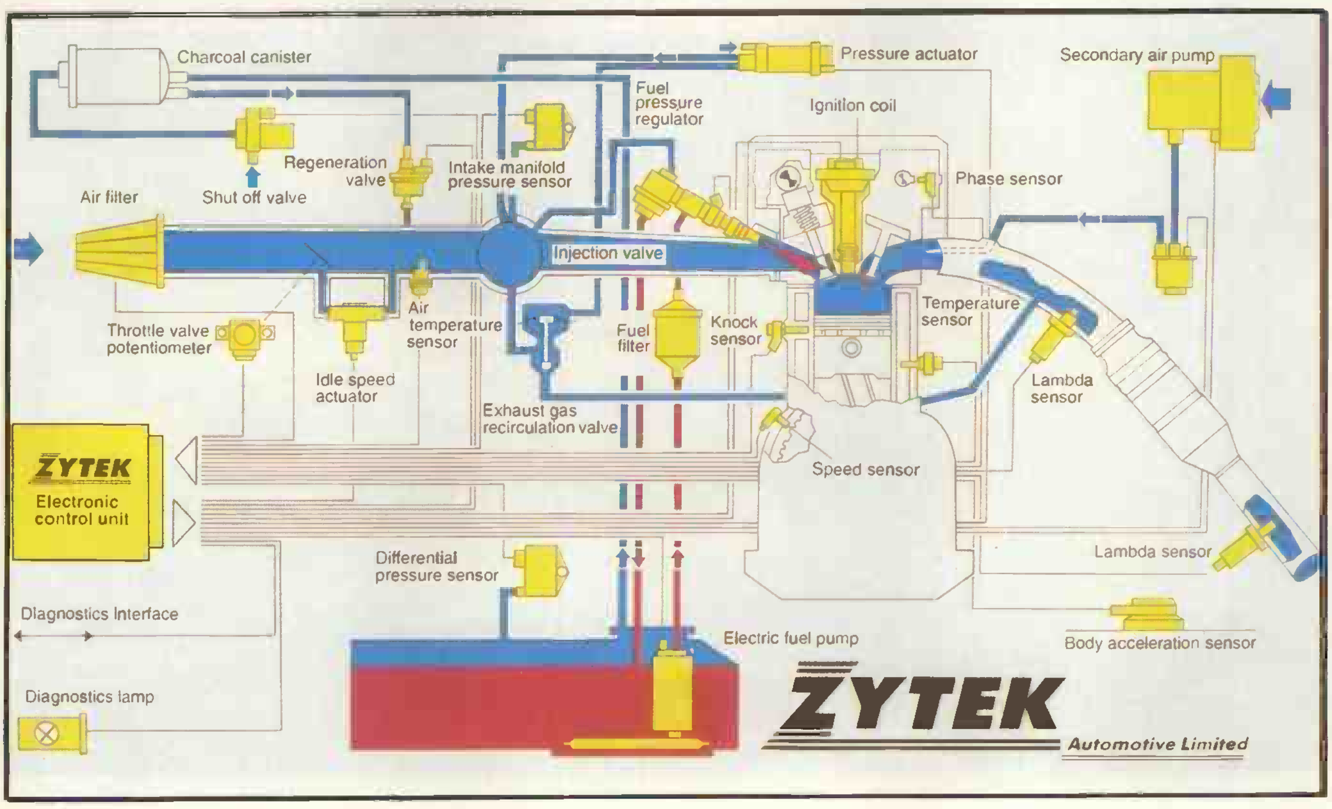
Optimisation provides the best combination of power, fuel consumption and exhaust emission for any given driving situation. It also means the smoothest idling and maximum engine efficiency during cold running. Systems respond to varying driving and atmospheric conditions as a journey progresses in contrast to mechanical adjustments which are fixed beforehand.

Engine management is the latest development in a trend which began with transistorised ignition in the early 1970s and has been helped by the lower costs of fuel injection due to volume sales. Motor sport's con-



Weber's engine-management system takes into account a wide variety of engine parameters from fuel flow to water temperature. The result is faster engine response, more efficient combustion and fewer noxious emissions.

Zytek's EMS3 has a 32bit Motorola processor allowing dynamic modifications to engine maps. Watertight housing is critical to successful microprocessor operation. Damp, dirt, dust, condensation and extremes of temperature are all enemies of electronics but are the standard environment for car computers. Vibration can also affect in time, pcb mountings, connectors and soldered joints. But car manufacturers demand lightweight control units – an additional problem for designers.



tinual quest to shave fractions of a second off lap times has helped drive development and that harsh environment has ensured robust products. The result has been a radical rethink of engine control, which is now spinning-off into volume cars.

Yet higher integration

One of the leading companies in the field, Weber Concessionaires, has taken EMS a step further with a system that has even greater flexibility and versatility – Alpha Plus. It is a development of the Alpha Engine Management System that is specified, amongst others, by Opel Racing on all their Class 2 touring cars, including the Vauxhall Dealer Sport team, Aston Martin and Caterham for their JE special. John Clelland has just won the British Touring Car championship for Vauxhall using an Alpha system. He clinched the title with two races in hand which indicates how far ahead the Vauxhall was.

A complete Alpha Plus kit can include – fuel injectors, engine speed sensor and trigger disc, high-resolution throttle position sensor, inlet air temperature sensor, barometric pressure sensor, water temperature sensor, lambda sensor, electronic control unit with integrated ignition amplifier and option of static or capacitive ignition system, cold start extra air valve, fuel pressure regulator, high pressure fuel pump, throttle bodies, wiring harness, inertia safety switch, ignition coil, fuel filter and connector unions, fuel hoses, manifolds, inlet air trunking, air filter mounting flanges and adjustable throttle linkage.

It can monitor external air temperature, turbo-boosted or barometric air pressure, water temperature, throttle angle, engine speed, camshaft timing, gear selection, idle setting and battery voltage. The on-board computer in the ECU processes all these inputs and can control injector opening, spark position, camshaft timing and variable inlet track length. The result is faster engine response, more efficient combustion and less noxious emissions.

To convey engine speed, the trigger disc is bolted to either flywheel or pulley. It is a flat ring with teeth on the outside, one per cylinder and one to indicate top dead centre, and has to be fitted with a concentricity of ±0.1mm. This accuracy, combined with the ring's low mass in comparison to crankshaft mass, means that no re-balancing is required. The speed sensor is a proximity type and generates a pulse every time a tooth passes. A similar sensor is required on the camshaft to identify cylinder status.

A potentiometer fixed to the end of the throttle spindle conveys driver action on the accelerator immediately to the ECU which controls engine devices through robust power transistors. These buffer the pcb components from such problems as transient spikes on the vehicle electrics.

Memory for more torque

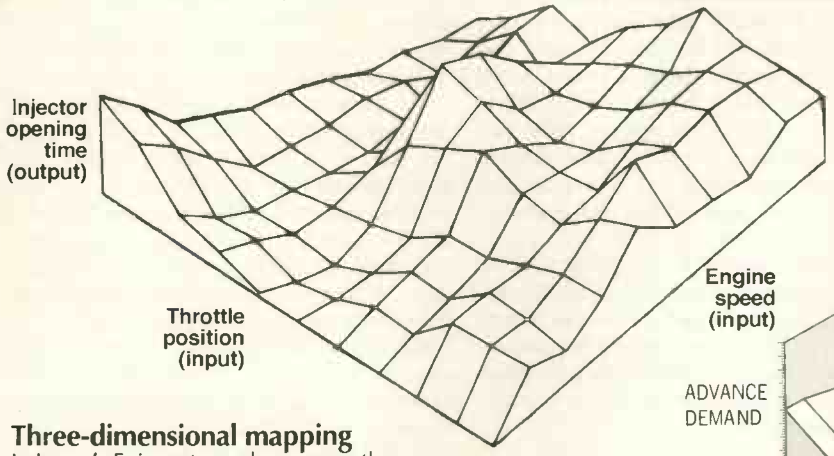
The microprocessor board also holds a memory chip which contains the calibration settings. These are the optimum settings for a range of driving conditions – up to 256 different combinations of engine speed and throttle position in the case of Alpha Plus. The result is a wider power curve, which provides more torque at low and high revs, and an increase in mid-range performance.

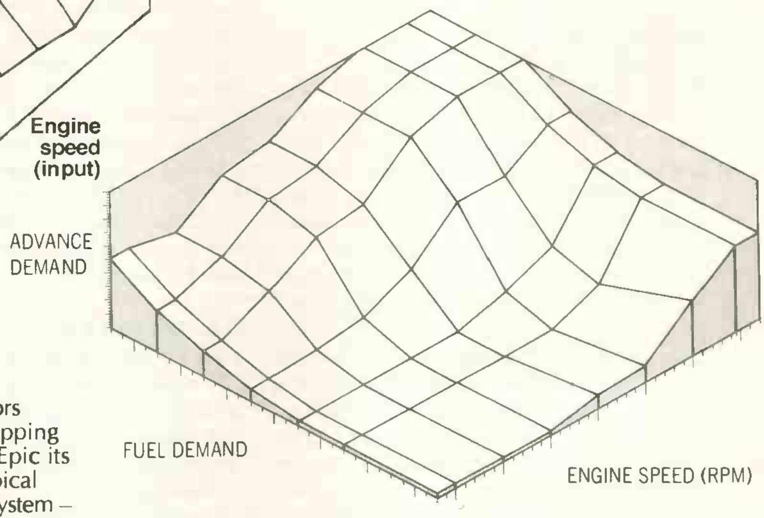
Alpha senses engine load from the throttle position sensor rather than monitoring air flow. This means an immediate response to changed throttle settings because there is no delay while air pressure changes. It is also felt that sensors in the air flow create turbulence and affect volumetric efficiency of the engine, reducing the benefits of EMS.

There is no one combination of settings to suit all drivers so an ECU has to be trained for the vehicle it works with. First, in a calibration session that takes several hours, an engineer feeds into a portable computer up to 256 settings of ignition, air and fuel against a range of throttle positions and engine speeds.

For calibration, either the engine is mounted on a test bed or the car is run on a rolling road where wheel speed, torque, bhp and exhaust emissions are monitored. The application

John Clelland's Vauxhall Dealer Sport team car – featuring Alpha Plus engine management – wins the British Touring Car Championship 1995.





Three-dimensional mapping
In Lucas's Epic system, above map, the electronic control unit converts signals from the various vehicle, engine and pump sensors to digital form. They are then used as inputs to a number of maps stored in memory. The control unit then sends signals to the pump actuators to implement the desired control strategy. Multi-dimensional mapping and accurate transient control of the electronic system that give Epic its prime edge over electro-mechanical systems. Right is another typical three-dimensional map created by Weber Concessionaire's Alpha system – discussed in the main text.

engineer runs the engine at a series of set speeds and tunes ignition timing and fuel injection using Weber Concessionaires' own software to store settings in a portable computer. These settings are logged against throttle positions and engine speeds.

Once the overall engine performance meets the customer's requirements, the settings are transferred to a memory chip, generally erasable-prom. This is then plugged into the EMS computer on the car. The electronic control unit now knows the precise settings for every expected driving situation and can accurately calculate and adjust fuel, air and ignition timing according to the driver's demands and outside conditions.

Computerised engine management has to monitor and control in real time. This means that the microprocessor has to act immediately on any input. Generally, computers store instructions while the central processor is busy, which then implements them after a short delay. While this may not be noticed in an office application, for example, it could be critical in open-loop engine management, where the computer uses information from combustion in one cylinder to amend within a few milliseconds the calibration of the next cylinder.

Further features the new system include monitoring and self learning of exhaust gas mixture. This helps catalytic converters perform more efficiently and maximises life expectancy.

Alternative management systems

Zytek also supplies sports car engine management systems and its EMS3 system is now 32bit, using the Motorola 68332 processor. Different set-ups can be stored on disk to suit individual drivers or racetracks. There are facilities to switch from one to two injectors per cylinder at a pre-determined engine condition and to transmit engine data over a telemetry link. Carbon fibre also has a place, in a lightweight housing for the ECU. The company also supplies Aston Martin and Rolls Royce.

MBE Systems designs, develops and manufactures engine management systems. Its 941 ECU is a fuel injection and distributorless engine management system for use at serious club and works level motor sport on engines with up to 12 cylinders. Here reliability can win championships as well as speed because regularly collected points can beat irregular first places. For this reason MBE Systems EMS allows users to retain the manufacturer's original parts for reliability and this also helps with availability of spares.

The main ignition and fuel injection maps on the 941 system are arranged with map points every 250rev/min with up to 16 programmable throttle points. There are also maps with up to 16 programmable pressure points. Cold start and warm up maps are available and alterable by the user. The system can compensate for inlet air temperature variations.

Fuel is cut off during overrun and a gearshift light acts as a driver prompt. Weight of the unit has been trimmed to 750g. MBE has developed all its software and hardware inhouse and the management systems need a PC installed with the company's Easimap software, which runs under Windows, and a mapping console. MBE says that using a proven system such as theirs enables vehicles to be certified for road use first time.

16-bit control

A 16-bit microcontroller controls all ECU functions including communication with the company's 933 Traction Control System. This was developed in conjunction with Vauxhall Dealer Sport for the Astra F2 Rally Car. It measures all four wheelspeeds, lateral G forces and can also accept footbrake and handbrake positions. Twelve levels of traction control can be selected from ice through gravel to tarmac. A serial link connects it to the ECU which then controls engine power output. It differs from many traction control systems by finely adjusting engine characteristics rather than simply cutting ignition or fuel.

Cars manufactured by TVR use EMS from MBE Systems. The current range includes Cerbera, Griffith 500 and Chimaera. All can reach 60mile/h in a fraction over four seconds with engines between 4 and 5 litres developing some 340 bhp and providing a top speed about 160mile/h. TVR Engineering has just developed its own engine, the AJP 8. This is an all aluminium 4.2 litre V8 which develops 350bhp at 6500rev/min.

Rover Group has developed its own EMS but sells it with an engine. This combination powers Morgans, the Lotus Elite and Reliant Scimitar.

A smooth ride has low priority in racing cars but this is not the case for luxury cars. In the latest Jaguar models, the automatic transmission can control ignition through the engine management system. Ignition timing is retarded during gear shifts to give seamless gear changes and to increase transmission life by smoothing the load on the gearbox components.

Vehicle security is also enhanced because fuel, ignition and cranking are inhibited when the ignition key is removed and computers cannot be hot wired. A coded transponder in the ignition key has over 1000 billion combinations to ensure a robust system.

Exhaust emissions are continuously monitored by sensors in the exhaust system. These communicate with the management system and the mixture to the engine is automatically adjusted to compensate.

Another problem for the ECU to solve in a Jaguar is the extra load from air conditioning compressors. When these switch on during idling, for example, engine speed has to increase to compensate for the extra load.

Wiring it up

Interconnection between ECU, sensors and engine devices increasingly uses serial transmission. Although the principle is the same as the long established RS232 standard, signals have to be much more robust to withstand the harsh electrical environment in cars. Controller Area Network, CAN, is the accepted standard in the automotive industry.

With 8km of cable harness in some cars at the moment, replacement by twisted pair will mean a great weight saving. Power is wired separately to the command wiring but because it runs as a ring main to all devices there is further weight saving over conventional harnesses. The communications protocol is complex to eliminate the problem of collisions of data on the network and the system can distinguish between permanent hardware failure and occasional soft errors. Defective nodes are automatically switched off the bus, implementing a fail-safe procedure.

Electronic management systems are in turn breeding advanced mechanical devices. Siemens Automotive has just announced one of the fastest acting injection valves and a lambda probe which fits into an exhaust to monitor emissions. The injector valve has been developed because engine management systems demand faster response from the engine to make most use of the available increased efficiency.

Siemens has achieved a response time of 0.1ms compared with 0.6ms of electromechanically operated counterparts. A multilayer stack of piezoelectric material expands to provide mechanical motion from an applied voltage. The amount of movement is measured in micrometres.

And the future?

Once engine control systems are fully established, interest will turn to vehicle handling and the ECU will process inputs from additional sensors. The most advanced system yet proposed comes from Mercedes.

This company's ECU is linked to engine, automatic transmission, brakes, accelerator, steering and a yaw sensor through a CANbus. The accelerator is electronically linked to the engine management system, bringing the Mercedes close to drive-by-wire. A variable resistor is rotated as the accelerator is depressed to give a much finer control than with mechanical linkage.

Mercedes' system prevents skidding by applying brakes to individual wheels to maintain the car's balance. Control signals to the brakes derive from a computer which compares steering wheel position with the car's direction of travel. When under- or over-steer is detected, the appropriate brakes are momentarily applied and engine torque is reduced. This brings the car back on line.

Key to the system is a solid-state gyro which acts as a yaw detector. Housed under the rear seat of a car it gives an output signal proportional to the rate of rotation about a vertical axis. A database holds all the parameters for optimum handling stability.

Engine management could also work in conjunction with anti-collision radar to reduce engine speed when a vehicle is too close to one in front. It already interacts with intelligent cruise-control systems.

Whether it is through direct involvement or through general interest there is more and more in today's cars to attract the electronics enthusiast. ■

A new sensor for medicine

Developed with medical applications in mind, Chris Lavers' new opto-electronic technique could make diagnosis quicker, equipment cheaper and sensors smaller.

Currently, there is a need for simple and rapid techniques to detect antigens of medical and veterinary importance. The immune system reacts to attack by producing specific antibodies which will only react with one intruding antigen. They will not react with structurally related compounds, in the same way that only a single key amongst many will fit a given lock. This is the central process of antibody immunoassay.

In the past, immunoassay has involved the use of radionuclides or sensitive detection of fluorescently labelled molecules. However, these methods are now considered unfavourable. They are also slow. As a result, the use of non-labelled molecules has both practical and psychological advantages over traditional techniques.

An optoelectronic solution

Optoelectronics is taking a leading role in the field of optical sensors, where changes in optical intensity – due to variations in sample environment such as refractive index or absorption – offer the opportunity of on-line sensors for clinical and industrial applications.

Semiconductor technology has allowed the fabrication of monolithic integrated chips to be combined with total internal reflection within glass waveguides.

Total internal reflection sensors utilise the existence of the so-called evanescent wave. At

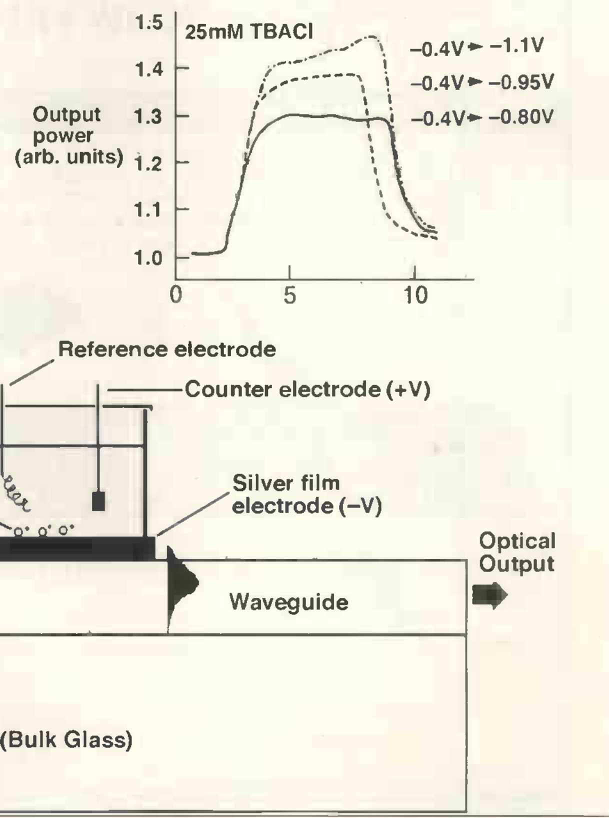


Fig. 1. Variations in refractive index near the waveguide surface modify output beam intensity. Applying voltage between the positive counter electrode and negative silver film electrode enables positive tetra-butyl ammonium – TBA^+ – ions to migrate towards the negative silver film. Ion migration results in changes of refractive index and hence output power. 25mM is 25 millimolars.

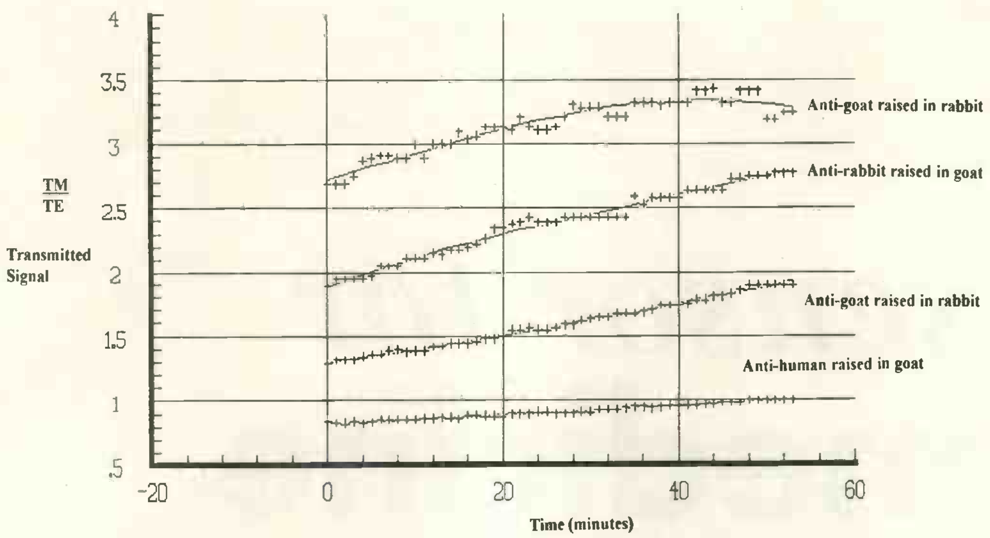


Fig. 2. Transmitted output power of the waveguide (TM/TE) is recorded as a function of time for the following sequence of antibodies bound to the waveguide – a) anti-human raised in goat, b) anti-goat raised in rabbit, c) anti-rabbit raised in goat and d) anti-goat raised in rabbit. As each subsequent layer binds to the preceding layer the transmitted output power is observed to rise in each case.

the boundary between two dielectric media, light incident from a denser medium may be reflected from a rarer medium if the incident light approaches at an angle greater than the critical angle. This angle is Θc, where $\sin\Theta c$ is n_2/n_1. Values n_1 and n_2 are refractive indices of the denser and rarer medium respectively.

In an evanescent wave, the electromagnetic field decays exponentially with distance from the reflecting boundary. It then probes above the glass surface to about a wavelength's depth into the surrounding medium. This medium may be, for example, a sample of body fluid.

Intensity of the surface wave may be amplified significantly via surface plasmon resonance – an electromagnetic wave travelling along a metal/dielectric surface.

Optical excitation of a surface plasmon resonance is achieved when radiation undergoes total internal reflection at the interface between the glass waveguide and a thin metal film deposited on top.

In the case of a planar waveguide, resonant coupling depends critically on the refractive index of the fluid adjacent to the metal. This technique can monitor small changes in index caused by either deposition of charged ions in an electric field, **Fig. 1** or by antibody-antigen binding near the surface of a waveguide, **Fig. 2**. The deposition technique was reported at OFS9 in Florence 1993 while the binding method is more recent.

Summary

Several advantages can be gained from using sensitive surface plasmon resonance optical sensors. These are small size, optical fibre compatibility, cheap production costs, and disposability.

Use of such multifunctional integrated optical sensors for rapid medical sensing – with monitoring of real-time antibodybinding and detection in minutes – is thus a realistic expectation before the end of this millennium. However, psychological aspects of rapid diagnosis of HIV, pregnancy and indeed any other immune response has not yet been considered fully. ■

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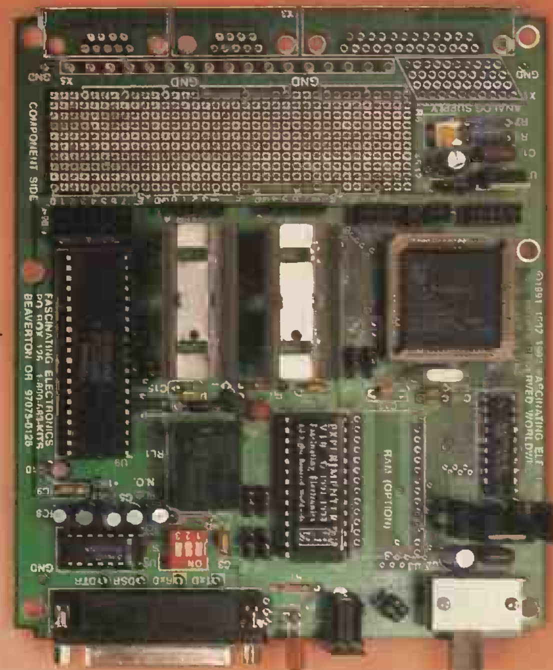

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ADC-11 **has eleven analogue input channels and takes up to 15000 samples a second. It plugs directly into the pc printer port. This 10bit converter takes inputs in the range 0 to 2.5V and has a digital alarm output. Data-logging software is supplied with the** ***ADC-11*** **as standard.**

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For current domestic tv sets, cathode-ray tubes still have no competitors. But crt technology will have to be replaced when large-screen tv comes on line. Peter Willis looks at a few of the options.

Technology for big TV

High-definition television needs large screens to display it – larger than is practical or feasible with cathode ray tubes. Flat panel displays using liquid-crystal displays have been in development for years, but have so far not been produced in sizes much above 20in, i.e. 50cm.

Now though, a variety of approaches appears capable of producing large flat panels – the commonly agreed target diagonal is 50in or 127cm – in the very near future. Several prototypes were on show at the Berlin IFA during August and September.

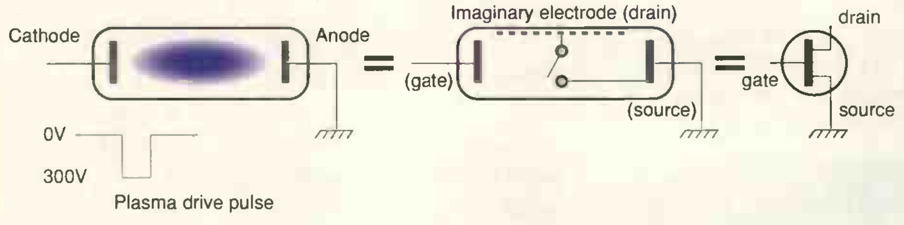
Some use plasma discharge, the result of passing high voltage through gas. Thomson showed a plasma screen deploying this phenomenon in the traditional way, as a light source, exciting further light from phosphors. Several Japanese companies are working on similar technologies.

Sony's Plasmatron however, uses plasma discharge as an electronic switch similar to a transistor, to control lcds in front of an independent source of light. The plasma switch takes the form of a channel across the screen, equivalent to a single scanning line, and inputs image signals for that line instantaneously into the liquid crystal.

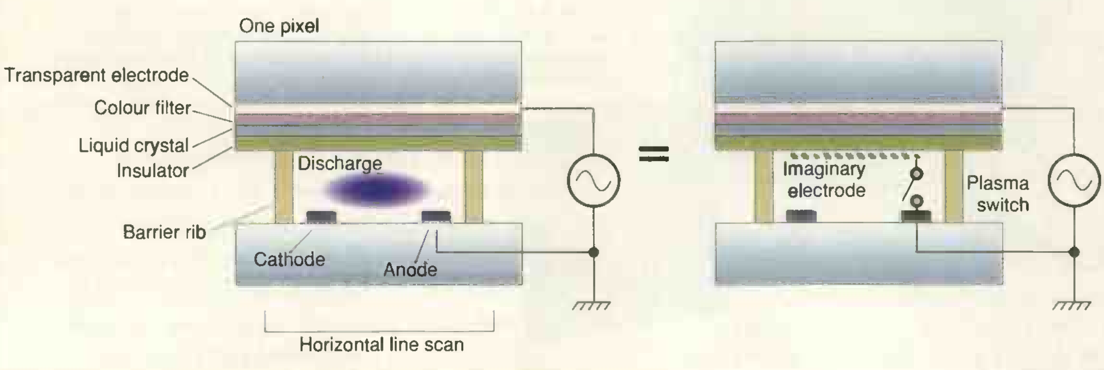
The system, called PALC, – plasma addressed liquid crystal – was developed by Sony in conjunction with Tektronix. According to Sony it has the advantage of relative simplicity and relatively low-cost production. Sony was showing three 30in prototypes in Berlin, and aims to have the first Plasmatron sets in production next year. However, the eventual size of these displays is not yet defined and the name not confirmed.

Each pixel is a rotating mirror

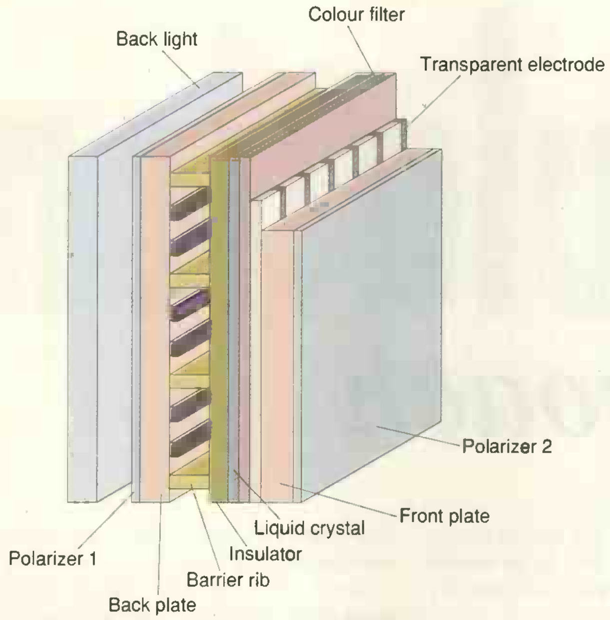
An entirely different technique developed by Texas Instruments has been adopted by Nokia who demonstrated it in Berlin. It uses an array of tiny movable mirrors, one for each pixel, which when



In the plasma-addressed liquid crystal, each plasma switch is the equivalent of a transistor. When the plasma switch is on, the whole channel is conductive. When the switch is off, the channel forms a good insulator.



When a line of plasma switch is on, video signal is written to the liquid crystal layer through the transparent column electrode and anode electrode.



Greatly enlarged structure of a plasma-addressed liquid crystal switch element.

tilted, reflect light onto a back-projection screen. The mirrors, numbering around 500,000 and making up a panel measuring only 1.5 by 1cm, are part of a semiconductor device. They are created by an etching process, and although seemingly microscopic – 900 fit under a grain of salt – they are large in semiconductor terms with a 16μm cell size. Each mirror is square, and mounted on a pair of torsion hinges at diagonally opposite corners. Output of the chip, instead of going to pins, is taken to electrodes beneath the array, which electrostatically attract individual micromirrors.

For each frame of the television picture, the image is built up by tilting the required mirrors to reflect high-intensity light source onto the screen. Brightness of each pixel is controlled by an 8bit command which can instruct the mirror to be on for the whole or part – down to 1/256th – of the operation. Three such operations make up each frame, one each for red, green and blue, with the results passed through a synchronised colour filter wheel.

Despite the small size of the picture source, there is no observable line structure on the screen. Developed by Texas Instruments under the titles digital light processing and digital micromirror device, the system can provide a picture of over 50in or 127cm diagonal in a cabinet only 15in or 38cm deep and weighing a relatively modest 35kg. The normal array of 440,000 pixels equates to standard definition television, and produces such a startling improvement in image quality as to call into question the need for high definition television. If it were required, a two million micromirror chip could be supplied.

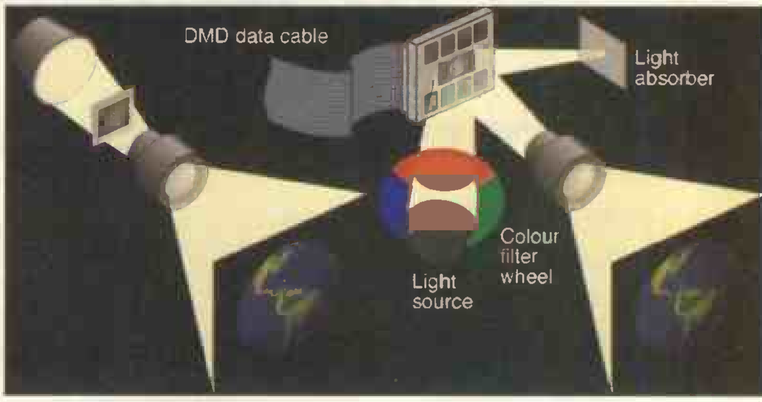
Digital light processing also provides a number of other advantages, including uniform picture geometry, high resolution across the screen area, high contrast over a wide dynamic brightness range and an almost complete freedom from flicker. It also offers the possibility of end-to-end digital video with the ability to address each pixel individually and freedom from noise.

Is mechanical breakdown a problem?s

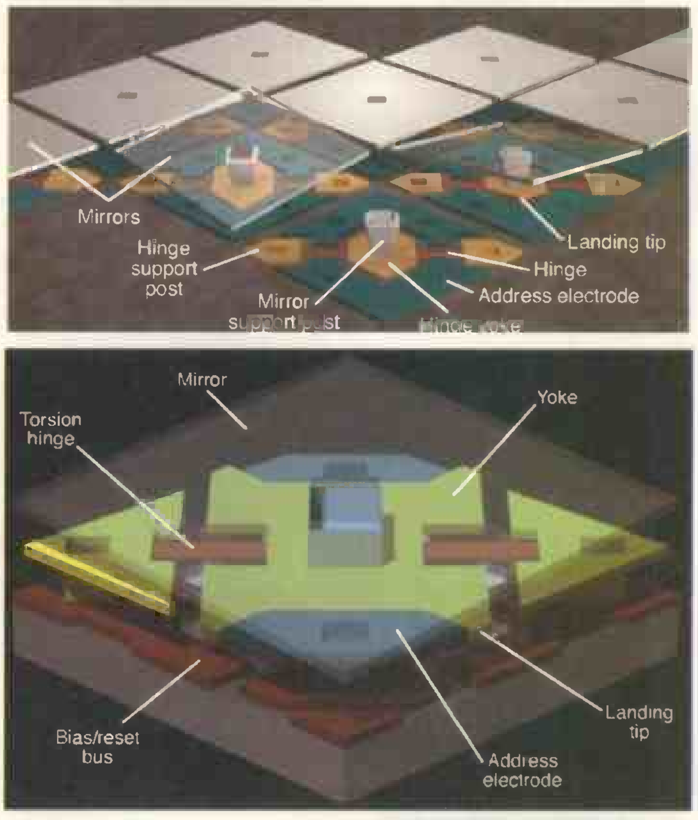
The reliance on what are in effect a great number of mechanical components might give some concern. One stuck mirror could create a dark, or more obtrusively, a bright spot on the screen. But according to Texas Instruments digital light processing manager, Adam Kunzman, the torsion hinge on which the micromirror's operation depends has been extensively tested and has "never failed". Nor is it expensive. Applicability of the system to a wide range of uses, including cinema projection, means that the volume and yield of production can be quickly increased, and costs brought down. Eventually consumer products costing under £1000 are thought to be feasible.

Nokia is the first consumer manufacturer to publicly adopt the system, and plans to launch a big-screen set in early 1997. Other partners of Texas Instruments so far announced include cinema, home cinema and business equipment firms but, interestingly Sony is also included. The system is particularly kind to large-scale vdu screens and ohp's. Talks are understood to be in progress with a number of other companies.

In the contest to produce the definitive flat-panel large-screen display – or even a reliable, affordable one – digital light processing may prove to have one clinching advantage over plasma-based systems. It does not rely on high-voltage pulses which could contravene regulations on electromagnetic interference. ■



Conventional projector, left, and a projection system incorporating the digital micromirror device, right. In the DMD system, light from the source passes through a colour wheel and is reflected to both the projection lens and to the light absorber on the right. Picture courtesy Texas Instruments.



Details of the digital micromirror device. 500,000 mirrors – each hinged and capable of rotating independently under digital control – make up a display measuring just 1.5 by 1cm. Pictures courtesy TI.

Demodulation a new approach

Archie Pettigrew's new demodulator concept uses amplitude-locked loop techniques to produce significant improvements in the quality of fm and am reception.

The amplitude-locked loop was developed to overcome a number of fundamental difficulties which have existed since the inception of both amplitude and frequency modulation – am and fm.

With the radio spectrum becoming more crowded each year, and carrier frequencies moving inexorably higher, two basic problems with am and fm transmission become more obvious. Amplitude modulation becomes highly distorted when the carrier fades – or in certain cases, vanishes altogether. Frequency modulation becomes highly distorted and unintelligible when another fm signal arrives at the antenna at the same time as the wanted signal which is equal in amplitude and of a similar frequency.

Both these breakdown processes are caused by interference in the form of multi-path[1] Doppler or quasi-synchronous[2] reception. All these forms become worse as frequency of the carrier is increased i.e. as wavelength is shortened and/or as transmission becomes mobile[3]. By using an amplitude-locked loop and associated circuitry, many of these interruptions can be avoided, and more reliable communications achieved.

This article describes in detail the operation of two demodulators, one for am and the other for fm, using the amplitude-locked loop.

Amplitude-locked loop

The amplitude-locked loop, ALL, is the dual of the phase-locked loop, PLL. It works in the magnitude domain rather than the phase or frequency domain. It consists of a linear multiplier contained inside a high gain, high bandwidth servo loop.[4,5,6]

A phased-lock loop is similar in that it consists of a voltage controlled oscillator contained in a high gain, high bandwidth servo loop. **Figure 1** shows a diagram of the amplitude-locked loop[7].

Carrier from the intermediate frequency stage of the radio enters the ALL at the first port of the linear multiplier. The second port of the multiplier is set to a nominal value of unity. The modulated carrier passes through the multiplier to the modulus detector which accurately detects the modulus of the carrier down to white noise levels.

A dc reference voltage compares and subtracts the incoming modulus voltage and a difference or error voltage is generated, $e(t)$. This voltage is integrated and reversed in sign. Output of the integrator is added to the restoration voltage which sets up the operating conditions – or bias conditions – of the loop.

When no carrier amplitude is present, the loop is out of lock. The integrator drifts to its maximum voltage and the multiplier is at maximum gain awaiting an input. When the carrier appears, an initial transient occurs as the loop pulls into lock. Servo feedback causes the carrier amplitude at the output of the multiplier to be fixed or locked to an amplitude defined by the loop reference voltage.

Let a simple amplitude modulated carrier $v_1(t)$ be described as,

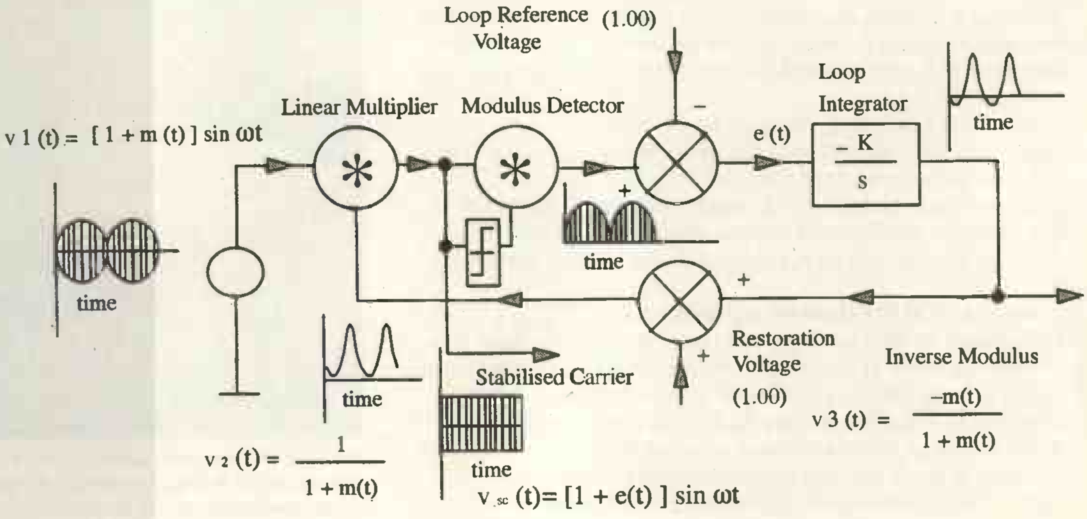


Fig. 1. Amplitude-locked loop consists of a linear multiplier, modulus detector and a high gain integrator. When the loop is closed, envelope variations of the carrier are reduced to insignificant proportions due to servo action and an error signal called the inverse modulus is produced.

$$v_1(t)=[1+m(t)]sin\omega t$$

where $m(t)$ represents the modulating function of time and $sin\omega t$ is the normalised carrier amplitude of ω radians per second.

After some mathematical analyses, a number of amplitude-locked loop identities become evident. Assuming that open-loop gain is sufficiently high that servo theory is valid, ie the value of K in the integrator is greater than 100 at the maximum frequency of interest, the stabilised carrier, $vsc(t)$, becomes,

$$v_{sc}(t)=[1+e(t)]sin\omega t$$

where $e(t)$ is the loop error voltage which becomes insignificant due to the high open loop gain. That is, $v_{sc}(t)=[10]sin\omega t$

Voltage $v_{sc}(t)$ represents a stabilised carrier with no envelope variations. Voltage at the second input to the multiplier must therefore be the reciprocal of the input modulation. As a result, $v_2(t)$ is $1/[1+m(t)]$ and $v_3(t)$ is $-m(t)/[1+m(t)]$ by subtracting unity.

Three signals have been obtained – the unmodulated carrier, the inverse of the modulus and the inverse modulus with the dc term removed. Unfortunately, there is no requirement to recover the unmodulated carrier in amplitude modulation. The demodulated signal is the reciprocal of the modulation which is a highly distorted version of the original signal. The signal at the integrator output is simply the reciprocal of the modulation but with an average value of zero. At first sight, nothing seems to have been achieved by this circuit so why investigate further?

Much the same arguments were used for the PLL when it was first suggested. For example, the PLL could easily have been replaced by a piece of wire and at a much lower cost etc.

Perhaps for the above reasons the concept of the ALL has never been investigated, even in the valve or tube era of electronics. If the ALL is not directly suitable for demodulation of am, can it be used to replace the limiter-filter in the demodulation process? Indeed it can as will be explained.

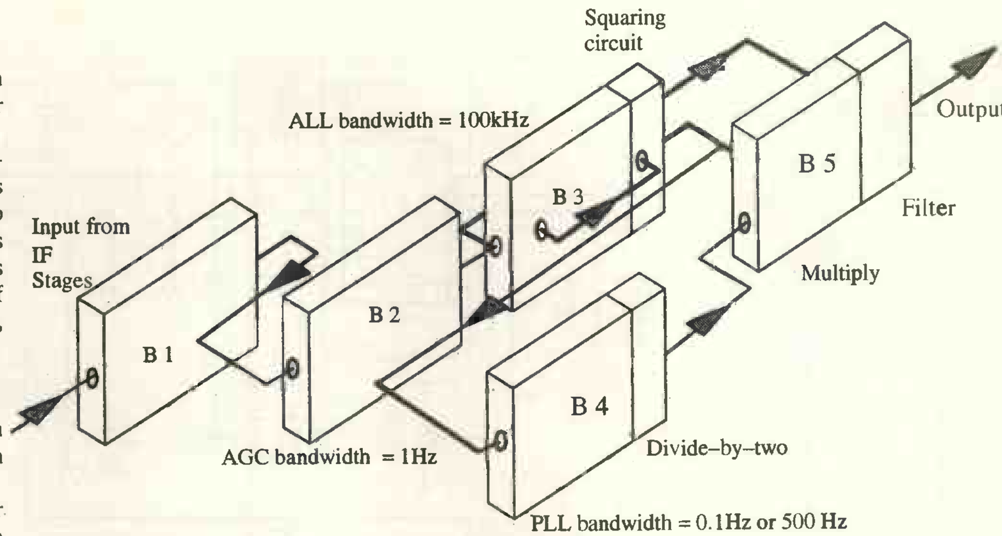


Fig. 2. ALL-PLL amplitude demodulator. The amplitude-locked loop alternates between in-lock and out-of-lock for strong and weak carrier signals. The PLL captures the carrier phase quickly but releases it slowly. A highly stable carrier is generated.

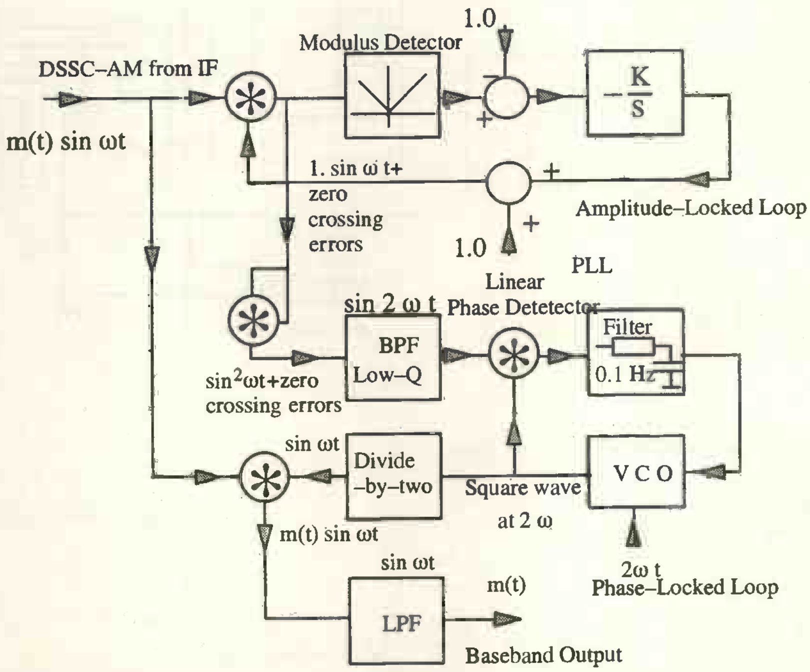


Fig. 3. Amplitude demodulator for double side-band suppressed carrier. The amplitude-locked loop generates a constant envelope from the input carrier. As the carrier approaches zero the amplitude-locked loop loses lock and the gain changes by 80dB. Phase-locked loop bandwidth is reduced by the same amount and synchronous demodulation is now feasible even in conditions of high noise.

Application to fm demodulation

When two fm carriers of equal amplitude are added, their envelope increases to twice the individual size and reduces to zero at the instantaneous difference frequency. This envelope variation will be eliminated by the servo action of the ALL. This is similar to the action of a hard limiter and a filter and fulfils the first requirement in fm demodulation – that am variations must be removed before demodulation.

A second signal is also available which is the inverse of the modulus of the two carriers. Could this second error signal be used constructively to improve fm demodulation?

Operating limits

Before continuing, it would be sensible to define the limits of operation of the first ALL unit. Starting with an intermediate frequency of 455kHz, amplitude and phase information is updated at twice the carrier frequency or 910kHz.

In a closed-loop feedback system, instability starts to occur at about one tenth of this frequency or 91kHz. So the ALL unity gain bandwidth was set to 91kHz giving an open loop gain at say 1kHz of 91 or 39dB. This was improved later by using a double integrator.

The dynamic range of the ALL was determined by the offsets and the characteristics of the linear multiplier, the *Exar 2208*. This was found to be to +20dB (10) to –6dB (0.5) or a linear lock range of 26 dB.

In practice, the ALL will track 26dB of amplitude variation up to a frequency of about 20kHz without significant error. This was found to be more than adequate for all narrowband fm speech channels. The lock-in transient is very short since the ALL is more linear than a PLL. Typically it was measured at about 4μs, i.e. the time required to reach 95% of the steady state value of stabilised amplitude.

Using a simulation package called *MATRIXx*, optimum circuit operation was established before any hardware was constructed. A much improved loop was designed using a double integrator with suitable phase advance for stability. This is described further in reference 8.

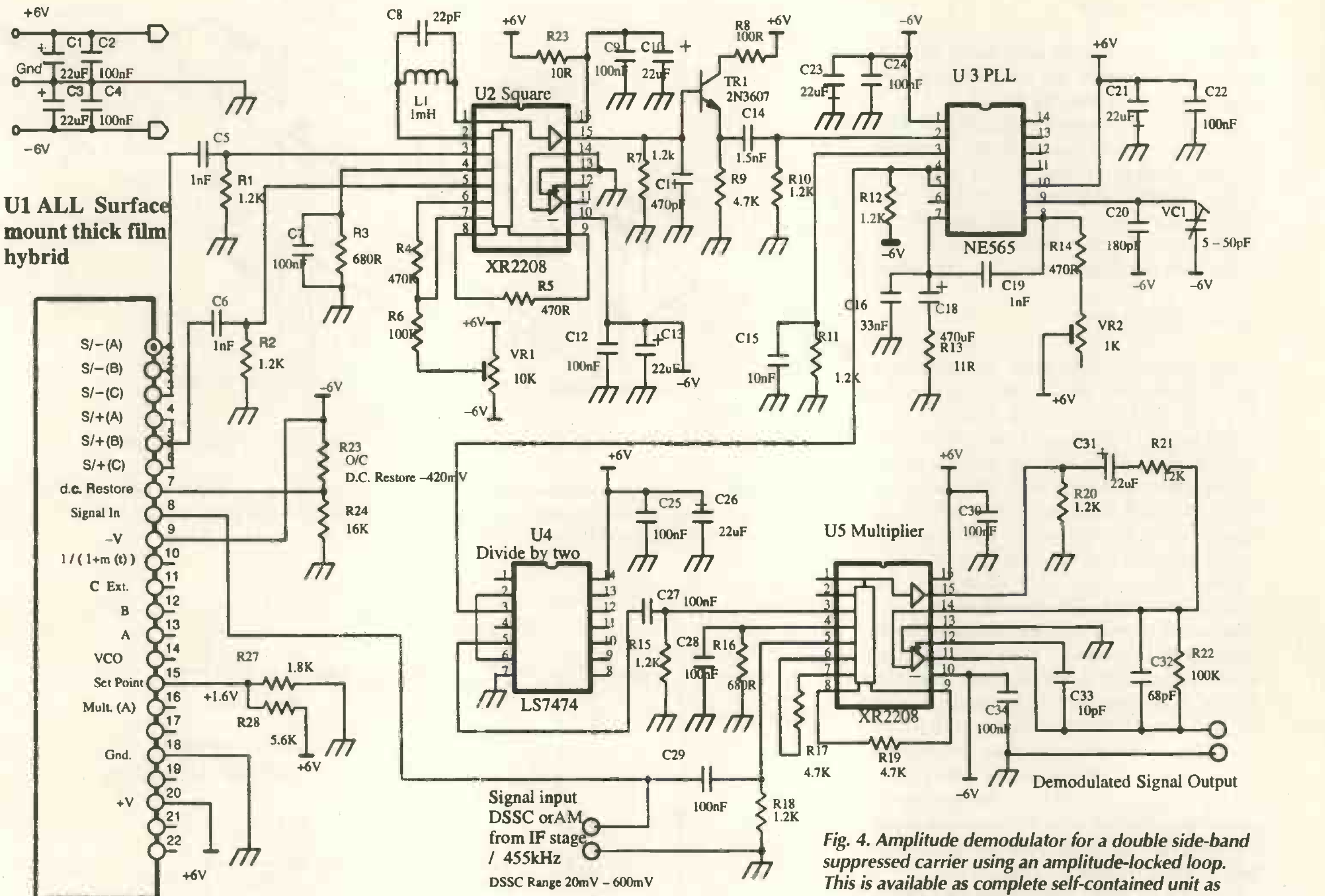


Fig. 4. Amplitude demodulator for a double side-band suppressed carrier using an amplitude-locked loop. This is available as complete self-contained unit as shown in U1. ICs U2-5 comprise a standard synchronous demodulator.

Applying the amplitude-locked loop

The first application of the ALL is to improve the amplitude modulated double side-band suppressed carrier. This represents the ultimate in carrier – or Rayleigh – fading since the carrier vanishes at every silence of the speech waveform, by definition.

The core of this problem is the recovery of a stable carrier. There is no carrier present during the silence between speech. The worst case occurs at the lowest modulating frequency and lowest amplitude of the signal.

Doppler effects may cause the carrier to be shifted by say 100Hz so that high-Q filters are not permitted due to their rapid phase changes at resonance. Should the system lose lock, then reliable re-lock must occur within say one cycle of the lowest operating frequency, say 300Hz, or in about 3ms.

The carrier recovery circuit must also be able to track frequency variations up to 100Hz to an absolute phase accuracy of less than say 45° error between the modulated carrier and reference carrier. Assuming that a PLL is available to regenerate the carrier at 455kHz, then two conflicting conditions need to be met simultaneously.

Since amplitude of the DSSC signal is continuously varying, the envelope must first be

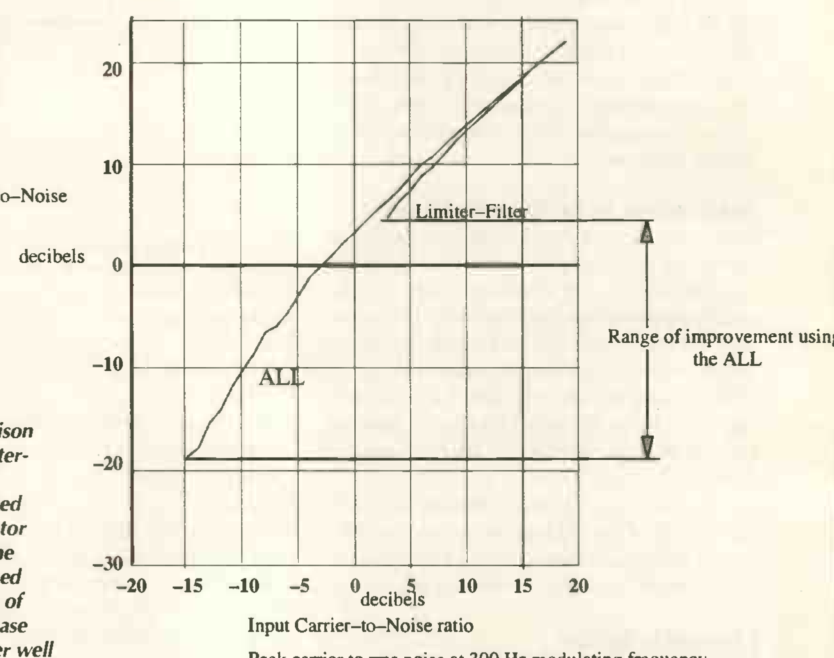


Fig. 5. Comparison between a limiter-filter and an amplitude-locked loop demodulator showing that the amplitude-locked loop is capable of extracting a phase coherent carrier well down into noise levels.

made constant. The PLL must have a wide capture and track range for fast lock-in and frequency tracking, yet it must have an extremely narrow noise bandwidth for stability during every speech silence.

The solution to these seemingly conflicting requirements is shown in **Figs** 2 and 3 in block diagram form and in circuit form in **Fig.** 4. This could be could be done by limiting and filtering which would be successful at high instantaneous amplitudes. A major problem occurs at low amplitude and low frequencies with noise. Noise captures the limiter, the voltage controlled oscillator becomes unstable and the phased-locked loop loses lock. System failure ensues. If the limiter is replaced by an ALL, a different process takes place.

At high instantaneous amplitudes, the large negative feedback of the loop flattens the amplitude variations giving a constant envelope at the output of the linear multiplier. At low instantaneous amplitudes, the ALL drops out of lock since its track range has been exceeded.

Gain of the 'loop' drops to the gain of the multiplier alone and not the combined gains of the multiplier, modulus detector and the integrator. At 300Hz, this represents a change from 80dB to 20dB. The noise level is not amplified, and in effect, the system closes itself down.

White noise is not permitted to overtake the signal, as would happen in a limiter. This is the advantage of the in-out action of servo feedback.

Carrier generation

A pure squaring device follows the analogue-locked loop to generate a coherent carrier at 2ω. When the ALL is out of lock, the PLL is being driven by a zero level carrier. Since the bandwidth of any linear PLL is a direct function of the input amplitude, its closed loop bandwidth drops to zero.

The voltage controlled oscillator free wheels on the long open-loop time constant of the PLL since there is no significant noise energy to cause perturbations. After a divide-by-two circuit, normal demodulation takes place.

Thus the PLL has effectively two bandwidths. The first is with signal present and the ALL in lock. With values suggested in Fig. 3 this bandwidth measures about 500Hz. When no carrier is present, i.e. with the ALL out of lock, there is no signal present at the input to the PLL.

The effective open loop gain of the PLL is reduced to zero assuming a linear phase detector.

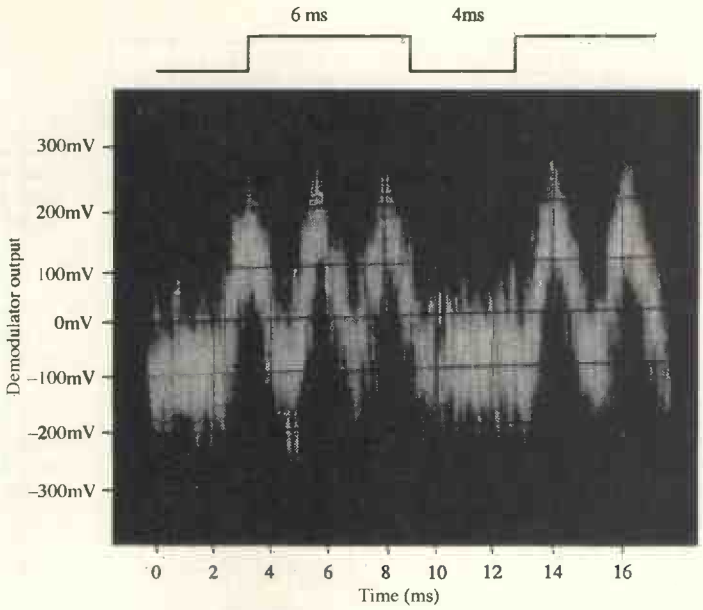


Fig. 6. Oscilloscope measurement of demodulated output showing a 300Hz signal gated on and off at 6 and 4ms intervals at a carrier-to-noise ratio of 0dB. Note stability of the noise during the period of zero level carrier. This is due to the amplitude-locked loop and the phase-locked loop both shutting down and awaiting the resumption of the signal.

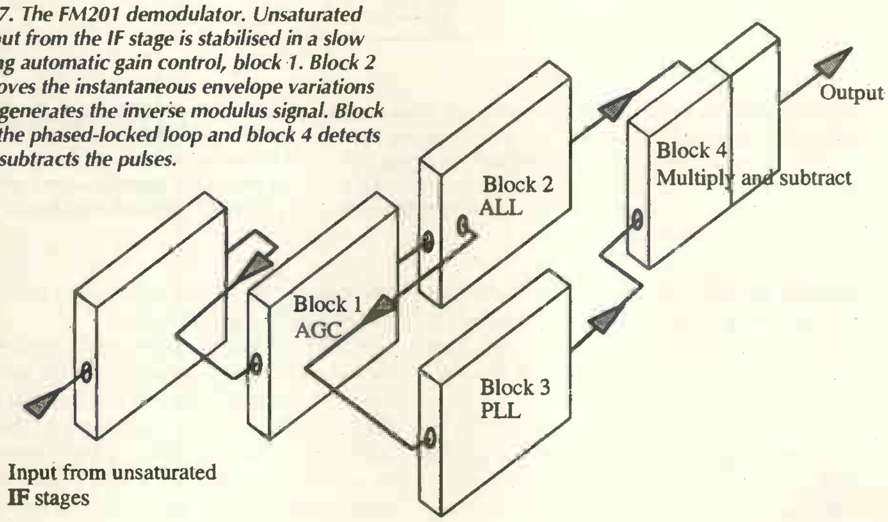


Fig. 7. The FM201 demodulator. Unsaturated output from the IF stage is stabilised in a slow acting automatic gain control, block 1. Block 2 removes the instantaneous envelope variations and generates the inverse modulus signal. Block 3 is the phased-locked loop and block 4 detects and subtracts the pulses.

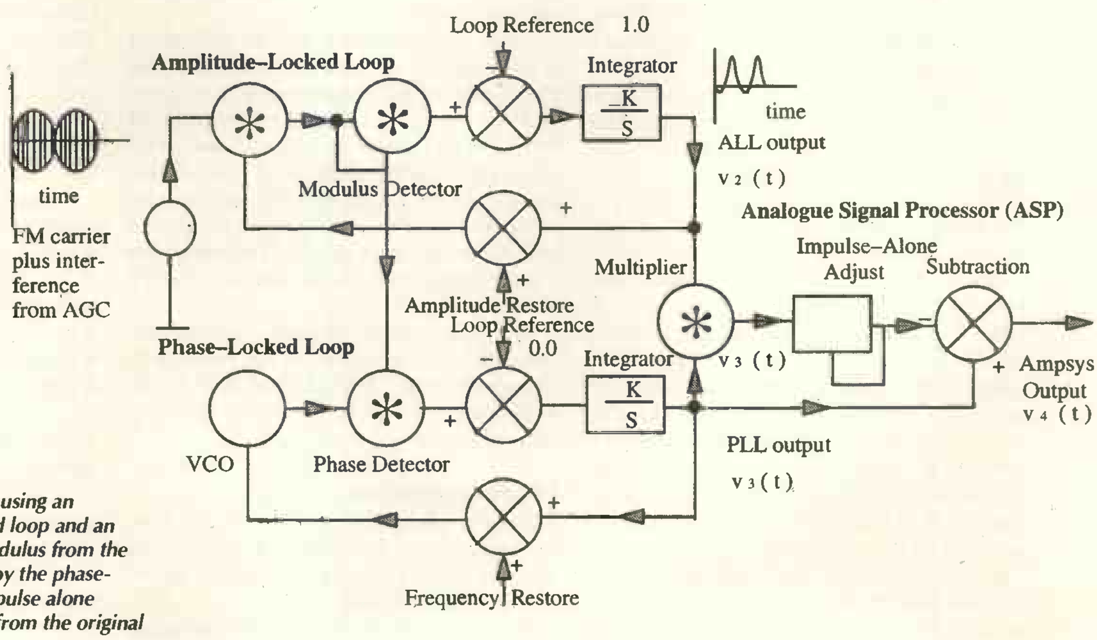


Fig. 8. Complete FM201 demodulator using an amplitude-locked loop, a phase-locked loop and an analogue signal processor. Inverse modulus from the amplitude-locked signal is multiplied by the phase-locked loop output to produce the impulse alone signal which is scaled and subtracted from the original phase-locked loop signal.

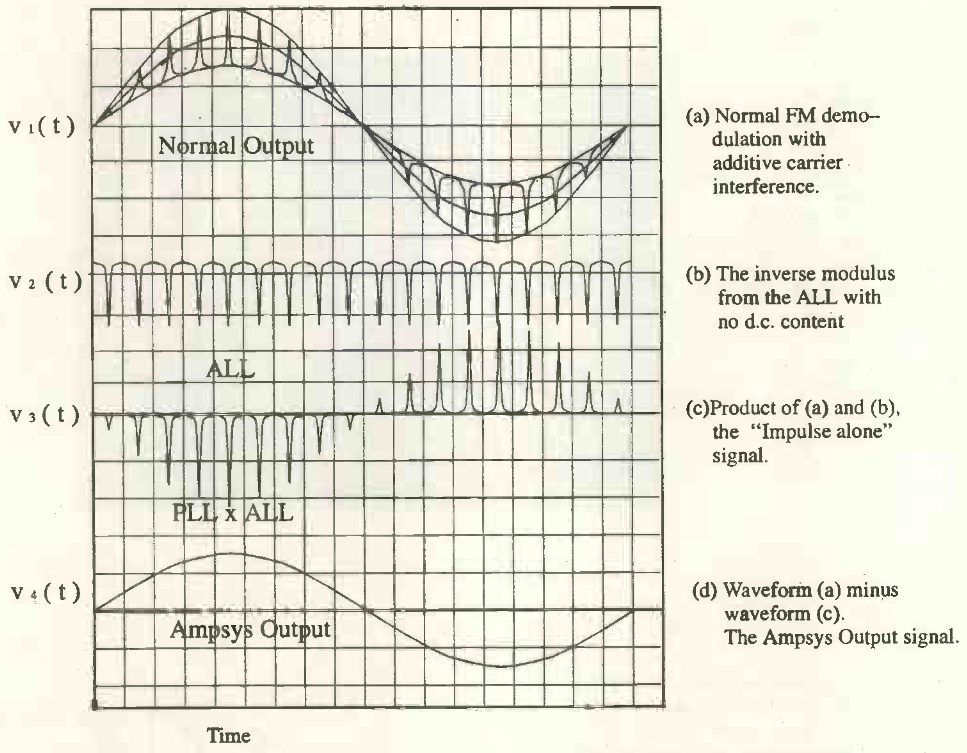


Fig. 9. Waveforms in the FM201 demodulator. The inverse modulus $v_2(t)$ is multiplied by $v_1(t)$ and subtracted from the phase-locked loop output to give $v_4(t)$. Fig. 9a).Normal FM demodulation with additive carrier interference – phase-locked loop output. Fig. 9b) inverse modulus from the amplitude-locked loop with zero average ie no dc content. Fig. 9c) shows the 'impulse alone' signal – product of the signals 9a) and 9b). Fig. 9d) final demodulated output where all harsh impulses or spikes have been removed.

Stability of the voltage controlled oscillator is then determined solely by the time constant of the filter following the phase detector. This can be made large ie of the order of one second.

Carrier stability is thus maintained due to this very long time constant. By use of the ALL, phase-capture transients are very short when signal is present and phase loss transients are long when the signal is absent. By this technique, the coherent carrier can be recovered reliably even during periods of poor carrier-to-noise ratio. The circuit diagram of this system is shown in Fig. 4.

The ALL is contained in the hybrid block U_1. The circuits which follow the ALL represent the normal synchronous AM demodulation technique, namely, a pure squaring device (U_2) followed by a narrow track range PLL (U_3), a divide-by-two, (U_4) and finally a synchronous multiplier (U_5).

Results obtained for this demodulator are presented in Fig 5. **Figure 5** shows the comparison between a demodulator using a limiter and filter in place of the ALL. Whereas the limiter-filter ceased to operate effectively at about 3dB carrier-to-noise ratio, the ALL circuit still maintained synchronism until well into noise. Cycle slipping occurs in both demodulators at about the same relative position but does not result in complete loss of intelligibility.

It is interesting to note that there is no threshold effect present as would be the case in fm or angle demodulation. The output signal-to-noise ratio tracks the input carrier-to-noise ratio in a linear manner.

Results obtained from the above demodulator exceeded the performance of the normal synchronous demodulator in that carrier recovery could be achieved down to and below unity carrier-to-noise ratios. **Figure 6** shows an oscilloscope trace of a 300Hz sine wave signal which is being gated on and off at 6 and 4ms intervals. Carrier-to-noise ratio with signal present was 0dB. Noise and carrier amplitudes were equal.

Due to carrier stability, system white noise is demodulated in a coherent manner. The PLL has an effective phase capture bandwidth of 500Hz and a phase release bandwidth of 0.1Hz. This phase capture-release phenomenon is a direct consequence of utilising the two in-lock and out-of-lock characteristics of the ALL and PLL simultaneously to make a near perfect am demodulator. This represents a major improvement in the state of the art on am demodulation.

This demodulator operates reliably and completely independently of the presence or absence of carrier. It is therefore ideal for the reception of am during multipath or quasi-sync. conditions.

FM demodulation

Frequency modulation is transmitted at constant amplitude. Any amplitude variation at the point of reception must be due to interference or noise acquired en route.

According to perceived wisdom, amplitude variations must be removed by hard limiting and filtering of the carrier on reception. If not, two forms of degradation will occur at the demodulator output. The first is due to envelope variation and the second to phase variation.

In reality, the fm carrier is degraded not only by naturally occurring phase noise but also by amplitude noise. This is converted to phase noise in the limiting process. These two processes combine as the input carrier-to-noise ratio approaches a low value of typically 12dB.

The catastrophic fm threshold effect begins and rapid deterioration of the output signal-to-noise ratio then follows.

This same effect causes fm reception to be rendered unintelligible if two fm transmissions arrive at the antenna at equal or near equal strength to each other – assuming co-channel frequencies.

The corrupting carrier may be another transmission – co-channel – or a delayed version of the wanted carrier – multipath. It could even be a version of the same broadcast from an equidistant transmitter – simulcast or quasi-sync reception. Harsh acoustic spikes are demodulated which are inband and cannot be filtered.

Capture effect

In the past, much has been made of the 'capture effect' in fm. Generally, this means that if one carrier is say 10% stronger than the other, say 1dB, then capture takes place and the weaker station is completely suppressed. This was the argument put forward by Edwin Armstrong the inventor of fm. It is true – but it is not the whole story.

The unwanted carrier is suppressed but not into silence, which would be ideal. On the contrary, the co-channel interference is demodulated into strident noise, or large impulses which are intolerable to the ear and destructive of all intelligible communication. So destructive is this interference that all fm transmissions start to break down in the region where either carrier is within 6dB of the other. This is sometimes called the 'distortion zone' – when its existence is admitted. The 'capture effect' is not an advantage but is in fact a major disadvantage of fm in a crowded radio spectrum.

Frequency modulation works well when;

- carrier strength is high,
- there is only one single carrier,
- there is no co-channel interference,
- there is only one direct signal path,
- modulation depth is virtually unlimited
- transmission power is virtually unlimited.

These conditions prevailed some fifty years ago, but unfortunately not in today's overcrowded spectrum.

Ideal requirements of the modern demodulator would be a circuit technique which would make fm demodulation linear at low carrier-to-interference ratios but still have the co-channel rejection properties at high carrier-to-interference ratios. Quasi-sync and multipath reception would then be improved by the addition of the intelligence in the carriers and co-channel reception would be equivalent to crossed lines in telephones.

The Ampsys FM201 demodulator

Using an amplitude-locked loop for the first time, an fm demodulator has been designed and tested which demonstrates the above requirements. It is designated the Ampsys *FM201*. Its block diagram is illustrated in **Fig. 7** and a system diagram in **Fig. 8**.

In the *FM 201* demodulator there are four separate processes or stages. The first process, after the normal intermediate frequency filtering, is to stabilise the wanted carrier and the interfering carrier to a fixed average value using a slow automatic gain control circuit. This process is necessary in order to present the ALL with a fixed average signal level.

The fm was originally transmitted as a fixed amplitude and the automatic gain control restores this long term average. Saturating limiting is always avoided. The automatic gain control block has a bandwidth of 10Hz.

In the second stage, block 2, the ALL removes all short term variations leaving the carrier similar to the output of a hard limiter and filter. This stabilised carrier is then applied to the input of the PLL, block 3, which is regarded as normal demodulation.

A second output from the ALL, the modulus reciprocal less the dc term, is applied to a multiplier in the analogue signal processing stage, block 4, Fig. 7. Output of the PLL is applied to the second port of this multiplier. A product is formed at the output of this circuit – the 'impulse alone' signal. 'Impulse' refers to spikes superimposed onto the baseband signal by the demodulation process.

This new baseband signal is scaled in size and subtracted from the original PLL output. Care must be taken to ensure that any phase delays through the ALL and the PLL are equal otherwise subtraction will not be possible.

A demodulated signal is created which is free of harsh spikes and is now perfectly intelligible even when the carrier and interference are identical in magnitude. A simplified version of the relevant waveforms is shown in **Fig. 9a-d**. Voltages $v_1(t)$ to $v_4(t)$ correspond to those marked on Fig. 8.

Worst-case fm reception

Figure 10 is an oscillograph of the demodulator output when the interfering carrier is located at the centre of the the intermediate frequency passband and is of equal amplitude to the wanted carrier. This represents one of the worst case conditions in fm reception. It would result in the carrier vanishing and doubling alternately at the instantaneous difference frequency. It is equivalent to a fade of infinite depth.

The normal output signal-to-noise figure is much less than zero and is unmeasurable by normal instrumentation. Acoustically, all intelligence is lost and the channel would be muted. With the *FM201* demodulator, signal-to-noise ratio rises to about 14 dB unweighted. This is acceptable in a communication channel and represents 100% intelligibility.

The link has been preserved, so avoiding the call being dropped. In normal demodulation when Gaussian white noise is added or when both carriers become weaker, distortion and noise effects become even more severe and generally intelligibility is lost just after the fm threshold point. This means that there can be a 'distortion zone' or failure gap as wide as 12dB. In simple terms this means that if one carrier is more than one quarter the size of the other at the antenna, failure ensues rapidly.

With the *FM201* demodulator however, the spikes are removed, as are the 'Rician' spikes due to Gaussian white noise. The net result is a much improved communication channel with almost 100% intelligibility well below threshold. Further testing has given the following observations.

- When the interfering frequency is offset from the centre of the passband, similar subtraction can be achieved by inserting an offset voltage at the input to the final multiplier. The interfering frequency however must be fixed.

- When quasi-synchronous reception occurs, baseband signals combine and an improvement in the signal-to-noise ratio of approximately 26dB is measured at the equal amplitude reception point. Multipath distortion causes a small reduction in signal-to-noise ratio.

- When two modulated carriers are present, the result is similar to that of a crossed line in a telephone. Although this is not ideal, it is preferable to complete loss of intelligence.

- With a very weak carrier, harsh spikes are removed and noise subjectively more acceptable. White noise can never be removed since there is never enough unique information.

- When the carrier-to-interference ratio is high, the 'impulse-alone' product diminishes rapidly since there are no envelope variations. The weaker transmission is suppressed as in normal demodulation and all beneficial characteristics of fm are retained, for example, quieting and co-channel suppression.

To summarise

A new circuit concept has been proposed called the amplitude-locked loop which can be used in conjunction with a phase-locked loop to improve the quality of fm demodulation.

By using the fundamental property that fm is transmitted at fixed amplitude, and that a unique relationship exists between the reciprocal of the modulus and the fm phase perturbations, a new signal has been derived called the 'impulse-alone' signal.

By a simple subtraction process this new signal can be used to eliminate spikes generated in fm demodulation. A fundamentally improved method of fm demodulation has been proposed which meets the criteria set out above for fm demodulation in today's overcrowded radio spectrum.

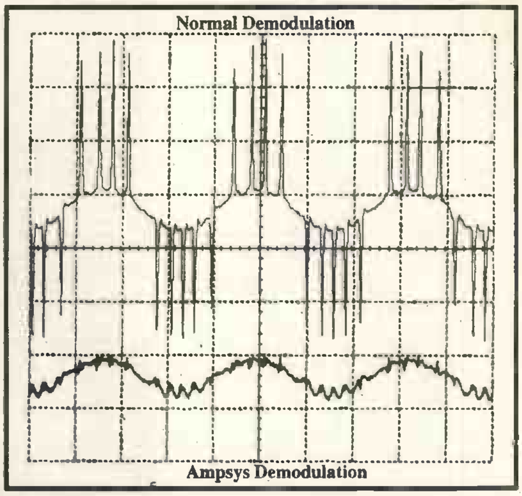


Fig. 10. These spikes could not be removed by any form of baseband filtering. At the interference-to-carrier ratio of unity the spike size is reduced by a factor of 20 fold, 26dB. Normally all intelligibility is lost. With Ampsys, demodulation is 100% intelligibility.

Two demodulators using the ALL have been built and tested and are available for evaluation purposes, from Ampsys, one for am demodulation and the other for fm.

Acknowledgments

I would like to thank Dr T. J. Moir of the Department of Electrical and Electronic Engineering at the University of Paisley for his help in the preparation of this paper and to the Governors of the University of Paisley for their support in the development.

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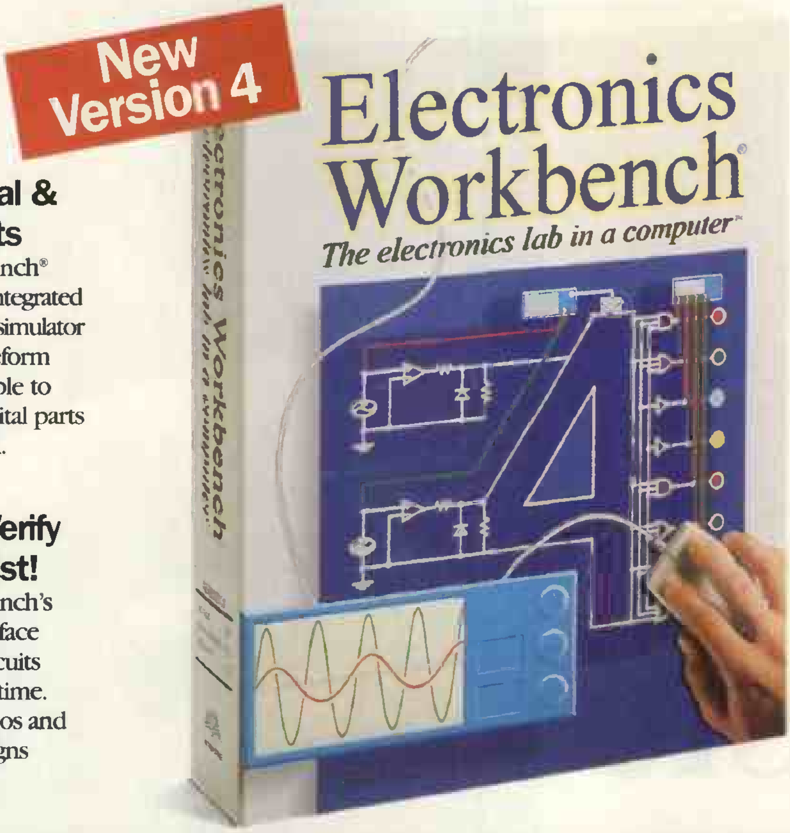

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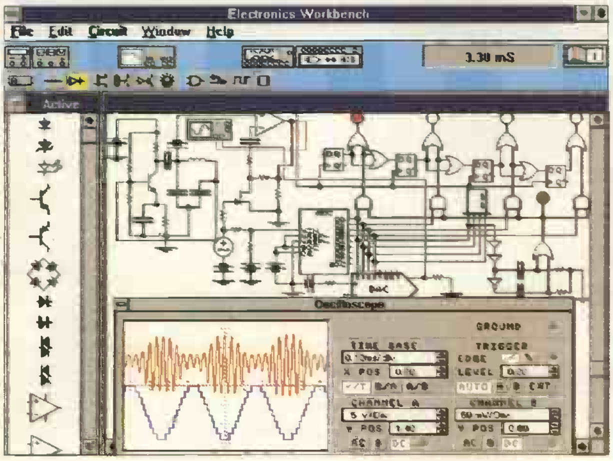

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Classic valve power

Williamson's valve power amplifier was published in *Wireless Word* in 1947, and set a standard of performance that was years ahead of its time.

The input stage is the standard common cathode triode with 20dB of global negative feedback applied from the loudspeaker output to the cathode. The phase splitter is a concertina circuit, direct coupled from the input stage. It feeds a differential pair using both halves of a *6SN7*, Fig. 1.

Operating in Class-A

The output stage is a push-pull pair of *KT66* beam tetrodes operated as triodes. These provide 15W output in Class AB1, operating mostly in Class A. In Fig. 1, the left hand potentiometer adjusts the dc balance of the output valves in order to minimise distortion due to the transformer core. The second potentiometer sets quiescent current to 125mA for the entire stage.

Rather than breaking a wire to insert an ammeter, quiescent current can be set by adjusting the

In this extract from his new book 'Valve Amplifiers', Morgan Jones looks at some of the classics in valve power amplifier design – including the Williamson.

second potentiometer for 18.75V across the 150Ω resistor; the amplifier should then be allowed to warm up and stabilise. The amplifier should then be switched off, and the true – hot – value of the 150Ω resistor measured. From this, a corrected voltage for 125mA could be calculated and set.

Linearity and drive capability of each stage is good due to the careful choice of operating points. Because this amplifier has four stages enclosed by the feedback loop, stability needs to be taken very seriously.

The input stage initially has an output resistance of 7.5kΩ, but this is raised by feedback to around 47kΩ. In combination with 12pF of input capacitance from the concertina, this gives a high frequency cut-off of about 280kHz.

However, this has been modified by adding step compensation components – an *R* and *C* in series – to the left-hand anode circuit of the dual triode. This circuit puts a step in the amplitude response which begins to fall at approximately 130kHz, but phase response remains virtually unchanged until 280kHz.

Output resistance of the concertina is approxi-

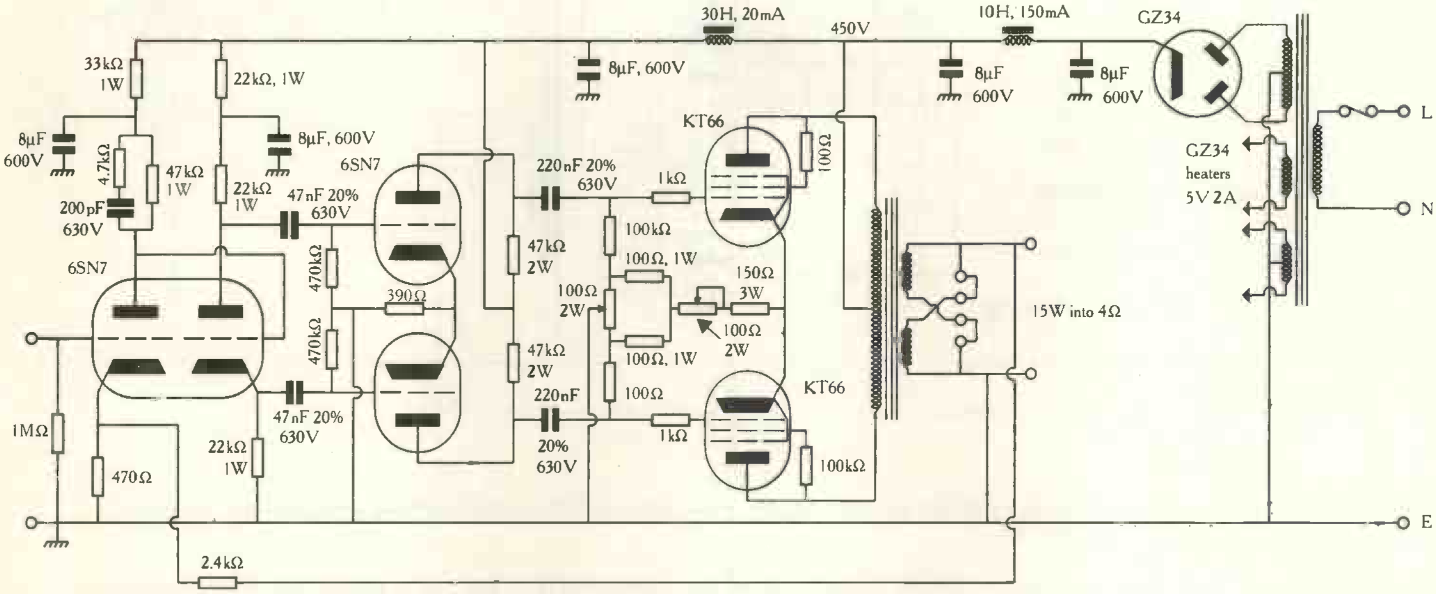

Fig. 1. Williamson's amplifier set a standard of performance that was years ahead of its time.

mately 1kΩ from the cathode and 22kΩ from the anode. As a result, there is an imbalance of high-frequency cutoffs in driving the 60pF input capacitance of the driver stage. This is because the anode cut-off is around 120kHz, while the cathode cut-off is 2.6MHz.

In practice, since the driver stage is a differential pair, as opposed to a cathode-coupled phase splitter, its gain will only fall by 6dB from 120kHz to 2.6MHz. At this point, it will fall at the normal 6dB/octave. The net result is very similar to the step compensation in the input stage. In this amplifier, adding a build-out resistor to the cathode will probably cause oscillation, and is not recommended.

The driver stage has an output resistance of ≈8.7kΩ; with 55pF of input capacitance from the output stage, cut-off is ≈330kHz, and the output transformer is specified to have a cut-off of 60kHz.

Step networks

The number of hf cut-offs within the feedback loop have not been minimised, and the dominant hf cut-off is too close to the next most dominant. The only remaining way to achieve stability at hf is to adjust the phase response independently of amplitude response by means of step networks.

At low frequencies it is more useful to consider time constants rather than –3dB points. Since the input stage is direct coupled to the concertina this can be ignored. The concertina feeds the driver stage with a *CR* of around 22ms, as does the driver to output stage.

The output transformer is set to 48ms. In view of this, it is not surprising that low-frequency stability is questionable – as was conceded in the original *Wireless World* article. A little of this may also be due to the fact that the input stage is not operated from the same ht as the concertina; in my experience, this fact alone can induce motorboating.

Remember that in 1947, circuits were designed using long multiplication, log tables, and even slide rules. Computer aided ac analysis was not an option. Most amplifiers were designed as carefully as possible, and then adjusted on test for best response.

Mullard's 5-20 power amplifier

This is a 20W 'ultra-linear' design from Mullard in 1955. It was introduced to sell the company's *EL34* pentode. There is a great deal of similarity between this design, the Mullard 5-10 (10W *EL84*) and some Leak amplifiers, **Fig. 2**.

The input stage is an *EF86* pentode. This device is responsible for the high sensitivity, but poor noise performance, of these amplifiers. Most of the cathode bias resistor is bypassed, since it would otherwise reduce gain from around 120 to 33. This would be a waste of open loop gain that could be used to correct distortion from the output stage.

Unadorned, the pentode has an output resistance of 100kΩ, and drives around 50pF of input capacitance from the phase splitter. This would give a cut off of 32kHz but it is modified by the usual compensation components across its anode load.

A slightly unusual feature is that the g_2 decoupling capacitor is connected between g_2 and cathode, rather than g_2 and ground. In most circuits, the cathode is at ac ground. As a result, there is no reason why the g_2 decoupling capacitor should not go to ground. In this circuit there is appreciable negative feedback to the cathode. Therefore g_2 has to be connected to the cathode in order to hold g_2–k (ac) volts at zero, otherwise there would be positive feedback to g_2.

The cathode coupled phase splitter is combined with the driver circuit using an *ECC83*; when loaded by the output stage, A_V is 54 for the *ECC83*, but overall gain is half this at 27.

The anode load resistors have not been modified to give perfect balance. With the 470kΩ grid leak resistors of the output stage in parallel with the 180kΩ anode loads, the effective anode load is 130kΩ. This means that the right-hand triode should have an ac anode load 3% higher, and R_L would then be 187kΩ. Mullard did actually state this, but probably assumed that most constructors would not have access to sufficiently high precision resistors to use the information.

A better solution is to use the 180kΩ value of anode load with the unloaded gain of 71. This then results in a value of 185.6kΩ, which still leaves the output resistances out of balance. Output resistance of left-hand triode is approximately 52.19kΩ, while the right-hand one is approximately 52.66kΩ, requiring a 470Ω build-out resistor.

The output stage has an input capacitance of about 30pF, combined with 53kΩ output resistance of the driver stage. This gives a poor cut-off at 100kHz.

Differential pair

Looking at the stage as a driver, investigate whether it is capable of driving the output stage. A total of 85V will be wasted across the 82kΩ tail resistor, but with 410V of supply rail, this still leaves you with 325V.

With component values given, this puts the operating point at 240V on the 180kΩ dc load line. Drawing the ac 130kΩ load line through this point shows that the stage would generate about 4% second harmonic distortion at full drive. This would result in an output of 18V rms, if it were not operated as a differential pair. Mullard claimed 0.4% distortion for the entire driver circuitry.

Although distortion appears satisfactory, the

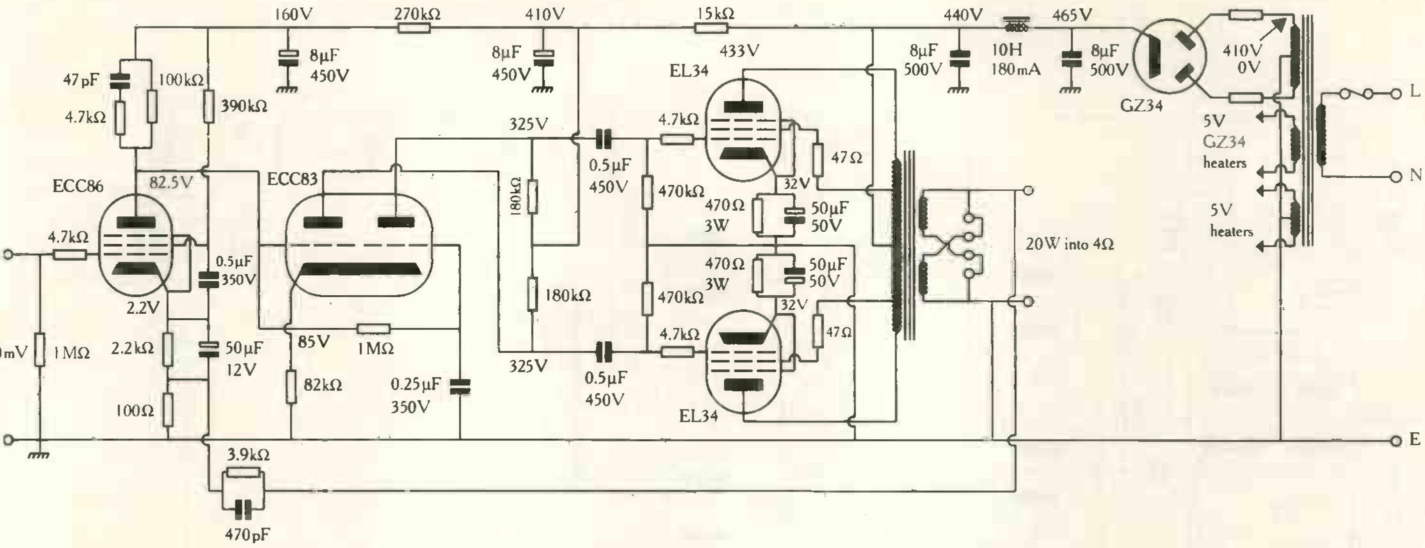

Fig. 2. Mullard's 5-20 was an ultra-linear design with high sensitivity but poor noise performance due to the input being pentode, rather than triode, based. Circuit courtesy Philips Components Ltd.

driver stage has only 10dB of overload capability. When output stage gain begins to fall – due to cathode feedback or input capacitance of the *EL34* loading the driver – global feedback will try to correct this by supplying greater drive to the output stage. This margin will quickly be eroded.

Driver circuitry was designed to produce an amplifier of high sensitivity – even after 30dB of feedback had been applied. This has forced other factors to be compromised. Whereas the Williamson sacrificed stability for linearity, the Mullard 5-20 achieves stability at the expense of linearity.

Ultra-linear output

The output stage is a pair of *EL34* in 'ultra-linear' configuration, with 43% taps for minimum distortion. Unlike the Williamson, there is no provision for adjusting or balancing bias, and this might seem to be a backward step.

Bias adjustment implies connecting the cathodes together and using a proportion of grid bias to provide balance adjustment. Because biasing is firmly set by the potentiometers, there is no self-regulation of bias current. As the valves age, balance will need to be reset.

In short, providing this adjustment ensures that it has to be used regularly. By contrast, the Mullard 5-20 has separate cathode bias resistors and relies on automatic bias to hold the anode currents at their correct, and therefore equal, levels.

In practice, this works quite well, although it does not quite achieve the low transformer core distortion of a freshly balanced adjustable system.

System drawbacks

A disadvantage of this system is that the individual cathode bias resistors applies series negative feedback to the output valves, raising their output impedance.

The output transformer could be redesigned to maintain the match to the load, but this is undesirable as it would require a higher primary-to-secondary turns ratio. This makes a high quality design more difficult to achieve.

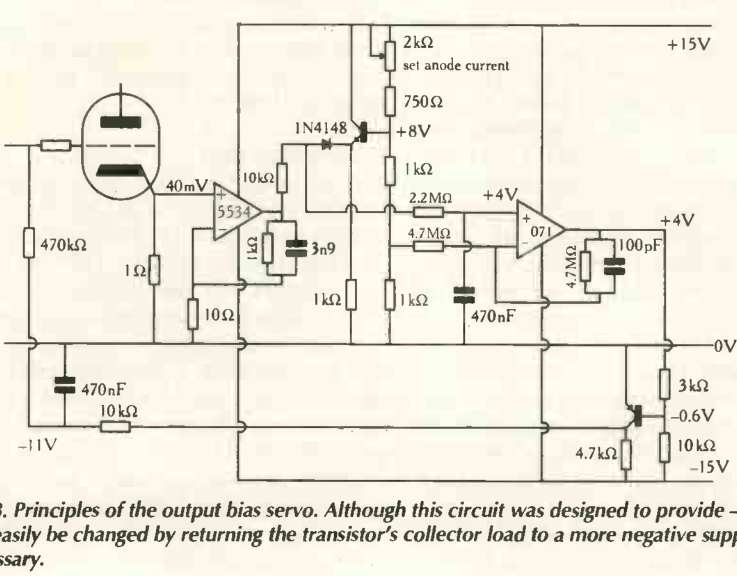

Fig. 3. Principles of the output bias servo. Although this circuit was designed to provide –11V bias, it can easily be changed by returning the transistor's collector load to a more negative supply, as necessary.

Because of this, the cathode bias resistors are usually bypassed by capacitors, resulting in several problems.

The capacitor is a short circuit to ac and so prevents feedback. But, as in the simple common cathode triode amplifier, at very low frequencies it will no longer be a short circuit, and will allow feedback.

Because the output stage is load matched, feedback causes an immediate rise in distortion and reduction of output power due to the mismatch. The obvious solution to this is to fit a large enough capacitor to ensure that the low frequency cut-off for this combination is below all frequencies of interest, say 1Hz.

Remembering that the resistance that the capacitor sees is R_k in parallel with r_k, you can calculate the value required. For a pentode, r_k is $1/g_m$; a typical output pentode has a g_m of 10mAV^{-1} at its working point, so r_k of around 100Ω. This is in parallel with a bias resistor of 300Ω, giving a total resistance of 75Ω. For 1Hz, a capacitance of 2000μF is needed.

Capacitors rated at 2000μF, 50V were simply not available at the time, and were not fitted. They are readily available now, but there are two reasons why you might wish to use a smaller value.

- A 2000μF capacitor will have considerable inductance, allowing feedback at high frequencies. By using low inductance electrolytics designed for use in switch-mode power supplies, and/or by bypassing with smaller values, this problem can be overcome.

- If the output stage is driven into Class B by overload, each cathode then tries to move more positively than negatively. It cannot turn off any further, but it can certainly turn on harder. The capacitor smooths these changes into a gently rising dc bias voltage, which biases the valve further into Class B, and the problem continues.

The effect of this is that a momentary overload can cause distortion of the following signals – even though they would normally have been within the capabilities of the amplifier. As the cathode bias capacitor becomes larger, this recovery time from overload lengthens. Theoretically, one never overloads amplifiers, and this would not be a problem, but occasional overload is inevitable, and should be considered.

The ideal way to deal with all of these problems is to reduce the cathode bias resistor to an ohm or less such that it no longer causes noticeable feedback, and measure the current through it using an op-amp. This signal then feeds an asymmetrical clipper. When the valve strays into Class B and clips one half-cycle, the clipper removes an equal amount from the other half-cycle before feeding the processed signal to an integrator. The integrator can have an *RC* time constant of almost any value – 10s is not unusual.

Output of the integrator is a smoothed dc voltage proportional to anode current. This can then be compared to a fixed reference. The difference between the two levels drives an amplifier whose output feeds negative grid bias to the output valve.

If anode current of one valve is set as a reference, then other valves can share this reference, forcing the anode currents into balance. Increased complexity of this scheme is partly offset by its improved performance and reduction in ht voltage required, since the cathode bias scheme wastes ht, **Fig. 3**.

Sensing anode current

Figure 3 was designed to sense a 40mA anode current by developing 40mV across the 1Ω resistor. As this circuit is based on the 40mV signal, the sense resistor should be changed to suit if a different current is to be sensed.

The *5534* has a gain of 100, and amplifies the mean dc level to 4V, with ac peaks rising to 8V. Any peak above 8V is clipped by the diode/transistor clamp; the other half-cycle will already have been clipped by the valve.

The clipped signal is integrated by the 2.2MΩ resistor in combination with the 470nF capacitor, giving τ=6.5s. The *071* compares this smoothed dc with a reference derived from the potential divider chain, and uses this to control the bias transistor. Reference and clamp voltages are made adjustable by the 2kΩ variable resistor in order to allow for fine adjustment of anode current.

Although this circuit was designed to provide –11V bias, this can easily be changed by returning the bias transistor's collector load to a more negative supply as necessary; no other changes are required.

Quad II power amplifier

The Quad II is an unusual design, which at first sight does not look too promising, but works because the design is synergetic. In this design, the phase splitter has been combined with both the driver stage and input stage.

In order to achieve the necessary gain, pentodes have been used. Output impedance is therefore high, as is input noise. To make matters worse, a variation of the see-saw phase splitter has been used. The output stage has local feedback, which increases the voltage swing required to drive it, **Fig. 4**.

Driving the loudspeakers

A pair of *KT66* beam tetrodes with anode and cathode loads split in the ratio 9.375:1 comprise the output stage. The cathode therefore provides little drive to the loudspeaker. This may be considered to be series feedback from the output transformer. Cathode current in the output transformer however is the sum of the anode and g_2 currents. It has been found that this summation reduces third harmonic distortion by a further 8dB over that due to the negative feedback.

The effect of this feedback on output resistance is the opposite to what might be expected (Williamson and Walker, 1954). If a cathode resistor is left unbypassed it will generate series feedback which increases r_a, whereas the transformer coupled feedback reduces r_a. This can easily be explained by assuming a short circuit as a load.

Clearly, the output stage will be unable to drive any voltage into this load, but conversely, there will be no feedback signal applied to the cathodes. The grids are then driven by the full input signal, rather than the input signal minus the feedback. Therefore the output stage is driven harder as it attempts to maintain its output into a short circuit. This action is directly equivalent to reducing output resistance. The new value of output resistance can be found using the normal feedback equation.

Transformer primaries are equivalent to 3kΩ anode to anode. With tetrodes, this low value of anode load results in a reduction of third harmonic distortion, and an increase in second harmonic. This is then cancelled by the output transformer.

There is no provision for balancing anode current, and the automatic bias is shared. As a result, you can expect an increase in distortion at low frequencies due to saturation of the transformer core. Curiously, the cathode resistor was only rated at 3W, yet it dissipates 3.8W. If your Quad II is distorting, a burnt out cathode bias resistor may well be the cause.

Even with pentodes, there is not a great deal of gain from the driver circuitry, and input sensitivity is low, at 1.4V for full output. This is an excellent choice of input sensitivity for a power amplifier. It not only guarantees impeccable noise performance – even from a pen-

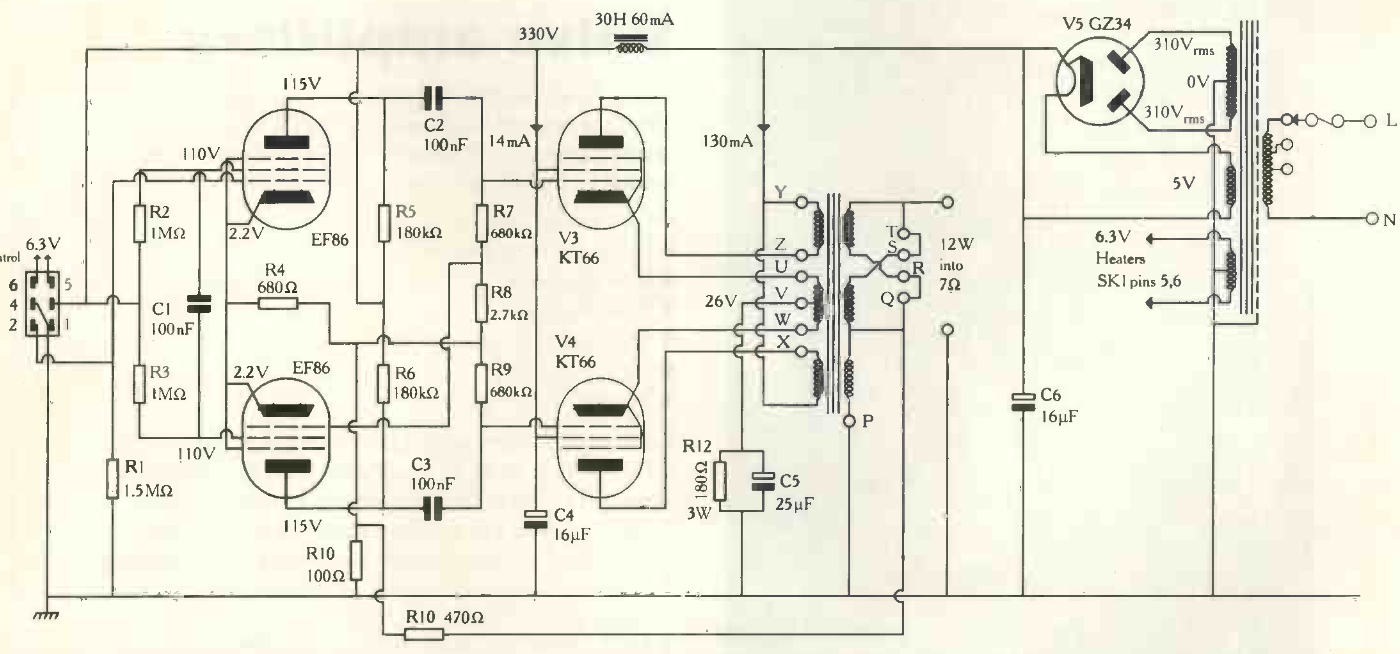

Fig. 4. The Quad II is different. Not only is the phase splitter combined with the driver, but it is also combined with the input stage. Circuit courtesy Quad Electroacoustics.

tode – but it also means that the input is far less susceptible to hum and noise from input cables or heater circuitry.

The Quad II was only beaten in signal-to-noise performance by the Williamson, which was quieter because it had a triode input stage.

Balanced output

Although the phase splitter is a variation of the see-saw phase splitter, it does not rely on feedback for balance, and its operation is quite elegant. Output valves must each have a grid leak resistor, so instead of applying additional loading to the driver valves, a tapping is taken from one of these to provide the input for V_2.

In theory, if this tapping had an attenuation equal to the gain of V_2, then the output of the phase splitter would be balanced. Because of component variation, this will not always be true, and so the cathodes of the two valves are tied together to improve balance.

This has been further modified by applying global feedback to the other end of this potential divider, which is why the gain of V_2 is not equal to 680kΩ/2.7kΩ.

Pentode over triode

Pentode stages have output resistance approximately equal to R_L. Since R_L for the Quad is 180kΩ, this would appear to be very poor at driving the 30pF input capacitance of the output stage, resulting in a cut-off of around 30kHz.

Apart from the output transformer, this is the only high-frequency cut-off in the circuit, and is therefore not a problem. Each output valve requires a swing of around 80V pk-pk. This is easily provided, because pentodes can approach 0V more closely than triodes. Also, *LC* filtering is used on the ht line, rather than *RC* filtering, which would reduce available ht.

This *LC* filtered supply also feeds g_2 of the output valves. This has the valuable advantage of reducing hum, since the anode current of a tetrode or pentode is dependent on g_2 voltage rather than anode voltage.

In the input stage, pentodes need to have g_2 decoupled to ground. Instead of each valve having a capacitor to ground, one capacitor is connected between each g_2. This has three advantages:

- If you had two individual capacitors, they would effectively be in series with a centre tap to ground. Since each valve is supplying an equal but opposite output, the centre tap would be at ground potential even if it were to be disconnected from ground. Disconnecting the centre tap from ground results in two capacitors in series. These can be replaced by a single capacitor whose value is equal to half that of one of them.

- Since this one capacitor is connected between two points of equal potential, it need not necessarily have the full voltage rating to ground. However, it is as well to consider the effect of fault conditions when determining the voltage rating. as a result, this is not a great advantage.

- Connecting g_2 of each valve together at ac helps maintain balance in the same way as commoning the cathodes.

Although substituting one stage that combines the functions of input, phase splitter and driver does not achieve the linearity of purpose designed stages, it achieves better linearity than the Mullard circuit. This is because less gain is demanded from it.

With only a simple driver circuit and output stage within the feedback loop, the Quad II has no stability problems. ■

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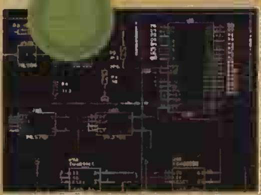

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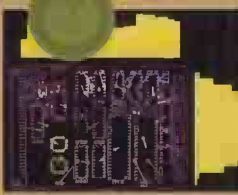

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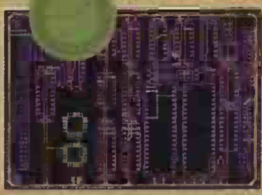

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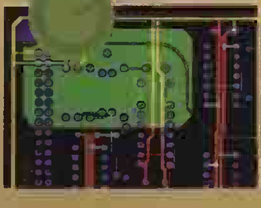

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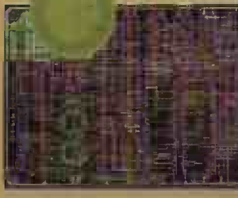

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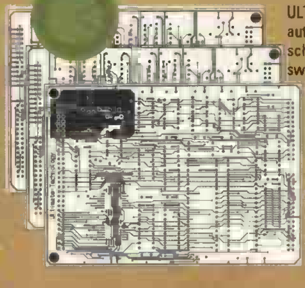

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Signal processing in a toolbox

Allen Brown reviews an extension to the maths modelling package Matlab, which is intended for designers of dsp circuitry.

PC software package *Matlab* – reviewed in *EW+WW*, March 1995 – is a product designed, amongst other things, for the modelling and simulation of physical processes. Whether solving electrical, biological, mechanical or economical problems, *Matlab* is a versatile tool for tackling a whole range of modelling requirements.

To support the package, its makers, Math Works, provide a range of 'Toolboxes' designed for specific applications. A number of these – one of which is the signal processing toolbox – are likely to be of interest to electronics engineers .

Over the past ten years, signal processing has aroused a considerable amount of interest, due in part to the availability of high-performance digital signal processors, both fixed and floating point. Also, the pc is playing a significant role in realising signal processing operations. The wealth of high speed data acquisition expansion cards and the availability of high quality software is contributing to the continuing interest in signal processing.

Normal engineering practice involves the modelling of electronic circuitry before it is constructed and it is customary to use such software packages as *PSpice* or *Electronics Workbench* for this purpose. For signal processing modelling the designer, who is familiar with *Matlab*, has the option of using the signal processing toolbox. This comprises a library of prewritten signal processing functions that can be evoked from the command line within the *Matlab* environment.

The toolbox has over one hundred functions whose operations cover the spectrum of commonly used processes. Groups of processes include filter analysis, linear systems transformations, digital filter design, transforms, statistical processing, windowing and parametric modelling. In fact, the range is as extensive as you are likely to find in any pc signal processing software.

Analysing filters

There is a number of interesting tools in the filter analysis group for analysing systems. It should be emphasised that the toolbox is not constrained to digital operations only – it is also possible to model system behaviour in the analogue Laplace S domain. For example, determining frequency response of the following transfer function,

$$H(s)=\frac{0.2s^2+0.3s+1}{s^2+0.4s+1}.$$

The command line code for this task is,

```
b=[0.2 0.3 1];a=[1 0.4 1];
w=logspace(–1, 1);
freqs(b, a, w)
```

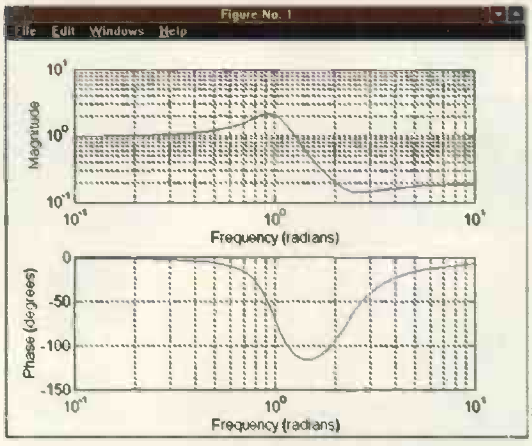


Fig. 1. Using the signal processing toolbox, a transfer function in the Laplace S domain can be easily realised in a plot.

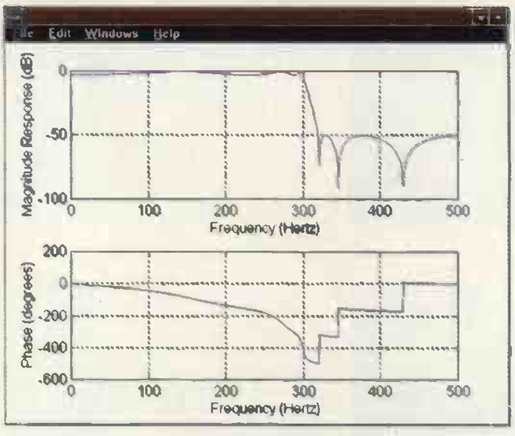


Fig. 2. Transfer function generated graphically in the digital z domain.

The graphical result, **Fig. 1**, shows the magnitude of the response and the phase in degrees. Coefficients of the transfer function are defined in the first line and log spacing for the plot in the second line. The freqs function in the third line performs plotting in the frequency domain.

Plotting the response of a digital filter is just as easy, consider a sixth-order lowpass elliptical filter with a cut-off frequency of 300Hz and a 3dB ripple sampling at 500Hz. The command line code would be,

```
[b, a]=ellip(6, 3, 50, 300/500);
freqz(b, a, 512, 1000)
```

The first line defines the specifications of the filter, and freqz in the second line performs transfer and plotting in the *z* domain. The graphical result is shown in **Fig. 2**. You will also notice that a pleasing feature of the phase plots of the signal processing toolbox is the absence of fly-backs at the 360° boundary.

Fly-backs are customary on almost all signal processing software and confuse the information in the phase plot. It is good to see that this product eliminates this problem. The user may be interested to see the impulse response of the filter. This can be accomplished via the impz function. For example,

```
[b,a] = ellip(10, .05, 80, .4);
impz(b, a, 50)
```

The command line code defines a tenth-order elliptical filter with a 0.05 ripple in the pass band, a 80dB attenuation in the stop band with a normalised cut off frequency of 0.4. A normalised frequency of 0.5 is equivalent to half the sampling frequency. The graphical result of the above instructions is shown in **Fig. 3**. Including the instruction zplane(a, b) produces the pole/zero plot of **Fig. 4**.

Alternatively within the linear systems transformations group of instructions there are some powerful operations for changing one system representation into another. These can be evoked with relative ease. For example, state-space to transfer function, ss2tf, or zero-pole to state-space, zp2ss. All in all some useful features for analysing system responses in both analogue and digital domains.

Designing digital filters

Functions for designing finite impulse response and infinite impulse response digital filters are included in the signal-processing toolbox. Infinite impulse response filters comprise the well known Bessel, Butterworth, Chebychev, elliptical and the not so well known Yule-Walker. There are also instructions for determining the order of the filters.

This example illustrates how the instructions are used. Consider designing a bandpass Chebychev II filter (ripple in the stop band), transmission from 100-200Hz with 3dB ripple and a stop band attenuation of 100dB. Given a sampling frequency of 1000Hz, the instructions are shown with comments in **Table 1** and graphical output for the filter in **Fig. 5**.

Once the user has learned the command line syntax, it is a relatively straight forward task to design filters. However, there does not appear to be any function for quantising coefficients – usually to 16bit – and converting them for use with a fixed point format for digital signal processing. It is sometimes necessary to model the performance of the filter once the coefficients have been quantised – there appears to be no function in the signal processing toolbox for performing this operation directly.

The option of designing finite impulse response filters using the Parks-McClellan method, based on the Remez Exchange Algorithm, is not as clean as it could be. There

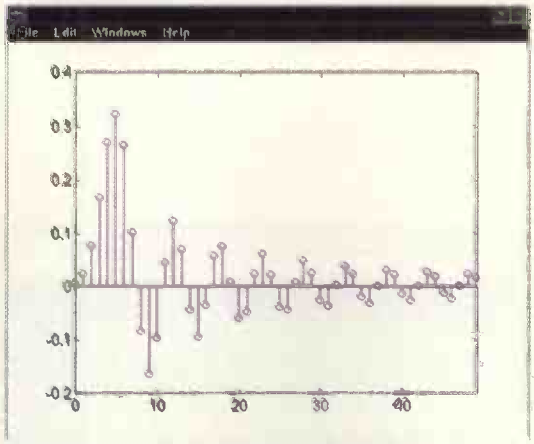


Fig. 3. When subjected to a single impulse, the impz function allows the response of the system to be visualised with ease.

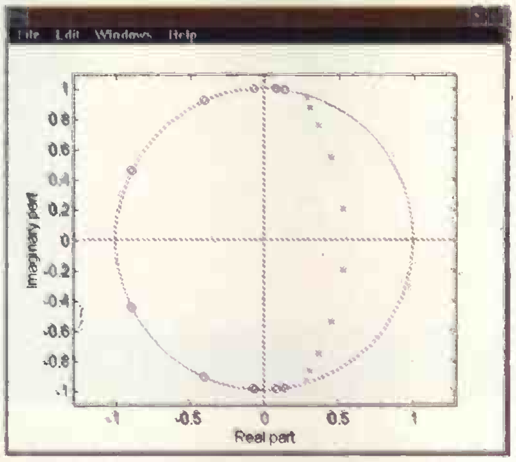


Fig. 4. Determining the positions of the poles and zeros in the z-plane unit circle gives information on possible problems that may occur for poles lying close to the perimeter. If through quantisation effects they should stray outside the unit circle the system becomes unstable.

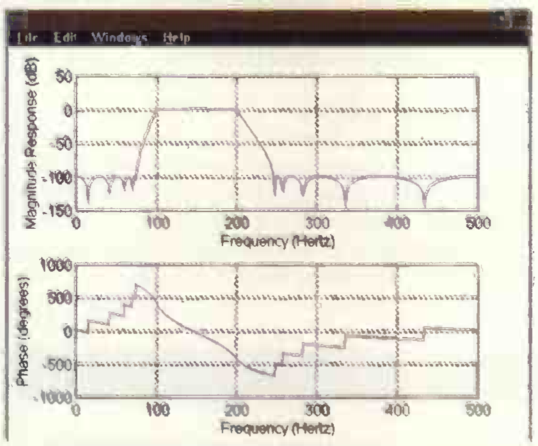


Fig. 5. Graphical results from the design of a digital filter using the instructions in Table 1. This example shows that Elliptical filters are unsuitable for systems which require linear phase response.

Table 1. Instructions required to design a bandpass Chebychev infinite impulse response digital filter.

```
% Specify the passband and the stopband corner frequency ranges,
      Wp=[100 200]/500;Ws=[50 250]/500
% Specify the attenuations in decibels
      Rp=3;Rs=100;
% Calculate the filter order n and the actual corner frequencies
      [n, Wn]=cheby2ord(Wp, Ws, Rp, Rs);
% Perform the design and calculate the raw coefficients
      [b,a]=cheby2(n, Rs, Wn);
% Calculate the coefficients in terms of poles and zeros
      [z, p, k]=cheby2(n, Rs, Wn);
% Convert the pole zero values into second order coefficients
      sos=zp2sos(z, p, k);
% Plot the frequency characteristics of the filter
      freqz(b, a, 512, 1000)
```

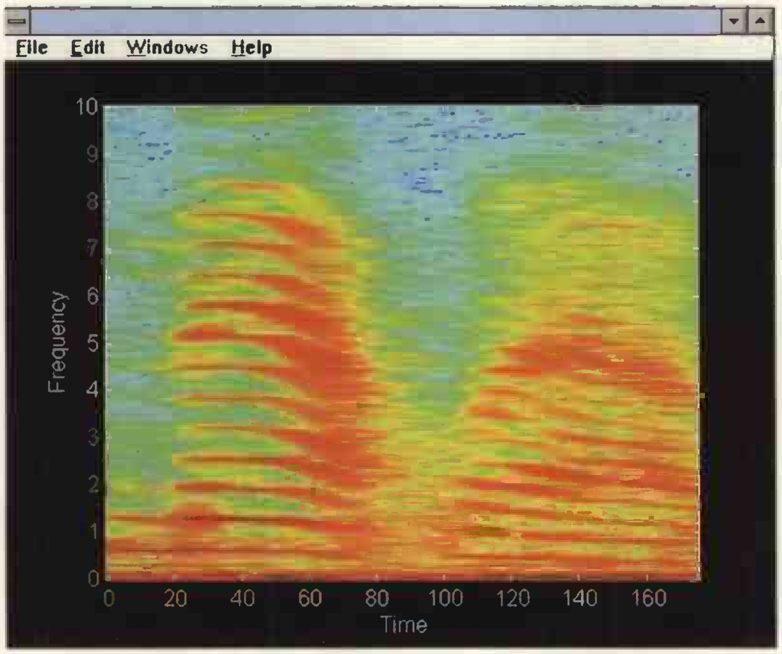
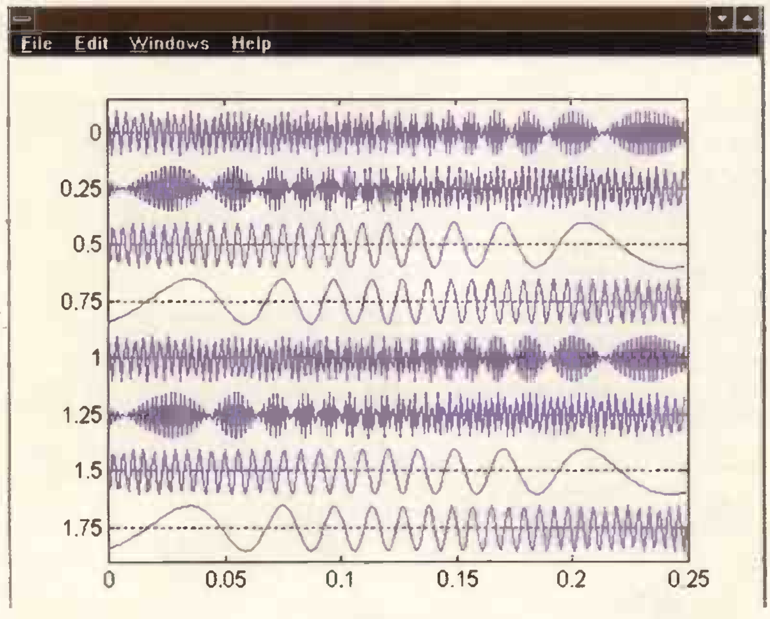


Fig. 6. Spectrogram of the word matlab. The function specgram can be used with great effect to analyse speech signals.

Fig. 7. A frequent requirement for analysing rapidly changing signals is dissecting a signal into several equal segments. This is achieved using a strip plot.



is now a well established method of specifying the required filter characteristics. However the two functions remez and remzord in the toolbox have to be set up in a manner which departs from conventional wisdom thus rendering them less amiable than they could be.

Analysis of spectra

The toolbox has a number of useful functions for investigating the spectral content of signals, both periodic and signals contaminated by noise. In addition to the usual FFT, DFT there is a choice of the Chirp z-transform and a reasonable array of window functions to pre-process the data before performing spectral analysis.

An interesting function also included is *specgram* ie spectrogram for analysing speech signals. It displays frequency and magnitude information *versus* time. **Figure 6** is an example of the spoken word *matlab*. Magnitude is colour coded – red for the largest magnitude and blue for the smallest. For noise contaminated signals, the user can first perform auto-correlation, xcorr, on the data before subjecting to the Fourier transform. In many instances this improves performance of the spectral analysis.

Generally the toolbox has a reasonable array of spectral analysis tools, however one deficiency is the waterfall function. This function would partition a signal into several segments, perform a FFT on each segment and display a three dimensional plot – magnitude *versus* frequency *versus* time. There is however a feature for partitioning a signal into segments only and displaying them as a strip plot, strips. An example of its output is shown in **Fig. 7**.

The signal processing toolbox possesses a number of alternative tools for deriving spectral information. These are categorised as parametric modelling and are useful for analysing non-stationary signals – for example speech.

User manual

The manual accompanying Toolbox is well written, beginning with a tutorial section covering each group of the signal processing toolbox's functions. In the reference section of the manual there are the all important examples which can save an enormous amount of time when trying to master their use. The manual is quite indispensable when using this package.

Conclusion

Most functions in the signal processing toolbox can be accessed with relative ease and incorporated into a user's *Matlab* model. They are not as comprehensive as I would like and there are some noticeable omissions such as adaptive filters.

There could also be more functions for implementing lattice filters. However, in general this package will prove a very useful tool for the electronics engineers involved in digital signal processing and systems design. ■

Availability

Toolbox is available from Rapid Data, Crescent House, Crescent Road, Worthing, West Sussex BN11 5RW. Tel. 01903-202819, fax. 01903-820762. Price of *Matlab* package is currently £1376 and Signal Processing Toolbox is £325, both excluding VAT. These prices include documentation and technical support. Add £25 for delivery of *Matlab* with the signal processing Toolbox or £15 for the Toolbox alone.

Transparent i/o via LPT

The main disadvantage of using the printer port for analogue and digital i/o is that you have to unplug the printer. Or do you? Alan Bradley explains.

The usual way of expanding a pc is via an internal card. This has the disadvantage of requiring an accurately made double-sided pcb with gold-plated edge connectors, and the designer needs to understand the pc's 8/16bit expansion bus signals.

An easier way is to use the Centronics parallel printer port. This port is standard, whereas internal expansion buses vary from ISA, PCI, MCA, to none at all. Extra Centronics printer cards are very cheap – between £9 and £18, and a pc can support up to three such ports if required.

In this design, a printer pass through facility is included, allowing a printer to share the same port. This obviates the need to open the pc to insert an extra LPT expansion card, and prevents the possibility of address clashes.

The printer port has eight data output lines. Since most pcs don't have bidirectional data lines, the printer port's status lines are used as input lines instead. For example, 8bit data can be read as four bits at a time using four of the port's five input status lines.

The LPT port also has four control output lines. These can be used as digital output lines and for selecting either the upper or lower input nibble.

The pc's LPT port

Each LPT parallel printer port has three 8bit registers – the data latch, the status register and the control register.

Writing to the data register, $d_{7\text{-}0}$, causes the byte sent, to be latched and to appear on the parallel port's 25way D connector, pins 2-9. In this design, this data is then sent to either the d-to-a converter or digital output port.

Normally, reading this register returns the contents of the latch. The status register, $b_{7\text{-}0}$, is a read-only register holding the state of BUSY, /ACK, PE, SLCT and /ERROR input lines from the printer. These are on lines $b_{7\text{-}3}$ respectively. Input BUSY is inverted between the D-connector, pin 11, and the register. Bits $b_{2\text{-}0}$ are unused.

The control register is an output latch holding the four printer control signals – /SLCT-INP, /INIT, /AUTOFEED and /STROBE. These are bits $b_{3\text{-}0}$ respectively. Interrupt bit b_4 determines whether an interrupt is generated by a falling /ACK input – 0 represents interrupt disabled. I always disable this bit as it is rarely used by printer software. This leaves IRQ channels 5 and 7 available for other expansion cards.

Bits 0, 1 and 3, ie /STROBE, /AUTOFEED and /SLCT-INP, are inverted between register and the D-connector output pins. This inversion is corrected again when the control register is read. Parallel port signals are at ttl levels and are presented on a 25-way D connector, shown on the circuit diagram.

The pc parallel port is specified as having open collector outputs, but the pc/at version uses standard ls-ttl devices. The three sets of addresses that can be assigned to parallel ports LPT1, LPT2 and LPT3 are,

Groups of pc i/o addresses reserved for LPT use.

Port	Data	Status	Control
LPT(n)	378_{16}	379_{16}	$37A_{16}$
LPT(n)	278_{16}	279_{16}	$27A_{16}$
LPT(n)	$3BC_{16}$	$3BD_{16}$	$3BE_{16}$

Note that register addresses assigned to each parallel port vary depending on the pc type.

Information on parallel port addresses and the number of ports fitted can be found in the Bios data area,

Bios addresses containing information on pc parallel port address locations.

Address	Description
$0{:}408_{16}$	Base address of parallel port 1 (LPT1), low byte
$0{:}409_{16}$	High byte for above
$0{:}40A_{16}$	Base address of parallel port 2 (LPT2)
$0{:}40B_{16}$	High byte for above
$0{:}40C_{16}$	Base address of parallel port 3 (LPT3)
$0{:}40D_{16}$	High byte for above
$0{:}410_{16}$	Hardware info., bits 15-14 = No of LPT ports
$0{:}411_{16}$	High byte for above

Useful ICs

While designing the interface, I have found the following 74 series ICs useful. Each is available in a variety of forms, including LS, ALS and HCT. The type chosen will depend on a number of criteria, including budget, speed and power consumption.

The '139 is a 2-4 line decoder useful for selecting between four interface i/o ports while the '138 decodes three lines to eight and is useful for applications with eight ports. Selection between the upper and lower nibbles of the input byte can be carried out by a '157 quad 2-1-line multiplexer. Alternatively, a '241 dual or '244 quad tristate buffer can select the upper or lower nibble.

A '573 or '373 octal transparent latch can form an 8bit digital output port. With its enable pin high, the outputs follow the inputs; when it is low outputs are held latched. For 8bit digital input, a '245 octal transceiver can be used. A

Pin-out of the 25way female D connector used on the pc for LPT i/o.

Pin(s)	Bit	Status
1	STROBE	output
2-9	d0-d7	output
10	ACK	input
11	BUSY	input
12	PE	input
13	SLCT	input
14	AUTOFEED	output
15	ERROR	input
16	INIT	output
17	SLCTINP	output
18-25	GND	

'541 or '244 buffer can also be used here, depending on which produces the most convenient pcb layout.

When reading data through the status port, the BUSY bit must be inverted, for example by exclusive-or-ing it with 1000000_2. On fast machines, the interface-controlling program may need software delays to allow the parallel port control lines to settle after a change.

Design details

This design provides a fast 8bit a-to-d converter with integral sample and hold, a fast 8bit d-to-a converter and a pair of 8bit digital i/o ports. The a-to-d and d-to-a converter are both capable of audio sampling and playback.

In my prototype, the lower D-type plug shown in the drawing was pcb mounted. This plug connects to the pc printer port. The upper D-type, also pcb-mounted, connects to the printer, providing the 'pass through' facility.

Note that IC_6 – a *74LS157* quad 2 to 1 line multiplexer – must be a ttl type as its inputs are left floating when neither the a-to-d converter nor digital input port are selected. In addition, $IC_{1\text{-}4}$, and IC_9 should be ls-ttl types, as cmos ICs would require input static protection circuits.

The remaining pcb-mounted D-type socket brings out both 8bit digital ports, the digital ground rail and +5V into the real world.

Analog Devices *AD7569JN*, IC_{10}, is an 8bit a-to-d converter with integral sample and hold facilities and configured for mode 2 operation and for an input range of 0-2.5V. Bringing the read line low when chip select is active low selects the a-to-d converter, starting the conversion. The converter's busy line is not used so a software delay is needed to determine when the conversion will be complete. Conversion time should be less than 2.6μs.

Eight-bit a-to-d converter IC_{11}, a *ZN558*, has an integral transparent latch and a range of 0-2.5V. Single supply op-amp IC_{12}, an *LM358*, buffers the d-to-a converter output and its 2.5V reference.

Resistors $R_{25,26}$ help prevent the op-amp from oscillating due to cable capacitance and also protect it from short circuits. A dc path to ground to prevent cross-over distortion is provided by R_{24}.

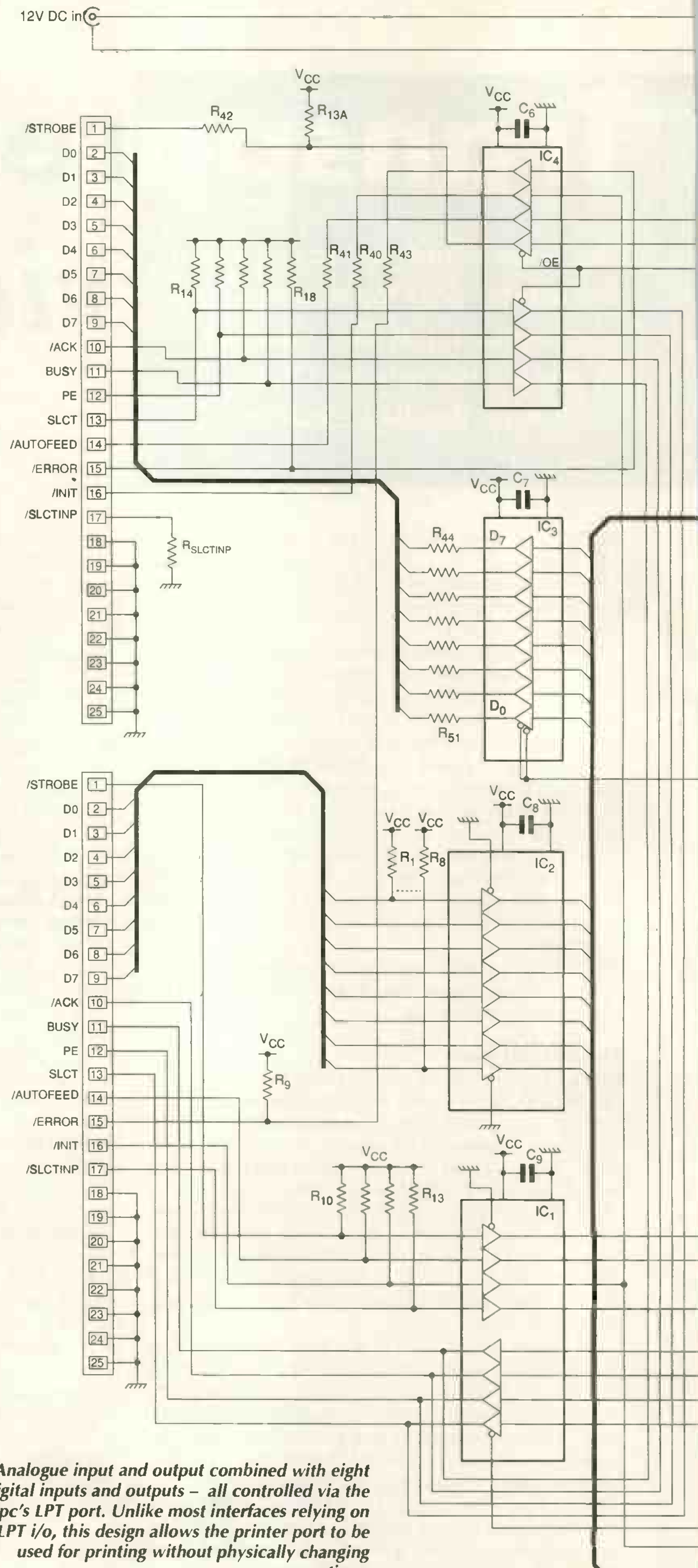
A further *LM358*, IC_{13}, is configured as a unity gain buffer to protect the *AD7569* voltage input. The *7569*'s input pin sources approximately 10μA and R_{27} helps the op-amp sink this near 0V.

A ready built 1A, 12V regulated power supply is ideal. Diode D_1 provides reverse polarity protection. A *7805*, IC_{14}, produces a +5V rail. An analogue 11.3V power rail was provided with a separate ground rail, which is also used for all the interface's analogue ICs.

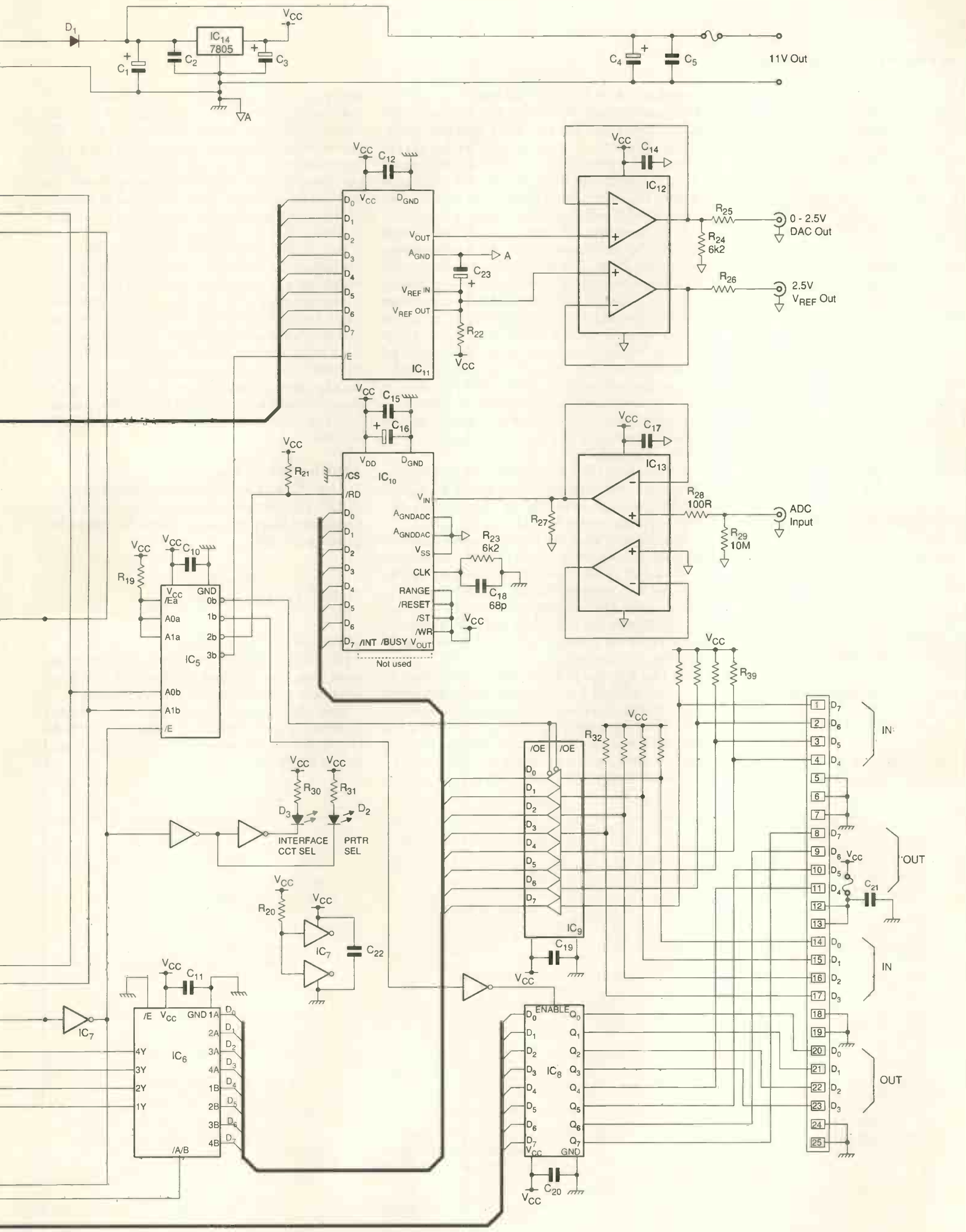
Within the pc, these control and status bits manage data transfer to and from the LPT analogue and digital i/o circuitry.

Bit	Control	Description
7	*	unused
6	*	unused
5	*	unused
4	IRQ DISABLE	0 disables when a & d i/o is selected
3	SLCTINP	0 selects a & d i/o, 1 selects printing
2	INIT	0 selects lower nibble 1 selects upper nibble
1	AUTOFEED	0 0 1 1
0	STROBE	0 d-to-a 1 a-to-d 0 dig o/p 1 dig i/p

Bit	Status	Description
7	BUSY	bit 3 from LS157 mpx
6	ACK	bit 2 from LS157 mpx
5	PE	bit 1 from LS157 mpx
4	SLCT	bit 0 from LS157 mpx
3	ERROR	not used by i/o card
2	*	unused
1	*	unused
0	*	unused



Analogue input and output combined with eight digital inputs and outputs – all controlled via the pc's LPT port. Unlike most interfaces relying on LPT i/o, this design allows the printer port to be used for printing without physically changing connections.


D1
IC14
7805
VCC
C1
C2
C3
C4
C5
11V Out
A
VCC
C12
D0 VCC DGND
D1
D2
D3
D4
D5
D6
D7
/E
VOUT
AGND
VREF IN
VREF OUT
C23
R22
VCC
IC11
VCC
C14
IC12
R25
R24
6k2
0 - 2.5V
DAC Out
R26
2.5V
VREF Out
VCC
C15
C16
VDD DGND
R21
/CS
IC10
/RD
VIN
D0
D1
D2
D3
D4
D5
D6
D7
AGNDADC
AGNDDAC
VSS
CLK
RANGE
/RESET
/ST
/WR
/INT /BUSY VOUT
Not used
R23
6k2
C18
68p
VCC
VCC
C17
IC13
R28
100R
R27
R29
10M
ADC
Input
VCC
VCC
C10
R19
VCC
/Ea
GND
0b
1b
2b
3b
A0a
A1a
IC5
A0b
A1b
/E
VCC
VCC
R30
R31
D3
D2
INTERFACE
CCT SEL
PRTR
SEL
VCC
VCC
R20
IC7
C22
VCC
R39
R32
VCC
/OE /OE
D0
D1
D2
D3
D4
D5
D6
D7
IC9
C19
VCC
IN
OUT
IN
OUT
VCC
C21
1 D7
2 D6
3 D5
4 D4
5
6
7
8 D7
9 D6
10 D5
11 D4
12
13
14 D0
15 D1
16 D2
17 D3
18
19
20 D0
21 D1
22 D2
23 D3
24
25
VCC
C11
/E VCC GND
1A
2A
3A
4A
1B
2B
3B
4B
4Y
3Y
2Y
1Y
IC6
/A/B
IC7
D0
D1
D2
D3
D4
D5
D6
D7
ENABLE
D0 Q0
D1 Q1
D2 Q2
D3 Q3
D4 Q4
D5 Q5
D6 Q6
D7 Q7
IC8
VCC GND
C20
VCC

I/O interface key components

IC_{1-4}	*74(A)LS244*
IC_5	*74(A)LS139*
IC_6	*74(A)LS157*
IC_7	*74(A)LS04*
IC_8	*74(A)LS373*
IC_9	*74(A)LS244*

For easier pcb layout, broadside ICs could be used – for example IC_2 *74LS541*, IC_3 *74LS245*, IC_8 *74LS573*, IC_9 *74LS245*.

IC_{10}	*AD7569JN* a-to-d
IC_{11}	*ZN558* d-to-a
$IC_{12,13}$	*LM358*
IC_{14}	*7805*, 1A version
D_1	*1N4001*
D_2	led 5mm green
D_3	led 5mm red
$C_{1,3,4}$	100μF
$C_{2,5,8-15}$	0.1μF cer.
$C_{17,19-22}$	0.1μF cer.
C_{16}	22μF 25V
C_{18}	68pF polystyrene 2%
C_{23}	4.7μF tant.

R_{1-8}, R_{10-13}, R_{13A-18}, R_{32-39}, 4k7 SIL8

R_9, R_{19}, R_{20}, R_{21}	4k7
R_{22}	390R
$R_{23,24}$	6k2
$R_{25,26,28}$	100R
R_{27}	330R
R_{29}	10M
$R_{30,31}$	led resistor 2k2
$R_{SLCTINP}$	220R
R_{40-51}	33Ω resistors required by printer pass through circuit when printer is switched off.

Fuses shown are 100mA a/s.

Interface board register usage

Data lines d_{0-7} are used for the d-to-a converter and digital output port. The four status input lines, BUSY, ACK, PE and SLCT, b_{7-4}, are used to read the upper or lower nibble of the input byte. The four control output lines are used to select either the upper or lower input nibble. They also control printer pass through, analogue input or output and digital input or output.

In the control register, bits 0, 1, and 3 are inverted between the control register and corresponding D-connector output pins. The SLCTINP line is normally used to select or deselect the printer, a low level signifying selection. It has therefore been used to switch between the analogue and digital i/o circuitry and the printer pass-through facility, selected when SLCT-INP is low as normal. Analogue and digital i/o circuitry is selected when SLCTINP is high, and the printer is deselected.

Consequently, when b_3 of the control register is high, the corresponding pin 17 on the D connector is low and $IC_{3,4}$ are selected. At the same time, the four status line buffers of IC_1, b_{7-4}, are tri-stated and IC_5's enable pin is held high.

The pc parallel port is now connected via buffers to the printer thus allowing normal use of the printer. Resistors R_{40-51} in series with the output buffers driving the printer need to be 33Ω.

Alternatively, when line b_3 of the control register is low, the corresponding pin 17 of the D connector is high and $IC_{3,4}$ are tri-stated, isolating the printer. Resistor R_{13A} pulls the STROBE line high when $IC_{3,4}$ are tri-stated. The four status buffers of IC_1 are selected and R_9 pulls the error status pin high. This pin is not used by the analogue and digital i/o sections. In addition, IC_5 is enabled.

Resistor $R_{SLCTINP}$ pulls the printer SLCTINP pin low, ie printer selected. Leds $D_{2,3}$ indicate whether printer pass-through or analogue and digital i/o is selected. Line b_2 of the control register, the INIT control line selects between upper and lower nibbles of the input byte from the a-to-d converter or digital input port.

Control lines STROBE and AUTOFEED are used to select one of the four i/o ICs – a-to-d, d-to-a, digital input or digital output – using the '139 decoder. When printer pass-through is selected, the '139's enable pin is pulled inactive high and all four i/o ICs are deselected. The d-to-a converter and digital output port retain their last value during printing. To prevent spurious intermediate values due to one of the AUTOFEED/STROBE pair changing state before the other, it is best if only one is changed at a time.

The a-to-d converter starts conversion when it is selected, ie the read line is pulled low. During conversion the a-to-d converter places data from the previous conversion on its data pins. An attempt to start a new conversion while the a-to-d converter is in the middle of a conversion results in an incorrect result.

If a switch is made directly from one output port to the other, the old port's value may briefly appear on the new port's outputs. To prevent this happening, an input port must be selected and the new port's output byte sent to the data register before the new port is selected.

Applying the design

The *LM358* input and output buffers have a slew rate of 0.5V/μs so limiting the analogue input and output frequencies to approximately 50kHz. Maximum frequency is given by,

$$f_{max} = \frac{\text{slew rate}}{2\pi \times \text{sinewave amplitude}}$$

where slew rate is in volts/s.

The *AD7569JN* a-to-d converter is configured for mode 2 operation. Bringing read low while chip-select is low selects the device and starts conversion.

Since the converter's busy line is not used a software delay is needed to determine when the conversion will be complete. Conversion time should be less than 2.6μs. ■

Control software example

This Qbasic program demonstrates how the analogue and digital i/o circuit can be controlled. It reads the a-to-d converter, prints the value on the screen then sends it to the d-to-a converter.

```
DEFINT A-Z
'string ch$
'int uppernibble, advalue, dloop
'int d is used in dummy INP loads
'int PDATA, PSTATUS, PCONTROL hold
' LPT1 reg addresses

DEF SEG = 0
        'FIND ADDRESSES OF lpt1
PDATA = (PEEK(&H409) * 256) + PEEK(&H408)
DEF SEG

PSTATUS = PDATA + 1
        'PRINTER PORT
PCONTROL = PDATA + 2
        'REGISTERS
PRINT HEX$(PDATA)

OUT PCONTROL, &H3
        'disIRQ 0, SLCTINP 0,lower nibble,
        'sel dig i/p port. 0000 0011
        ' ie printer deselected
d = INP(PDATA)
        'wait for ctrl lines to settle

DO OUT PCONTROL, &H1
        'dis IRQ,0,lower nibble,
        'sel ADC(&startconv),0000 0001
        'wait for conv to finish
FOR dloop = 1 TO 10
        'do nothing
NEXT

advalue = INP(PSTATUS) AND &HF0
        ' get lower nibble of ADC byte
advalue = advalue / 16
        'shift lower nibble to correct posn
PRINT "ADC lower nibble is  ", advalue

OUT PCONTROL, &H5
        'dis IRQ,sel iface brd,upper nibble,
        'sel ADC,0000 0101
d = INP(PDATA)
        'wait for ctrl lines to settle

uppernibble = INP(PSTATUS) AND &HF0
        'get ADC uppernible&setlower4bitstozero
PRINT "ADC upper nibble is  ", uppernibble

advalue = advalue OR uppernibble
        'combine up & low nibs to give byte
advalue = advalue XOR &H88
        ' correct BUSY inversions
        'ie invert bits 7 and 3
PRINT "ADC input byte is", advalue

OUT PDATA, advalue
        'set up dac value
d = INP(PDATA)
        'wait for lines to settle

OUT PCONTROL, &H4
        'dis IRQ,0,upper nibble,sel DAC,0000
0100
REM latch into DAC

PRINT "Press Q key to quit. Any other to
continue"
ch$ = INPUT$(1)
LOOP UNTIL ch$ = ("Q") OR ch$ = ("q")

        'SEL PRINTER PASS THROUGH
        '(if pass-through implemented) :-
OUT PCONTROL, &HC
        'dis IRQ,SLCTINP HIGH,0,00, 0000 1000
        'sets SLCTINPUT line low, ie printer
selected

'ENDPROGRAM
```

CIRCUIT IDEAS

Do you have an original circuit idea for publication? We are giving £100 cash for the month's top design. Additional authors will receive £25 cash for each circuit idea published. We are looking for ingenuity in the use of modern components.

£100 WINNER

Static noise limiter

Noise limiters usually rely on either low-pass filtering or companding methods. Both of these methods have disadvantages in either cost or performance.

This alternative uses a peak detector. The signal diagrams on the right show an ideal principle, in which the noise superimposed on the sine wave is averaged to leave just the signal – shown by the heavy line.

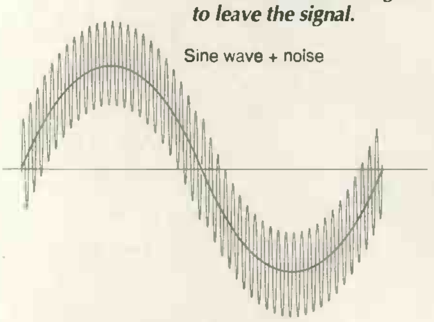
Since I have not found a way to do this, the alternative adopted is to detect the noise-plus-signal peaks to produce an approximation of the original signal, bottom right.

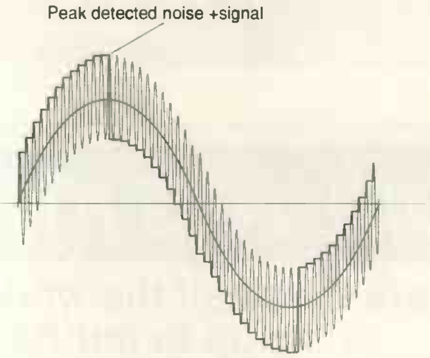
In the circuit shown, the input op-amp buffers the signal, which is applied to the bases of the peak-detecting transistors through C_1. Bias is provided by VR_1 and R_2. On a positive signal, C_2 charges via R_3 and C_2 holds positive peaks; when signal goes negative, negative-going peaks are held on C_2, which now charges through R_4. An output buffer completes the circuit.

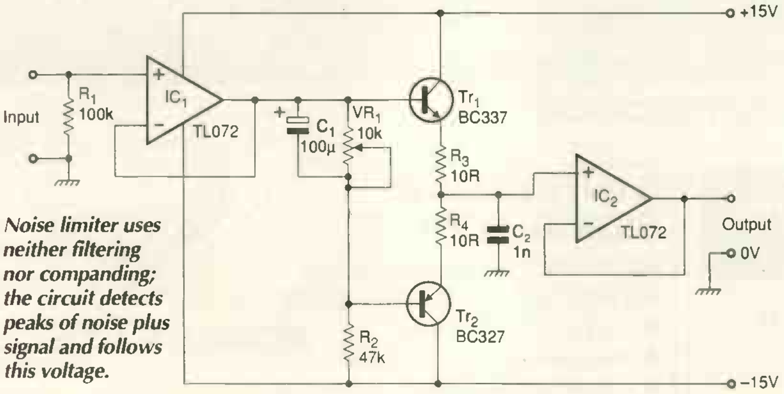
The amount of limiting is set by VR_1 and the dead zone is adjustable by VR_2 over a wide range. Some distortion is caused by the transistor switching, but in many circuits is less troublesome than noise.

J Macaulay
Chichester, Sussex.

Noise superimposed on the sine wave is averaged to leave the signal.







Noise limiter uses neither filtering nor companding; the circuit detects peaks of noise plus signal and follows this voltage.

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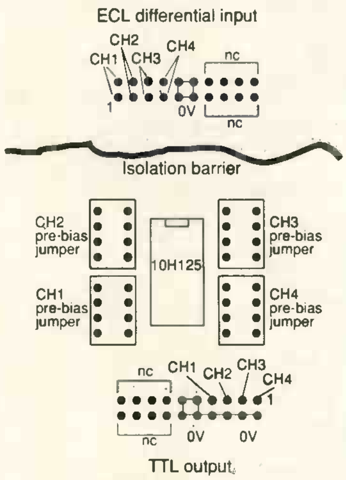
Fast-logic isolator avoids ground problems

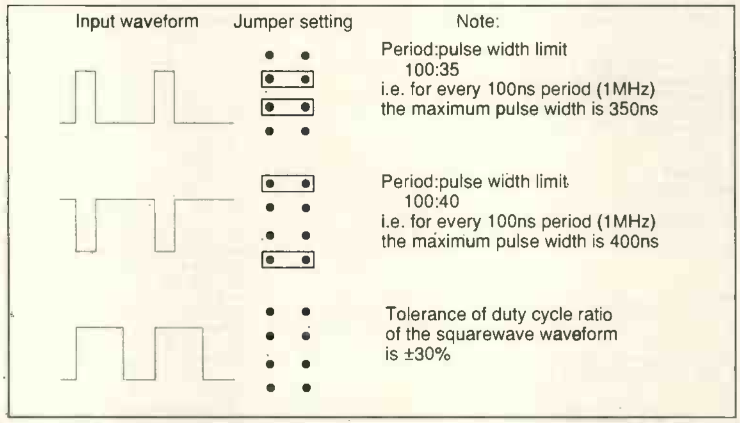
Two circuits, working at clock frequencies up to 100MHz, isolate the ground currents between two instruments to prevent the transmission of noise currents. Both circuits use twisted-pair lines. Ecl-ttl and ecl-ecl logic types are both catered for.

Opto-isolators do, of course, perform this function, but not at this frequency; their bandwidth is limited to around 40kHz. Therefore, a transformer is needed, the type used here having low capacitive inter-winding coupling with good inductive coupling at the frequencies of interest. Since the transformer cannot couple signals of less than 100kHz, with pulse widths of more than a few microseconds, a pre-biasing circuit is needed to define the output during steady-state periods by means of jumper settings.

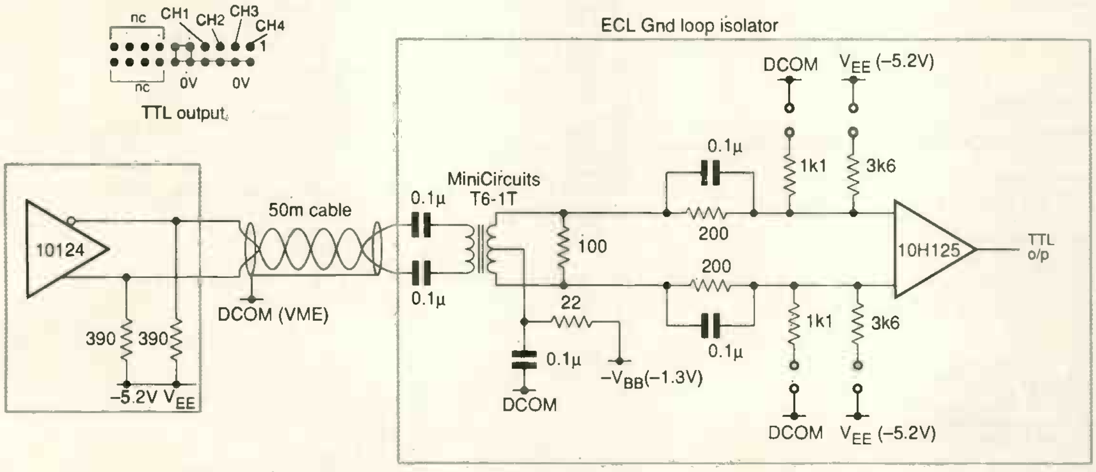
10KH ecl devices, rather than comparators, provide either ecl or ttl output because both cost and space occupied is less.

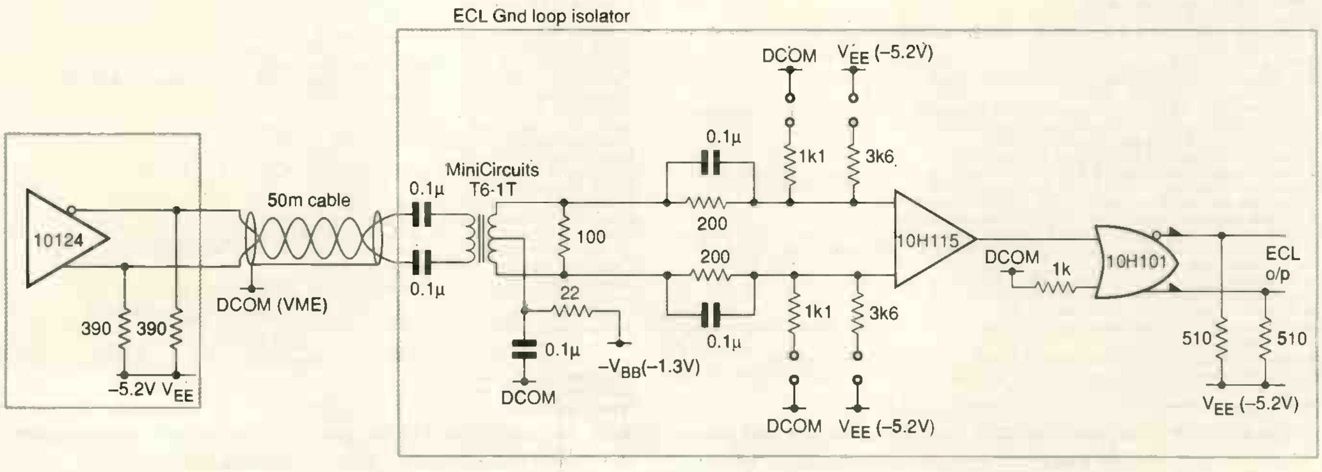
Richard Payne
Wimbledon
London





This jumper function diagram applies to both the ttl and ecl-output isolators.





Circuits to isolate or break the path between instruments and host computer to avoid ground currents interfering with analogue equipment.

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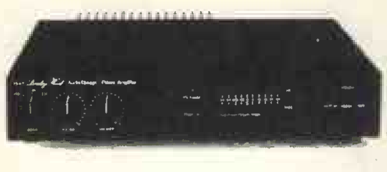

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Video transmission on vhf

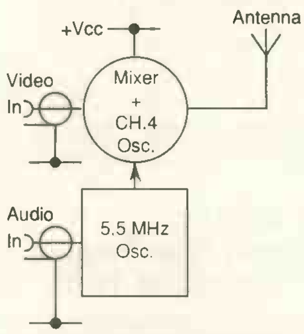
Only two transistors are needed to transmit audio and video to a tv capable of receiving channel 4 on vhf.

Transistor Tr_1 is a channel 4 oscillator and mixer. The second transistor is simply a 5.5MHz oscillator.

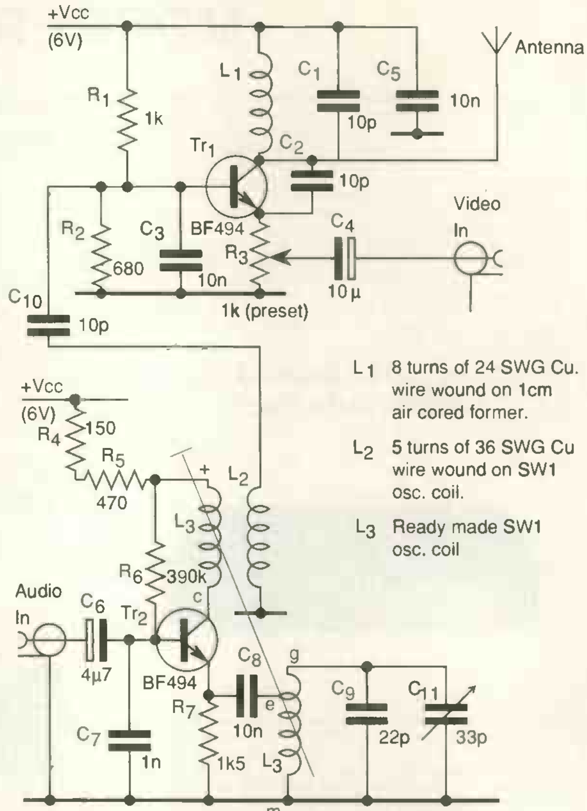
The coils are critical. Inductor L_1 is eight turns of 24SWG copper wound on a 1cm air-cored former. Coil L_3 is a ready-made and readily available SW1 oscillator coil. Five turns of 36SWG, wound around L_3, form L_2. Capacitor and resistor leads should be kept as short as possible.

To calibrate the transmitter, first set L_1 by stretching or compressing its coils until it is tuned to channel 4. The screen will blank when tuning is correct. Next set the 5.5MHz oscillator using either a frequency meter or a short-wave radio. Using a short-wave radio tuned to 5.5MHz, rotate variable-capacitor C_{11} and, if necessary the core of L_3, gently until you hear a hiss on the radio. The circuit is now ready to operate. Finally, with video and audio sources connected and with a tv tuned to channel 4, adjust R_3 for best results.

Raj K Gorkhali
Kathmandu
Nepal



Only two transistors are needed to transmit both audio and video on vhf channel 4.



Using a short-wave coil for the 5.5MHz oscillator greatly simplifies the transmitter's design.

Capacitorless solid-state relay

In this solid-state relay, there are no capacitors, hence no time dependencies. This, and the fact that the circuit can be made small, make the design suitable for implementing as a monolithic or hybrid IC.

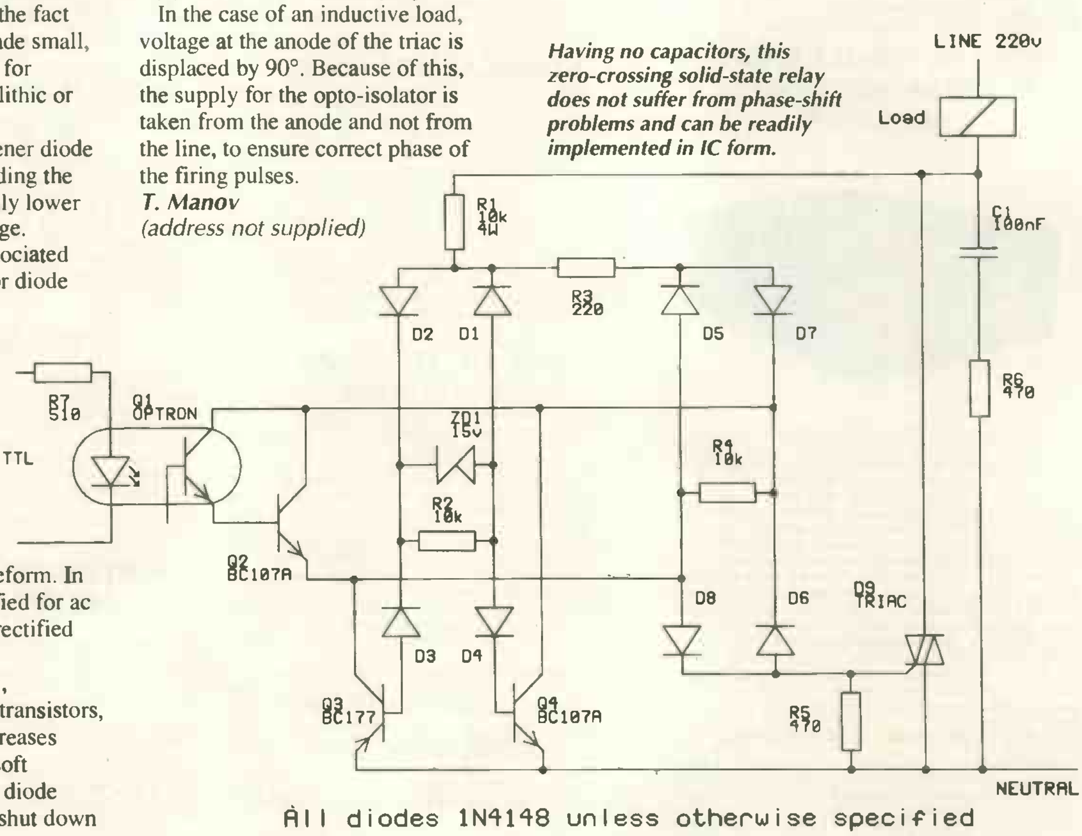
Bridge $D_{1\text{-}4}$ with the zener diode clamp the voltage, providing the opto-isolator with a supply lower than its breakdown voltage.

The isolator and its associated transistor act as a load for diode bridge $D_{5\text{-}8}$. They change the equivalent ac resistance across the input of the bridge. Because of this, the firing pulses always have the same polarity as the anode voltage and they appear every half wave to ensure a symmetrical output waveform. In this way, any load specified for ac will not be damaged by rectified mains voltage.

Zero-crossing circuitry, comprising two discrete transistors, reduces interference, increases reliability and provides soft starting. When the zener diode turns off, the transistors shut down the corresponding part of bridge $D_{5\text{-}8}$. Current is limited by R_3.

In the case of an inductive load, voltage at the anode of the triac is displaced by 90°. Because of this, the supply for the opto-isolator is taken from the anode and not from the line, to ensure correct phase of the firing pulses.

T. Manov
(address not supplied)



Having no capacitors, this zero-crossing solid-state relay does not suffer from phase-shift problems and can be readily implemented in IC form.

Prom and logic array in harmony
Due to an editorial error the last line in this piece should have read " of the prom on the address line A14". Our apologies if this was misleading.

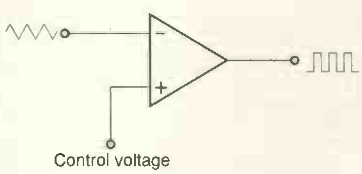
Simpler, but linear, pwm generator

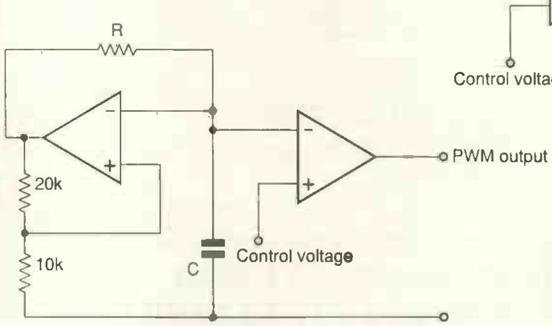
Since the width of an exponential rise and fall varies linearly with voltage, it can be used to replace a linear sawtooth in some applications. Here, it replaces the sawtooth in a pulse-width modulator, shown in the smaller diagram, where the sawtooth is taken to one input of a comparator and a direct voltage to the other. Mark:space ratio of the resulting constant-frequency square wave varies linearly with the direct voltage level.

In the second diagram, the exponential waveform across the capacitor of a standard op-amp relaxation oscillator is used for the same purpose; linearity with control voltage over the full range of zero to unity is within 0.5%. In this way, the circuit is cheaper than the sawtooth variety, since an accurate sawtooth generator in its simplest form consists of an integrator and comparator in a feedback loop.

At frequencies of less than a few kilohertz, an *LF353* dual op-amp is adequate, giving an economical design. For higher frequencies, a high slew-rate op-amp such as the *LM531* is needed and an *LM311* fast comparator will provide steep edges on the output.

D K Hamilton
Department of Engineering Science
University of Oxford





Replacing an integrator and comparator sawtooth generator with a simple relaxation oscillator still provides a linear relationship between control and output square-wave width in this cheaper pulse-width modulator.

Programmed bandpass filter

Clock frequency applied to a switched-capacitor filter determines the –3dB frequency, which means that such filters are programmable. It also means that, if two filters have slightly differing clock frequencies, a bandpass filter can be made.

If the output of one filter is made to be 180° out of phase with the other, by inverting in an op-amp, and the two outputs added, circuit output is low at low frequencies because the two outputs cancel. At high frequencies, output is again low because the filters have a low-pass configuration; and in the band between the two corner frequencies output is high since they do not cancel, one being too low in amplitude.

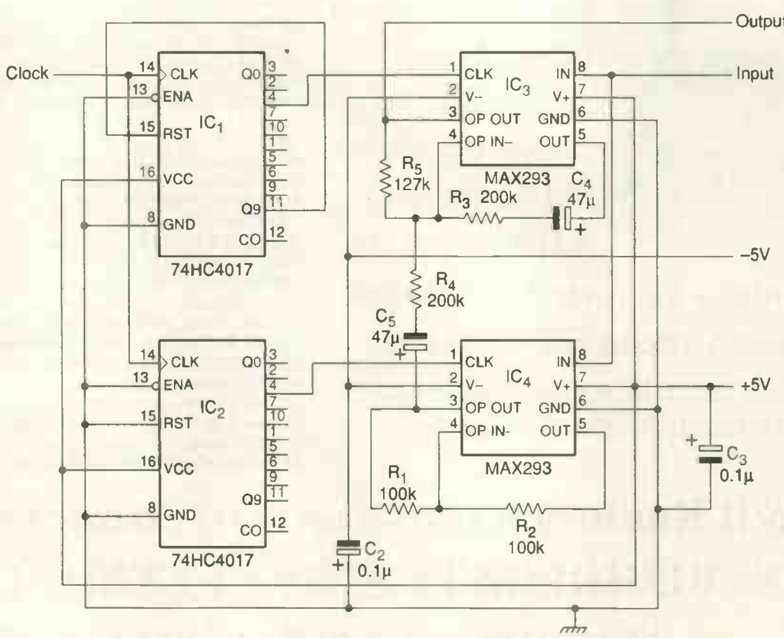
Here, the *MAX292* eighth-order, low-pass Bessel filter, which contains an uncommitted op-amp, exhibits a corner frequency of $f_{clk}/100$, two *4017* counters IC_1 and IC_2 dividing the clock input by 9 and 10 respectively.

Input goes to both filters, the output of IC_4 being inverted by the internal op-amp and added to the output of IC_3 by the device's op-amp, the resulting output having the characteristics described above.

The table shows what happens; the curve is steeper above the centre point than below, due to the behaviour of the eighth-order filters. Frequency of the centre-point depends solely on the clock frequency. ■

Yongping Xia
Torrance
California
USA

Peak frequency of this switched-capacitor bandpass filter is determined only by the clock frequency.



Frequency (Hz)	gain
20	0.06
30	0.10
40	0.15
50	0.19
60	0.27
70	0.33
80	0.42
90	0.58
100	0.81
103	1.00
110	0.74
120	0.22
130	0.03

Clock = 100kHz, input = 0.85V

Power modules *for quasi-resonance*

Relative to both monolithic and discrete technologies, power multi-chip modules provide an excellent compromise for quasi-resonant switch-mode power conversion. Paul Greenland, M. Ueki and M. J. Lee explain.

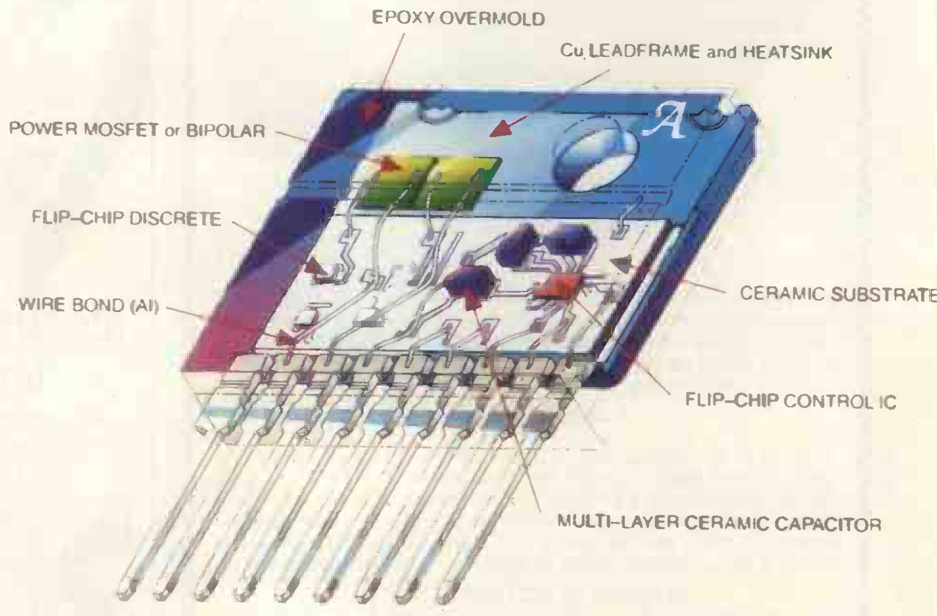
This article describes a new power multi-chip module – pmcm – which has evolved from consumer electronics applications and has considerable potential in general switch-mode power-supply designs.

Design of the module involves a complete systems discipline which has until now not been addressed by power device manufacturers. A single-ended, quasi-resonant flyback approach is described, along with an application-specific pmcm solution.

Successful power multi-chip module design is a 'holistic' procedure involving power-circuit, thermal and mechanical design. Until recently, non-monolithic approaches to power-circuit integration have been relatively unsuccessful. This is because traditional 'hard-switched' pulse-width modulated, pwm, power stages do not lend themselves to integrated packaging. The main reason for this is the high-frequency harmonic content of the characteristic current and voltage waveforms found in pwm circuits.

One principal reason that monolithic power integrated-circuit design is so difficult is that parasitic effects are omni-present, and must be considered at all stages of the design process. Fortunately, several conventional power topologies can be modified to reduce harmonic content without compromise on efficiency and component count. Furthermore, in most cases, savings elsewhere in the system

Fig. 1. Within the power multi-chip module packaging, a common copper heat sink carries the power switch and an alumina substrate with the control circuitry.



* Paul Greenland, Allegro MicroSystems, Inc.
M.Ueki and M.J.Lee, Sanken Electric Co.

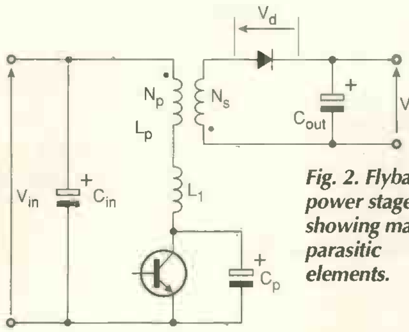


Fig. 2. Flyback power stage showing major parasitic elements.

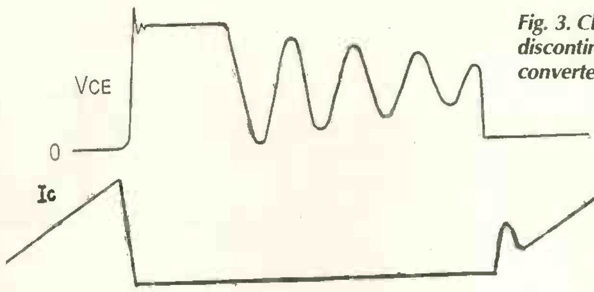


Fig. 3. Characteristic discontinuous flyback converter waveforms.

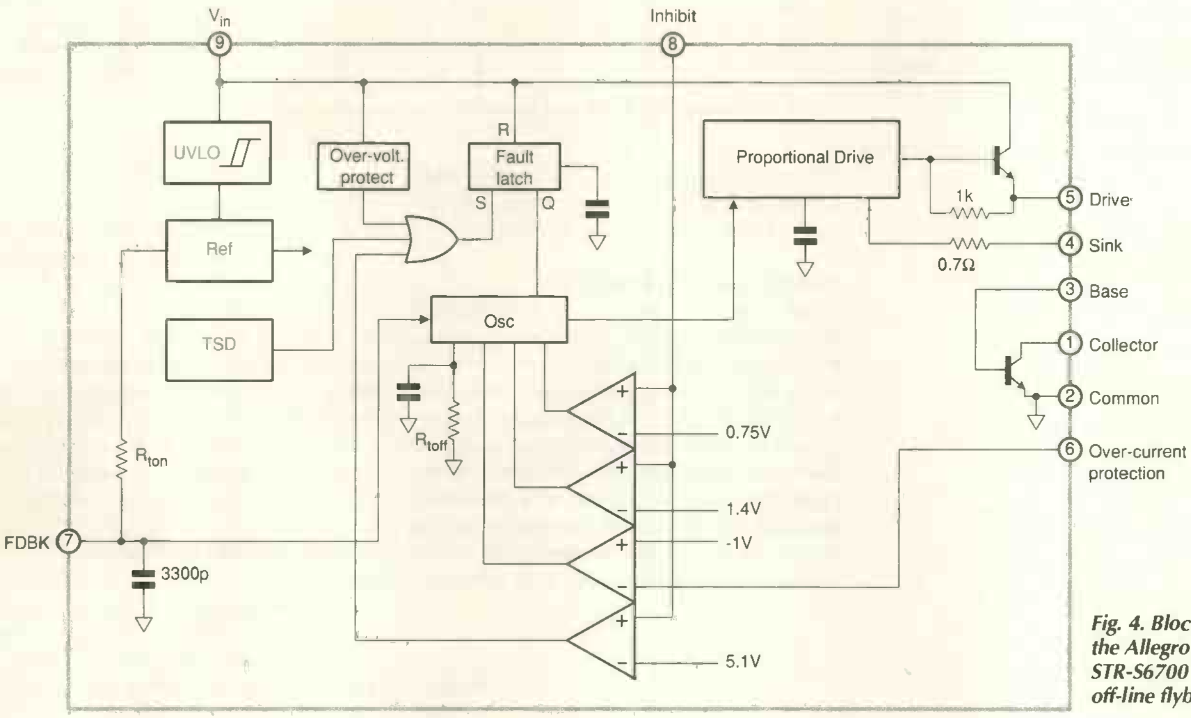


Fig. 4. Block schematic of the Allegro Microsystems STR-S6700 quasi-resonant off-line flyback convertor.

may be significant – particularly when power-supply ripple rejection of the load system is questionable.

Power module packaging

Originally developed for consumer electronics applications and used in areas such as tv sets, the *3GR* series package incorporates the control and power switching elements of a single-ended quasi-resonant flyback convertor. The package, **Fig. 1**, includes an internal plated copper heatsink carrying the switching power device – a bipolar transistor, mosfet or igbt.

The power device exhibits a thermal resistance to the mounting surface equivalent to what it would show in an over-moulded TO-247 package – typically no greater than 2°C/W. The copper heatsink also carries the control circuit on an alumina substrate with laser-trimmed thick-film resistors, discrete small-signal and driver devices, and local decoupling and timing capacitors. All active devices on the substrate – including the monolithic control IC – use 'flip-chip' interconnections.

Flip-chip interconnection reduces bond-wire count, and allows the monolithic control IC designer to optimise layout. Bond-wire parasitics are also eliminated. A drawback of flip-chip interconnect is that the active surface of the control IC is capacitively coupled to the collector or drain on the back side of the power device, which is subject to high dV/dt in conventional single-ended pwm topologies.

This effect can be reduced significantly by including an rf 'catcher' plate in the substrate. This will lower control-node impedances and includes blanking or switched attenuator circuits that reduce node sensitivity during the commutation interval.

These solutions, however, raise costs and reduce reliability. A better way is to introduce quasi-resonant techniques which make the switching transitions resonant, cut high-frequency harmonic content and reduce switching losses – applicable if fixed-frequency operation can be sacrificed. Moreover, these techniques make positive use of parasitic elements which can never be totally eliminated.

Parasitic elements

Figure 2 shows a flyback power stage with its major parasitic elements – leakage inductance L_l and primary capacitance C_p. In this circuit, L_m is the primary magnetising inductance, N_p and N_s are the numbers of primary and secondary turns, and V_d is the forward drop of the secondary rectifier.

The leakage inductance is a result of imperfect coupling between the primary and secondary windings. It may be as high as 5% of the value of L_m in a typical off-line power supply with a primary/secondary isolation of 3750V rms and a 4mm winding margin. The main effect of leakage inductance is to delay transfer of energy stored in the primary inductance during the 'on' time to the secondary during the 'off' time. During this delay, there is a voltage overshoot at either the collector or drain of the power switch, and the energy resonates between L_l and C_p with a frequency f_l given by the equation,

$$f_l = \frac{1}{2\pi\sqrt{L_l C_p}}$$

After the energy stored in the leakage inductance is exceeded, the voltage on the switch settles to the referred value V_{fb}, given by the equation,

$$V_{fb} = V_{in} + \left[(V_o + V_d)\frac{N_p}{N_s} \right]$$

The peak overshoot voltage is given by the sum of V_{fb} and V_{ring},

$$V_{pk} = V_{fb} + V_{ring}$$
$$= V_{fb} + \hat{I}\sqrt{\frac{L_l}{C_p}}$$

Once the energy has been transferred to the secondary and the secondary rectifier ceases to conduct, residual energy resonates between L_m and C_p with a frequency f_2, settling to V_{in} before the next conduction time,

$$f_2 = \frac{1}{2\pi\sqrt{L_m C_p}}$$

The primary capacitance is the sum of the capacitance of the power switch, the transformer's intra-winding capacitance and any stray capacitance. It is discharged at the beginning of power-switch conduction, and dissipates energy at turn-on.

Each parasitic element poses a problem in conventional pwm flyback converters – leakage inductance causes overvoltage, and primary capacitance causes overcurrent.

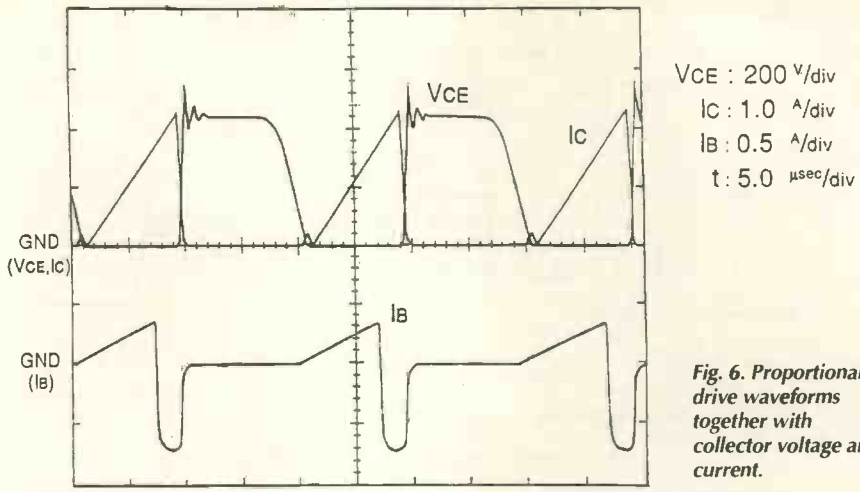
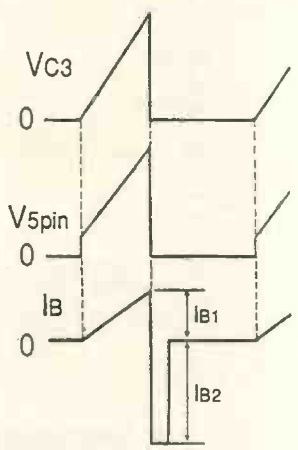


Fig. 6. Proportional drive waveforms together with collector voltage and current.

Making use of parasitics

Quasi-resonance makes use of these undesirable parasitic elements. The primary voltage and current waveforms are shown in **Fig. 3**. The point of lowest potential on C_p occurs one half-cycle after the core has unloaded its energy and the secondary rectifier has ceased to conduct. If conduction time commences at this point, turn-on loss will be at a minimum. Similarly, as turn-off loss occurs during the overlap of voltage and current at the end of conduction time, reducing the dV/dt with a primary capacitor at turn-off is beneficial.

In order to achieve these ends, independent control of 'on' and 'off' time is essential. In the pmcm, **Fig. 4**, 'on' time is controlled to a preset maximum by the voltage feedback loop. 'Off' time is controlled to a preset maximum by detecting the point at which the core has unloaded energy. This is achieved by monitoring the auxiliary winding referred to the primary, and then applying a half-cycle delay. Using this technique, all benefits of quasi-resonance can be realised in a manner which is predictable in high-volume manufacture.



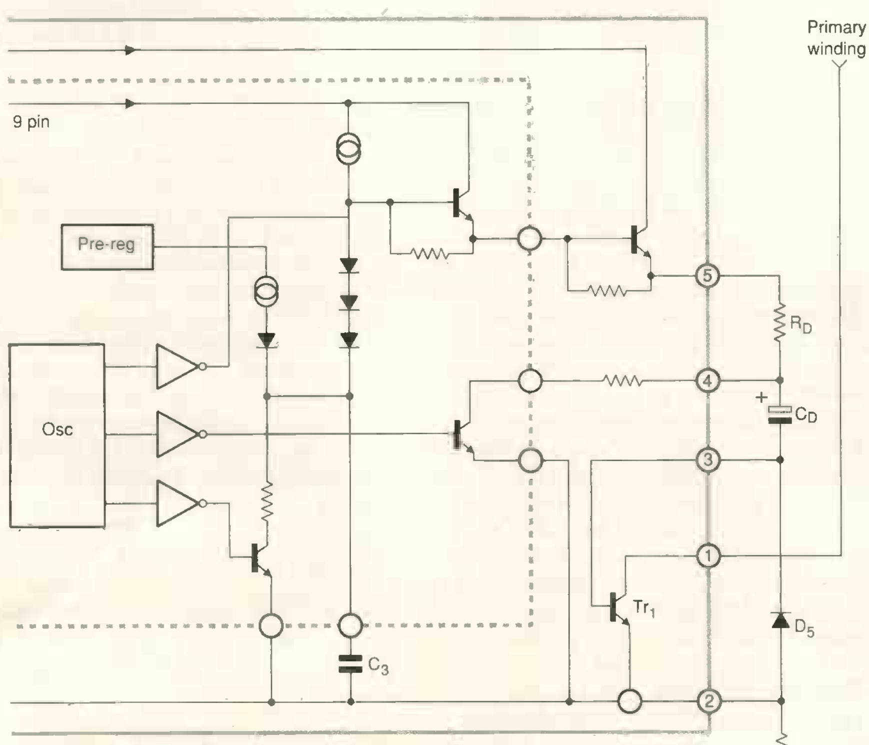


Fig. 5. Proportional drive circuit with characteristic waveforms. Using proportional drive helps the power supply cater for a wide load range. In the upper waveform diagram, $I_{B1(peak)}=(V_{9pin}-3)/R_D$

Power module realisation

Figure 4 is the block diagram for the *STR-S6700* power modules. These use the physical construction of Fig. 1. In this device the control circuit is subject to an undervoltage lockout, with hysteresis, which allows low-power startup from energy stored in the auxiliary capacitor fed by a high-value start-up resistor. Start-up current is not greater than 200μA.

Comprehensive protection is ensured by temperature-compensated cycle-by-cycle current limiting, latching overvoltage shutdown and thermal shutdown – which acts to protect the controller and the power switch. Latching functions are subject to a predetermined delay to prevent nuisance tripping. The oscillator has predetermined maximum on and off times.

Inclusion of a triple-diffused npn switching transistor rather than a mosfet may seem a retrograde step. This is not the case however, as the use of a proportional drive technique ensures the device is well matched to the application, and a $V_{ce(sat)}$ of 400mV gives low conduction loss.

Casual inspection of the primary current waveform reveals that it starts from zero – neglecting the leading-edge spike caused by discharging C_p. Fast switching at this point is undesirable, as this leads to excessive auxiliary current and high electro-magnetic interference. Furthermore, it is useful to have some proportionality between base and collector current for the leading portion of the primary conduction time to cater for a wide load range. Turn-off should be rapid to minimise switching loss, and a reverse bias should be applied to the base during turn-off voltage overshoot for reliable operation.

Figure 5 shows the proportional drive circuit and its associated waveforms. At switch-on, capacitor C_3 charges up, ramping the base current to a maximum set by R_D. At the end of the primary conduction time, current ceases to flow through pin 5, and the transistor drawing current from pin 4 switches on. This reverses polarity on the drive capacitor C_d, which has charged to a level set by D_5 in anti-parallel with the base/emitter junction of Tr_1.

Figure 6 shows the proportional drive waveforms, together with the collector voltage and current. Switching characteristics and low conduction loss of the bipolar switching device are matched to its application.

One of the main reasons that bipolar transistors have gained a poor record for reliability in switch-mode power-supply applications is the drive technique for the ringing-choke convertor used in early pcs. This can be seen from the drive waveforms in **Fig. 7**, which exhibit a large current at switch-on, creating a large switch-on current pulse as C_p is discharged.

During primary conduction, there is no proportional element, resulting in overdrive at light loads and excessive storage time. The turn-off is slow, resulting in large switching loss, and the reverse bias during the voltage overshoot on the collector is small. It is therefore hardly surprising that the major cause of failure in many ringing-choke converters is the bipolar power transistor.

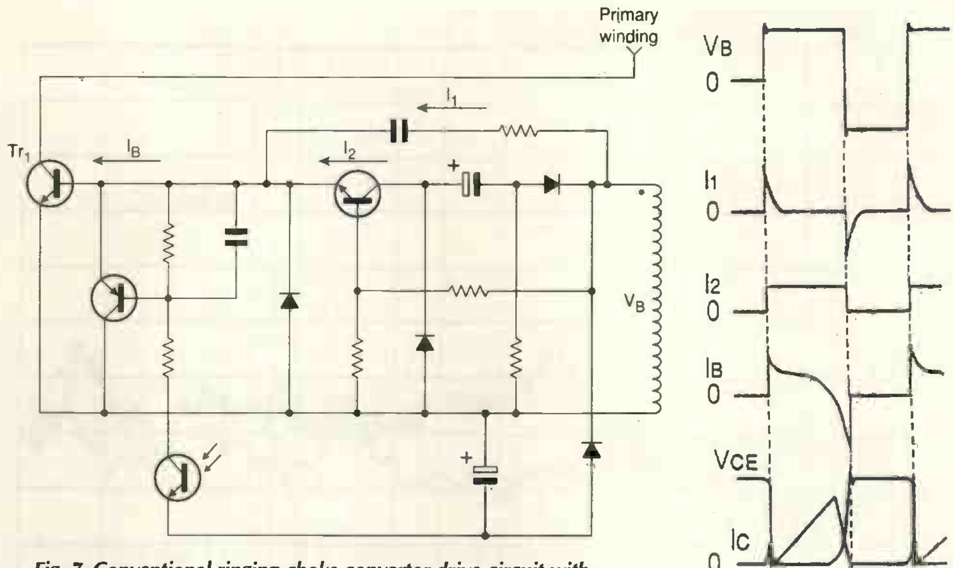


Fig. 7. Conventional ringing-choke convertor drive circuit with waveforms.

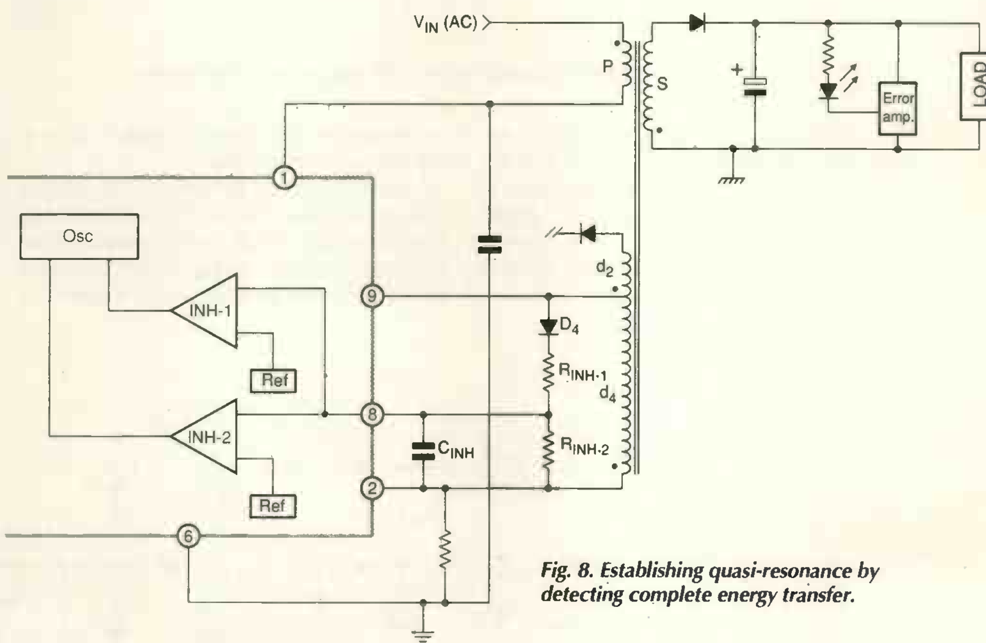


Fig. 8. Establishing quasi-resonance by detecting complete energy transfer.

Making energy transfer complete

Detection of demagnetisation, known as 'complete energy transfer', is a crucial element in establishing quasi-resonance.

In **Fig. 8**, the auxiliary winding has a sampling network connected ahead of the auxiliary rectifier. The sampling network applies a voltage to the inhibit pin, pin 8, during the auxiliary/secondary conduction time. Once this voltage exceeds V_{TH1} the drive to the power switch is suspended, and as it exceeds V_{TH2} the 'off' time is pre-terminated.

Once the core has unloaded its energy, the voltage at the inhibit pin, V_{INH}, drops, delayed by C_{INH}. As V_{INH} falls below V_{TH2}, the 'off' time capacitor and the proportional drive capacitor are re-initialised. Once V_{INH} falls below V_{TH1}, the drive to the power switch is enabled and primary conduction commences. **Figure 9** shows the complete timing sequence.

This technique allows quasi-resonance to be established, with the consequent benefits of lower switching loss, lower electromagnetic interference than with conventional ringing-choke converters **Fig. 10**, and utilisation of hitherto undesirable parasitics.

The negative aspects of quasi-resonance are a slight increase in conduction loss, a higher fundamental frequency component in the conducted harmonics, and a small increase in high-frequency ac loss in the transformer. The latter can be virtually eliminated by use of high-frequency transformer winding techniques – Litz, multifilar bundling and interleaving – to reduce ac losses.

Primary magnetising inductance is modified from the conventional value calculated in the classic ringing-choke converter transformer design procedure, as shown by the equation,

$$L_p = \frac{(V_{in}D)^2}{\left[\sqrt{\dfrac{2P_{out}f_o}{\eta}} + V_{in}\pi f_o D\sqrt{C_p}\right]^2}$$

where L_p is primary inductance adjusted for quasi-resonance, P_{out} is the output and auxiliary power in watts, f_o is the minimum switching frequency, η is efficiency, D is the duty cycle at minimum ac line potential and V_{in} is the minimum dc input voltage.

Isolated voltage feedback and regulation are achieved by augmenting the current charging C_{Ton} with opto-coupler current proportional to the error signal generated at the secondary.

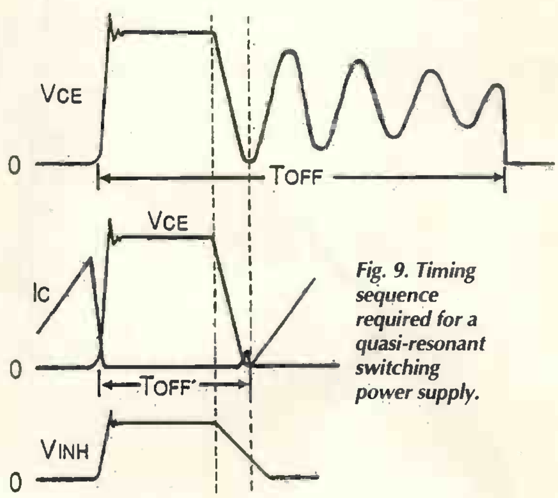


Fig. 9. Timing sequence required for a quasi-resonant switching power supply.

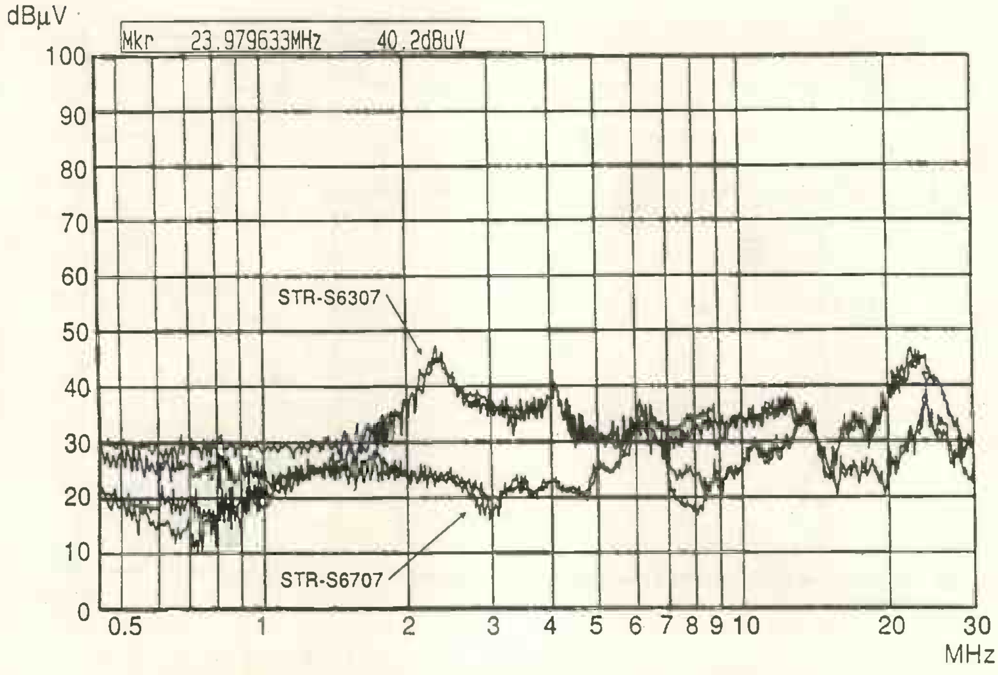


Fig. 10. Comparison of conducted EMI – ringing choke versus quasi-resonant techniques.

Thus, as the secondary load decreases or the applied line voltage increases, the slope of the voltage ramp on C_{Ton} increases, reducing primary conduction time. In most cases, simple integrator compensation around the secondary error amplifier will suffice. However, if more control-loop agility is required, pole/zero compensation and primary high-frequency bypass may be employed.

A useful feature of the *STR-S6700* series is standby operation. Originally intended for tv applications, the standby feature minimises the incremental power drawn by the convertor during light load operation. This feature is a by-product of the oscillator design and the quasi-resonant timing technique employed – refer to **Fig. 11**.

In standby mode, current normally drawn through R_{11} and D_9 is diverted into the base of Tr_3, turning it on. As current flows through R_{12} and D_{10} and Tr_2 is switched on, output and auxiliary voltages fall, and stabilise when S_1' reaches V_s, where $V_s=V_{R11}+V_{D10}+V_{BEQ3}$. At this time, standby power is supplied by C_{11} charging from the collector of Tr_2. As voltage on the auxiliary windings d_1 and d_2 tracks the secondaries, auxiliary power is supplied from the linear regulator transistor Tr_1 on d_2. In addition, as V_{INH} drops below V_{TH1} during the off time, the oscillator defaults to its maximum off time of typically 50µs.

In standby mode, the power supply operates with pulse ratio control, in which the 'on' time varies and the off time is held constant. This keeps the incremental power consumed by the power supply to an absolute minimum. This principle can be applied to save energy in printers or copiers which switch to standby mode if the system is inactive for a period.

A variant of the device without the standby mode is shown in **Fig. 12**. The *STR-S5700* Series is intended for indirect feedback applications using a primary referred sense winding, and component count is reduced at the expense of regulation.

Conclusion

This article has shown that integration of the primary power switch and control elements of

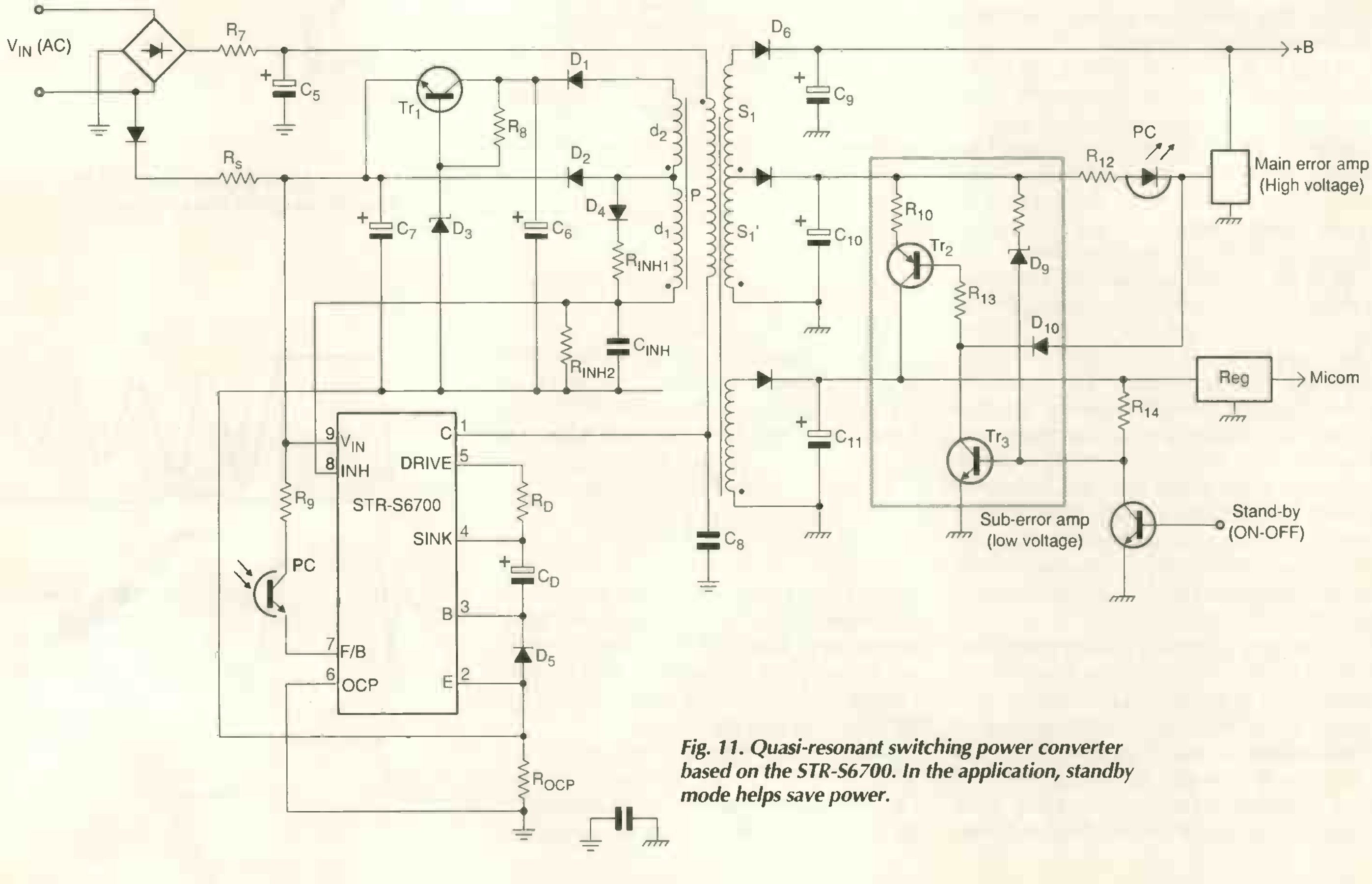


Fig. 11. Quasi-resonant switching power converter based on the STR-S6700. In the application, standby mode helps save power.

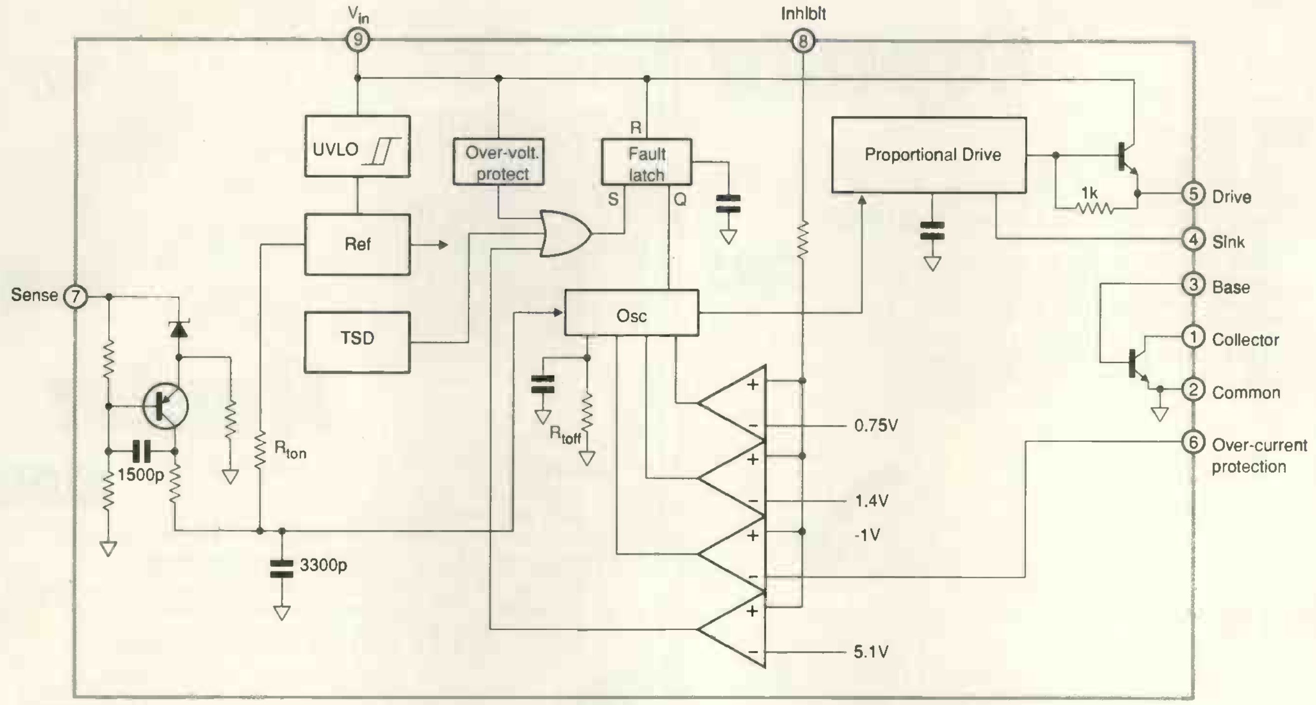
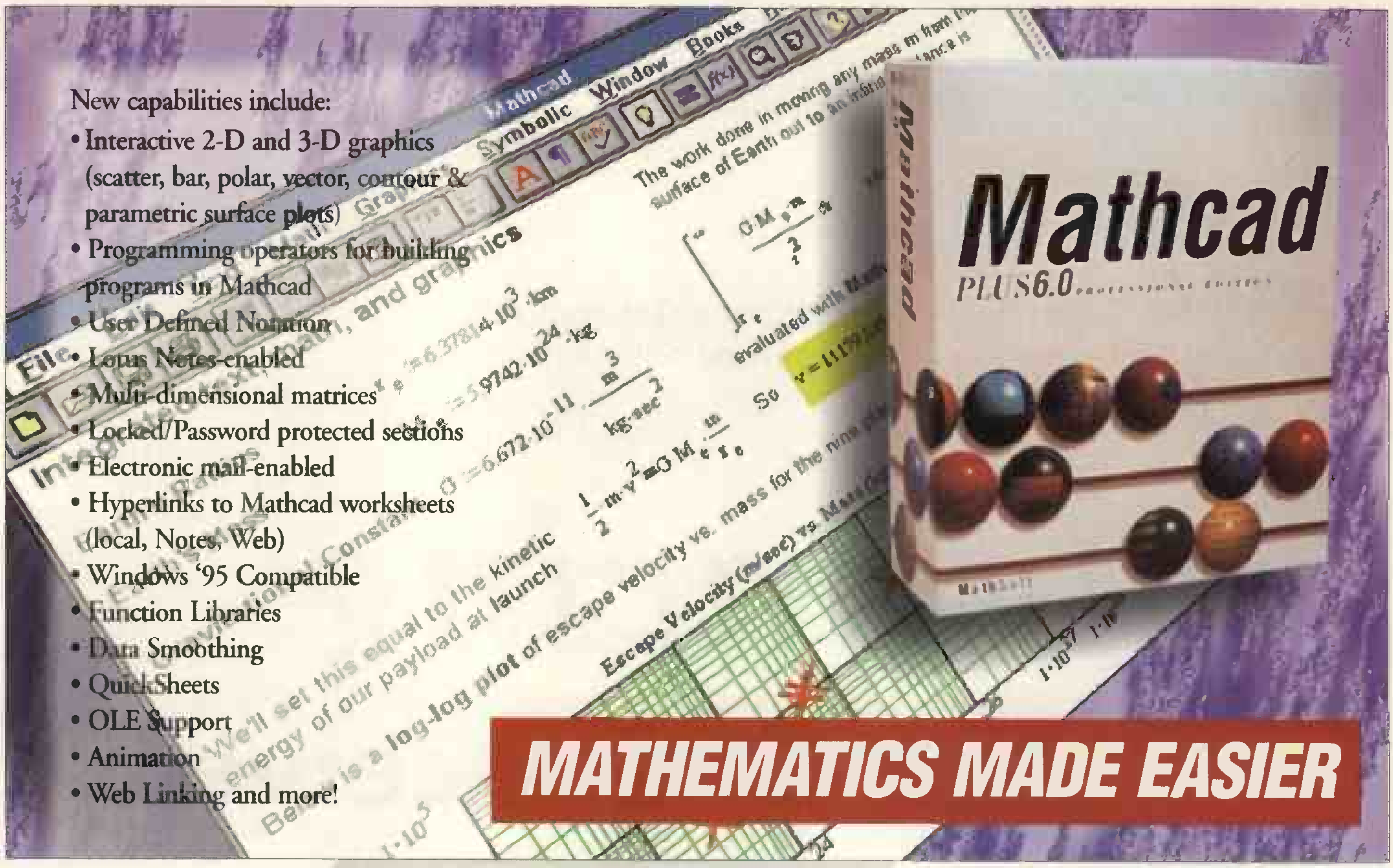


Fig. 12. Block schematic of Allegro Microsystems STR-S5700 converter.

an off-line switching power supply is viable once a true systems approach is adopted. Quasi-resonance can also be employed in off-line flyback converters to good effect. Furthermore, this technique brings the 'lowly' flyback convertor to the forefront of off-line power-conversion applications below 200W.

It can be shown that reduction in high-frequency harmonics, offsets many disadvantages of variable-frequency operation.

The quasi-resonant technique can also be adapted to current-mode control through a dual-purpose current limit which reduces pin count and is competitive with implementations using conventional controllers plus discrete components. ■



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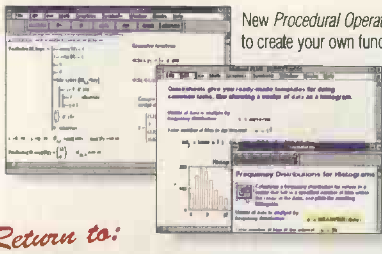

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CIRCLE NO. 138 ON REPLY CARD

Building on August's article describing a GPS receiver, Nigel Gardner presents a low-cost read-out interface for Rockwell's *MicroTracker* system.

Interfacing the GPS receiver

I chose the PIC microcontroller as the basis of this *MicroTracker* interface because of its ease of use and low development cost. It will handle all serial comms, sorting out and formatting of information, and display driving functions, needing a minimal number of external components to function.

The *MicroTracker LP* has two data output modes – binary and NMEA. This design is based on the NMEA format at 4800bit/s and interfaced to a *PIC16C74*. If data manipulation is required on the GPS information, then the binary format is the better option as there is a software overhead converting to and from the NMEA ASCII format. However, if the interface is only used to display and store data, the NMEA format is easier to work with.

This application is intended purely to display information, with no calculation done on information received.

Hardware design

The heart of the design is the *PIC16C74* microcontroller running at 4MHz, **Fig. 1**. An RS-232 link interfaces to the GPS serial interface module featured in the August 1995 issue and uses a *MAX232A* driver chip to level shift to the PIC.

Other items needed for the PIC are a *78L05* regulator, a 4MHz resonator and a few passive components for decoupling and reset. Power is drawn from the GPS serial interface board via a connection to the active antenna supply and is in the region of 15mA at 12V. This breaks down into 12mA for the *MAX232A*, 2mA for the lcd display, and 1mA for the *PIC16C74*. If the RS-232 interface is removed to give ttl connections between the MicroTracker and the PIC, a saving can be made on both current and components.

A reset switch is added to the PIC to assist in software development, but it could be eliminated. If the GPS serial interface module is used in conjunction with this design, then set the dip switches on the board to 1-5 on, 6-8 off. This sets NMEA mode at 4800bit/s on power up.

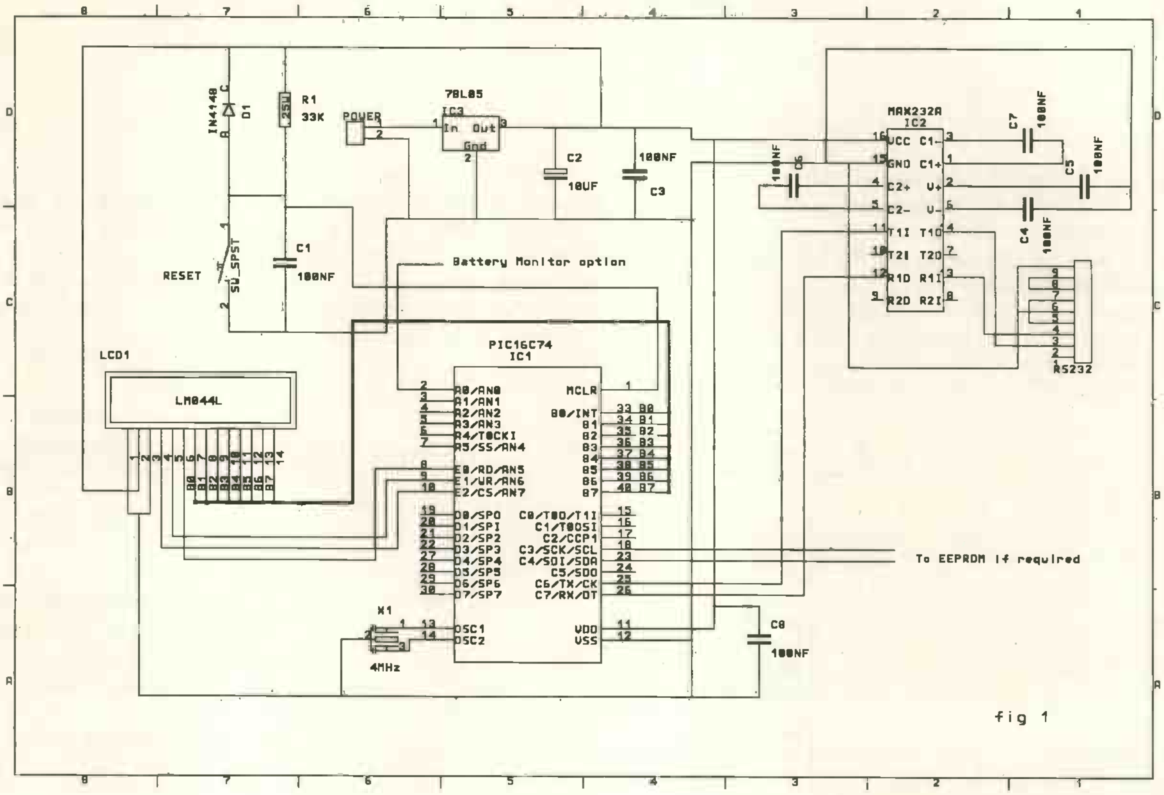
The lcd module is the Hitachi *LM041L* or *LM044L*. These have four lines of 16 or 20

Tips for enhancing the software

The current version of software does not look for a fix before starting to display information on the display. A modification could be made to examine data from the MicroTracker, looking for the valid signal character before passing it to the sign on message stage. Information received from the MicroTracker could be examined for a '.' symbol in addition to the ',' to truncate data displayed at the decimal thus reducing information presented to the end user.

Additional messages could be enabled and stored in alternative locations for extracting information not found in a specific message.

Fig. 1. Display controller for the MicroTracker is based on a PIC microcontroller and incorporates a liquid-crystal display module.



characters respectively and are interfaced in the 8-bit mode with full handshaking. This method of interfacing provides the fastest display update times but uses eleven i/o lines. If this design is transported to a 28-pin *16C73* then the 4bit write only approach can be taken with the display using only six i/o lines.

Software requirements

The flowchart for this program, Fig. 2, shows the basic operation. Source code is available from the TDC bulletin board for those wishing to evaluate the program and modify to their own requirements.

Following initalisation of the ports and other registers, the display is cleared and a sign on message sent. For this application, the default message formats of 'gga' and 'vtg' are turned off and 'rmc' turned on. The 'rmc' message format includes latitude, longitude, time, date, speed, heading, magnetic variation and magnetic heading. Other message formats can be enabled easily and relevant information extracted from the data string.

The 'rmc' message contains the following data – start field, 'utc' time, data valid, latitude, latitude direction, longitude, longitude direction, speed, heading, date, magnetic variation, magnetic direction, checksum <CR><LF>. Each field is comma delimited and will typically look like:

$GPRMC,234215.24,A,3339.686,N,11751.667,W,0.620,293.8,180595,14.0,E*79

Other message formats include information on altitude, number of satellites in use, track made good and ground speed. It can also be enabled and set to broadcast at increments of a

PIC object software for the GPS receiver display module.

```
:08000000E328820020342034C3
:0800080042346C347534653498
:08001000623469347234643477
:08001800203447345034533406
:080020002034203400344C347C
:0800280061347434203400340B
:080030004C346F346E342034AF
:0800380000346E342034DF3483
:08004000003454346934 6D34BE
:08004800653420340034243437
:08005000503452345734493496
:080058004C344F3447342C34C2
:0800600052344D3443342C34BA
:0800680041342C3431340D3415
:080070000A340034243450343A
:08007800523457344934 4C3472
:080080004F3447342C3447349F
:0800880047344134 2C34563496
:080090002C340D340A34003455
:08009800243450345234573473
:0800A00049344C344F3447345D
:0800A8002C3456345434473463
:0800B0002C3456342C340D34BD
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:0800F000530A000A00B7A28E6
:0800F800080089013830860080
:08010000B0209A2008008901DB
:080108000C308600B0209A20A3
:08011000080089010630860099
:08011800B0209A2008008901C3
:080120008600B0209A200800BF
:08012800890109158600B020D1
:080130009A2008008316FF303D
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:0801400009147820060809 10DB
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:0801680008008230 8F2008001E
:080170008030 8F200800C03030
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:0801900003A031900342508B0
:080198009420A40AC528A4006C
:0801A00024080120A500003A2B
:0801A8000319E02880309800E3
:0801B0002508990083169 81C34
:0801B800DB288312A40AD02801
:0801C0009030980000348 60124
:0801C8008901831602309F003B
:0801D0008030870089018601DF
:0801D8000C3099002030980062
:0801E0008312903098 00AB205F
:0801E8007220B8200230C4208F
:0801F0002730CF203A30CF2068
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:08020000FF281A08243A031D2F
:08020800FF28A301303084003F
:080210008C1E08291A080D3AA2
:08021800031912291A088000E5
:08022000840A08297220303025
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:080238001529B8202130C42073
:08024000840A00082C3A03199E
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:08025000BB201330C420840A16
:080258000000082C3A03191929D2
:08026000000894202B29BE20A8
:080268001830C420840A0008CC
:080270002C3A031919290008BA
:08027800942036 29C120840AFC
:08028000000082C3A03191929AA
:0802880000894203F291D30FD
:08029000C420840A00082C3A86
:0802980003191929000 8942044
:0202A0004929EA
:00000001FF
```

second. See the designer's guide on the *MicroTracker LP* for full specification and programming information.

Character input from the RS-232 port is examined to look for the 24 hour '$' symbol, signifying the start of the message string. On receipt of the '$' symbol, the characters are stored in the PIC's ram. On receipt of an end of message character – ie carriage return, $0D_{16}$ – the software branches into display mode. Information from the MicroTracker arrives at a minimum of 1s intervals providing enough time for the display to be updated.

The display section of the program looks at the characters in memory and when a comma, $2C_{16}$ is identified – field delimiter, the message type is then determined from a lookup table and sent to the liquid-crystal display. Additional text messages are sent prior to the received information to clarify to the user what is being displayed.

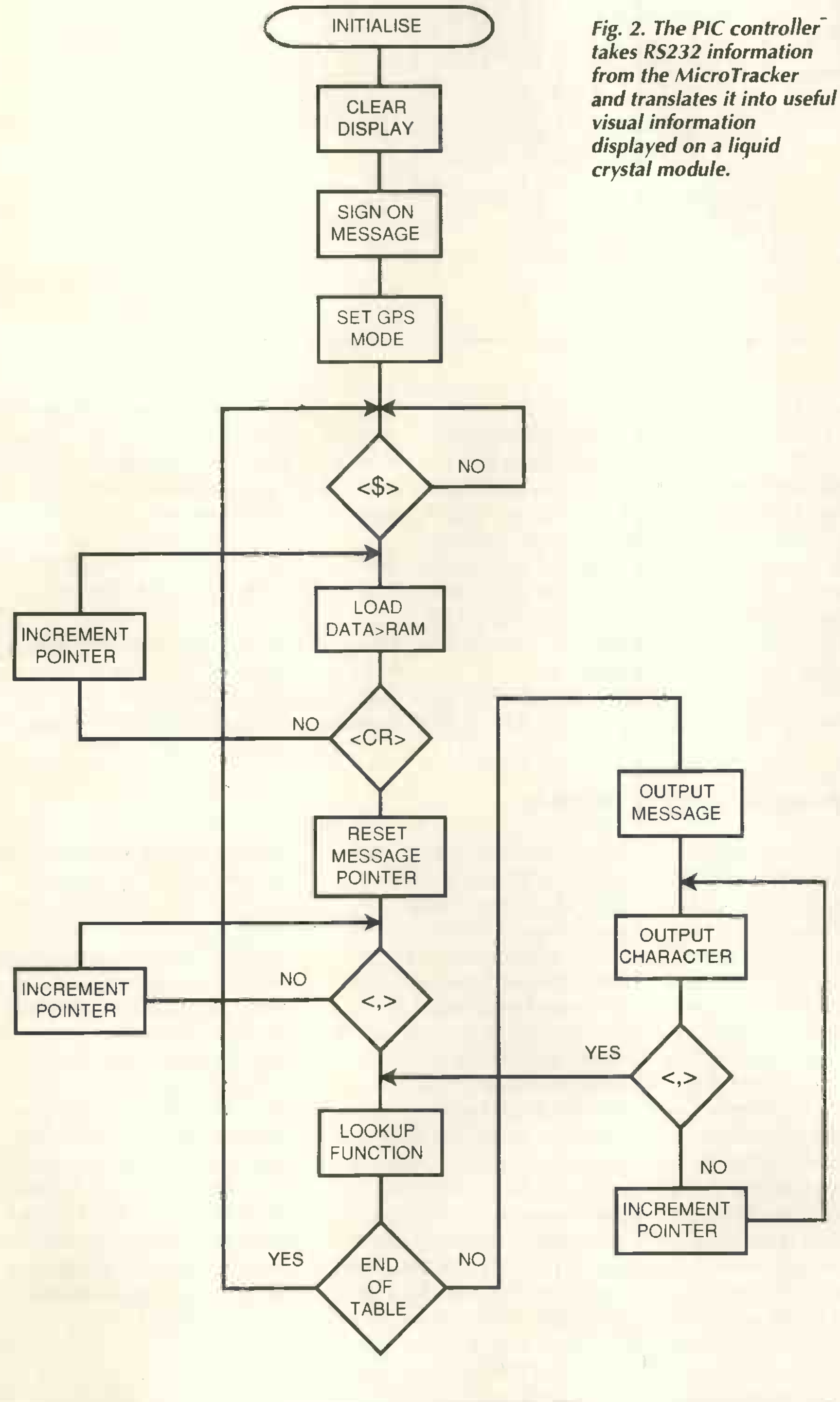


Fig. 2. The PIC controller takes RS232 information from the MicroTracker and translates it into useful visual information displayed on a liquid crystal module.

At the end of the information processing, the program jumps back to look for the next '$' symbol.

Bit rate calculation for the PIC serial comms port is straightforward. After transposing the formula in the data sheet, for low rates it is,

$$\text{Divide ratio} = \left\{\frac{\text{clock frequency}}{\text{baud} \times 64}\right\}^{-1}$$

This works out for 4800bit/s and a 4MHz clock at 12. This value is loaded into the SPBRG register and is common for both transmit and receive.

Designing a board

The prototype was built using a standard *PIC16C64/74* development board – Farnell order code 630-639 – with the display bolted on. This enables full access to all PIC pins and reduces development time and cost.

Laying out a board for the circuit in Fig. 1 should not present problems, but care should be taken in placing the resonator and 100nF capacitor as close to the PIC as possible.

Where next?

If an external electrically erasable prom is added to the PIC and connected via the inbuilt *I²C* interface, data could be logged in the eeprom upon closure of a switch providing a positional logging and ident system. Using an external eeprom, the *MicroTracker LP* can be updated on power up with the last positional information – speeding up the time to first fix.

As the *16C74* has eight analogue inputs, one could be used for monitoring battery voltage via a potential divider. This will alert the user to the time remaining before total loss of supply.

If cost is an issue, the *MicroTracker LP* can be connected directly to the PIC, bypassing the 'logic to RS-232 to logic' conversion circuitry and interface board. This brings component cost down to the *MicroTracker LP*, antenna, PIC, display and power supply. ■

Technical support

Rockwell MicroTracker LP Designer's Guide, contact Telecom Design Communications BBS. Phone 01256 332800 for connection details.
Beginners Guide to the Microchip PIC, Rev 2, ring 01628 777960 for details.
Microchip Databook 1994, Farnell/RS/Maplin.
The author runs Bluebird Electronics, 01380 725110 providing PIC training and support. This article is based on his forthcoming book 'PIC Cookbook – Vol 1'.

LETTERS

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

PhoneDay fiasco

At last, the dawn of realisation has broken over Oftel. As suspected, the new telephone number scheme introduced in April is what most of us who have to phone the USA have suspected – a great big waste of money.

I do not know how many telephones there are in the UK, but in the USA, with a population five times the size of ours, they have a simple and therefore easy to remember, three-digit area code, three-digit exchange code and four-digit number system.

There are far more phones per head of population there than here, and they even have the luxury of an exchange code for film and tv programme makers (555) to stop the nuts dialling the number on the tv.

Can any one explain why we couldn't have adopted such a simple scheme for our telephone numbering system? I don't want to appear too cynical by suggesting "not invented here".

Nic Houslip
Birmingham

Field hazards? Proof please

Your correspondent Roger Coghill – May 1995 – has suggested that the IEE Working Party's conclusion that there is no proven harmful effect due to low-level, low-frequency electromagnetic fields is the result of power-utility bias in its membership. As IEE Chief Executive, Dr. Williams, has pointed out – *EW+WW* August 1995 – that accusation is demonstrably untrue. What is true is that Mr. Coghill's own business has been concerned with advertising devices to 'reduce' the effects of power frequency magnetic fields. It might be thought to be in his commercial interest if any biological effect could be demonstrated. It is perhaps disingenuous of him not to mention this.

Bernard Jones – Oct 1995 – is "confident that magnetic fields produce brain cancer etc." although he gives no evidence to support his view. I hope that Mr. Jones understands that in science it is impossible to prove that something does *not* exist – there may always be some statistically insignificant effect that is masked by other sources. It is possible to show that an effect *does* exist when an effect that is statistically significant can be measured consistently.

Whatever your correspondents may say there is very little evidence that we have reached this stage. If they know otherwise perhaps they will refer your readers to the evidence.

Meanwhile, as an active, but independently minded, member of the IEE let me assure Messrs Coghill and Jones that whenever real significant evidence is available to support the hypothesis that low-level electric or magnetic power-frequency fields have a significant effect upon our health then I – along with many of our members – will work to ensure that the IEE gives it the widest possible publicity. Until then it will be in the public interest to stick to known facts rather than creating unnecessary concern.

Colin W. Davidson
Edinburgh

Is EMC approval necessary?

The recent EEC directive on emc requires all electrical equipment to be tested for radiation levels.

There are solid theoretical grounds for declaring that all equipment below a certain power automatically meets the requirements. This is because the radiated power cannot exceed the total power absorbed. The standard does not make such an exemption. If an exemption were made, and the power levels turned out to be reasonable ones, then the savings for electronics companies could be huge.

A piece of equipment only uses so much power. If all that power were converted into radiated electromagnetism, you could calculate from theory the maximum possible field strength according to distance from the device. So, for example, a device that absorbs only 2W could only ever radiate 2W of power as a maximum. In fact, it is hard to make a transmitter with 50% efficiency so it is reasonable to assume that only half the total power could possibly be radiated as electromagnetic radiation. Often, heat dissipation of the equipment is known and so you can subtract the heat power from the possible radiated power.

This would permit the radiation limit to be calculated purely from theory, based on the total power absorbed. For many items, that use very little power and so could never possibly radiate too much, such a calculation could avoid costly tests which would otherwise be necessary.

One might argue that the radiation field could be shaped in such a way that a tightly focused beam is emitted, so that the field in one direction might be very much higher than a theoretical average. To argue thus seems to me perverse, given that the equipment is not designed to generate focused beams and the difficulty of generating a focused beam. Also, if the equipment involved is small while the measurement distance is relatively large – 10m is the European standard – then you might argue that the field would be relatively uniform at this distance from the object. Similarly, the physical size of the equipment would have an effect – equipment that is small compared with the wavelength could not possibly generate a focused beam.

European standards make no allowance for such a theoretical calculation. I would be interested to know if such a calculation would be reasonable so that one can argue for a change in the standard or an automatic exemption – thus saving companies a lot of money.

I am not enough of an rf engineer to make the relevant calculations. However from my university text book on electromagnetic theory, it seems that if the equipment were a perfect dipole radiator then anything under 2W could not possibly exceed the emc limits. As a result, it would not need testing.

It would be interesting for someone with some expertise on this field to make the calculations.

Chris Bore
Bores signal processing

Sallen & Key distorted

Mr Ryder's conclusion that the Sallen and Key filter configuration is 'not generally usable for hi-fi' would be worrying if it were true – especially in view of the fact that I used this circuit repeatedly in the *LA100* Audio Analyser. This unit probably features more than any other, worldwide, in the automated checking of distortion and quality on the broadcast and studio equipment that originates his 'hi-fi'.

The truth is that the Sallen and Key filter has much to commend it, including low component count and unity gain – precise unity gain; for ever. These features are regardless of component values or temperature. This is especially valuable in multistage measuring equipment accurate to a hundredth of a decibel, but it is also useful in maintaining accurate gain and channel balance in multistage audio equipment without trimming.

The second-harmonic distortion problem he has discovered is not a direct function of resistance values, but is caused by a varying common mode capacitance which shunts his 470pF input capacitor. This causes signal-dependent modulation of its value which probably originates in reverse biased substrate-isolation or protection diodes in the op-amps. These behave as varicaps.

The effect can be puzzling because it only occurs near the cutoff frequency. At this point, the capacitors are working hardest. The effect arises mostly as a result of amplitude-dependant modulation of the filter phase shift.

Mr Ryder's choice of a high Q emphasises the problem in two ways: by requiring a large capacitance ratio, and hence a low value input capacitor, and by generating a large phase shift. His choice of values is far from optimal. Changing the 110kΩ resistors to 3.3kΩ, with a corresponding increase in capacitor values, will swamp the effect and produce very acceptable figures without loading the op-amp output too much.

In most applications, where a capacitor ratio of three or four to one is more common, the *TL072* works well. For critical applications, the *NE5532* performs much better in this respect, at the cost of higher power consumption.

Peter J Skirrow
Lindos Developments
Suffolk

Re-inventing the wheel

After reading the article 'maximising power in class-C' I must conclude that each generation of radio engineers is intent on re-inventing the wheel.

I was surprised to find that there was no reference to W.L. Everitt's book 'Communication Engineering'

the second edition of which was published in 1937. In it he gives a complete mathematical treatment for maximum power output from class-B and class-C amplifiers. I cannot say if the same analysis appeared in the first edition of 1931.
S F Brown
Oswestry
Shropshire

Modulating misguidedly?

I have just read 'Modulating linearly' by Ian Hickman in your July 1995 edition. While presenting an interesting method of reducing the intermodulation products at the modulator, I believe that the effort is misplaced. It is out of band products from ssb transmissions that are offensive and cause adjacent channel interference.

Most ssb transmitters employing dsb suppressed carrier modulators pass the signal through an ssb filter. This removes the unwanted sideband, frequencies outside the required modulating bandwidth, and incidentally out of band intermodulation products generated by the modulator. The offending out of band intermodulation products are usually generated at the power amplifier final stage where trade-offs between linearity efficiency cost and distortion are made.

Surprising though it may seem, in-band intermodulation products may be of advantage in voice ssb transmission improving modulation efficiency and transmitted power. By way of expiation of my statement it is speech processing incorporated in many ssb transmitters that proves this point. The peak to average form factor of the voice waveform is high which leads to low average transmitted power. Voiced sounds consist of harmonics of the vocal chord vibrations amplified and frequency shaped by resonant cavities in the vocal tract, head and chest. This results in a multi-tone spectrum of which a few tones are dominant in the mid-audio band of similar frequency to those used by Hickman in his article i.e. lkHz and 1.2kHz. If one applied these tones of equal level after ssb filtering to an ssb transmitter with auto level control so that peak to peak amplitude were the same as for a single tone, transmitter average power would be 50% that of a single tone.

Speech processing is often accomplished by clipping the ssb signal in the intermediate frequency after a stage of ssb filtering and removing the out of band products in a second stage of ssb filtering. Were the tones applied to a circuit of this type with 20dB of clipping then the transmitted power would be about 80% of that of a single tone.

In this instance a high degree of in band intermodulation is produced and it is in the intermodulation products that the additional transmitted power is contained.

These intermodulation products are not unrelated to the frequencies present in the original modulation and the brain interprets them as though the resonant cavities of the vocal tract were of lower Q and thus correlating the frequencies to produce additional intelligibility in poor signal to noise conditions.
John West.

Audio power – is current feedback the solution?

John Watkinson's letter in the October issue highlights the trend of designing ever-more complex audio amplifiers while the loudspeaker system remains the largest source of distortion.

Admittedly some amplifiers can handle the complex and variable impedances of various speaker systems better than others, but none seems to overcome the basic problems caused by the conventional moving coil speaker. Several manufacturers have attempted to improve speaker performance by using motional feedback or an active cross-over with separate amplifiers. However, this requires careful matching of both electronics and hardware.

The main purpose of an audio amplifier is to push current through the voice coil in order to move the cone. Traditional amplifiers use voltage feedback to ensure that the voltage applied to the speaker is controlled accurately while hoping that its impedance remains reasonably constant. Obviously any variation of impedance, either with frequency or cone excursion will cause some non-linearity in the current thus causing distortion. A look at the impedance curve of any good quality speaker will show surprising variations.

One possible solution that does not appear to have been considered is the use of current feedback rather than the more common voltage feedback.

A similar problem arose in the design of television receivers, where the vertical deflection coils are driven with a 50 or 100Hz waveform. Any linearity or amplitude errors are easily visible on the screen.

Early designs used a conventional voltage amplifier – often very similar to the audio amplifier – however the current amplitude and thus picture size usually varied as heating of the deflection coil caused its resistance to change. Thermistors were often added to try to compensate for temperature effects. However, the modern solution is to use current feedback which overcomes the problem of any change in load impedance: refer to circuit diagram.

Using current feedback should, in theory, null the effects of resistance changes caused by voice coil heating or by using different lengths or sizes of speaker leads. An additional advantage should be good short circuit protection without the clipping or distortion of a conventional protection system.

The current sensing resistor will cause some power loss, but in an 80W 8Ω system this would amount to less than 1W with a 0.1Ω resistor.

It would be interesting to hear if anyone has tried this approach and if it produced any improvement in the overall performance.
J.R Allison
Bradford
Yorkshire

Square-law rules – OK

Ian Hegglun is to be congratulated – not only for the content and design performance achieved but also, for actually quoting power output stage open loop performance measurements supported by waveform traces.

Prior to reading 'Square-law rules' in the September issue, the existence of an ongoing controversy and the absence of any reasonably comprehensive or explicit open-loop performance data had lead me to the conclusion that, as currently used, neither bjts or mosfets were the ideal audio power output stage device and that I would have to design my own avoiding current design conventions.

My own approach to a linear power output stage – ie not worse than 1% thd – was to investigate a fully differential solution using a self balancing active bridge configuration. An experimental prototype output module using power Darlington bjts has been tried. Provisional results with this output stage configuration have shown that nearly constant open-loop unity gain, ie V_{out}/V_{in}>0.99, is achievable. This is subject to gain/symmetry trimming, for a voltage swing of ±18V into a 15Ω resistive load.

Open loop frequency response is flat to at least 200kHz, output resistance was 0.5Ω, and input impedance around 600-800Ω. Quiescent power consumption was low, at 6 to 8W, from a single 25V dc supply. This reflects an optimum stable bias setting. Normal Class B/AB or A preconceptions do not apply to this circuit.

Listening tests confirmed measured expectations. These were effected by terminating one channel of my existing stereo amplifier (thd 0.2% at 1kHz and 10W) using the module as a relay amplifier to drive the right-hand speaker. This was a vintage 15Ω Goodmans Maxim.

Performance of this module is such that only a relatively modest amount of global feedback at a maximum of 36dB need be applied to the preceding voltage/impedance transformation stage. The level of 36db is also the maximum theoretical global feedback that will maintain a critically damped response in a single dominant pole design.

Thus for a cost effective solution I consider the answer to lie more in matching circuit to device and in designing the system to meet the needs of the application. This includes matching speaker ratings to amplifier ratings and speaker system to auditorium.
E J Chadderton,
Brokenhurst,
Hampshire.

Considering a valve design?

For readers considering a valve design, I would mention that an ultralinear connection enables a pair of *EL34*s to deliver up to 30W with less than 1% thd rather than 4% thd without feedback. This also compares with 14W max from triode connected *EL34*s at less than 1% thd.

A suitable push-pull output transformer for this connection, Gardners Type *AS 7034*, might still be obtainable from Gardners Radio, Christchurch. This had both 20% and 43% screen grid tappings.

I had also used the ultralinear connection to very good effect, many years ago, in a two valve 3.5W *EF86/EL84* single-ended Class A design for mono but regrettably replaced this with a pair of 10W Toby/Dinsdale when stereo was introduced.
E J C

Hot audio power

Jeff Macaulay's article 'Hot audio power' in the October issue contained two errors in the components list. Grid resistors $R_{6,9}$ should be 560kΩ, not 60kΩ and R_{11} is 680Ω, not 6k8. Apologies.

Advances in digital video

Reg Miles discusses developments highlighted at the broadcast and professional exhibition – Vision '95.

The first camcorder to use hard disks instead of tape is here. Developed by Avid and Ikegami for news gathering, its advantage lies in its facility to randomly access any part of the disk. This enables recorded clips to be viewed, modified and edited into sequences in the field for transmission by microwave or satellite link to the studio – as well as providing all normal functions.

Called *CamCutter*, the recorder itself has no moving parts. These are confined to a removable, sealed and shock-proof pack containing two 1.2Gbyte disks. These provide up to 20 minutes of storage.

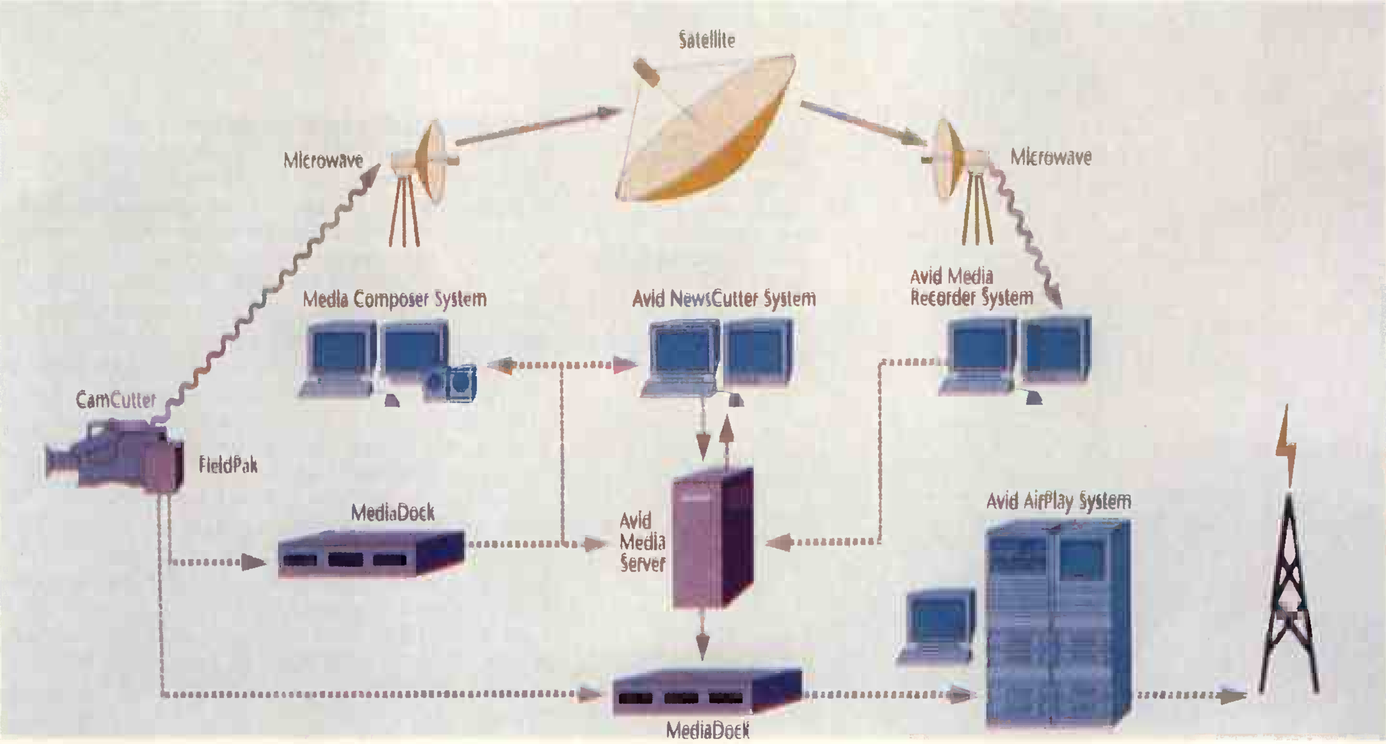
Video is recorded in component form, compressed at 7:1. Four 48kHz 16bit audio channels are available – although fewer can be used to increase recording time. An optional unit accepts up to three disk packs for editing and broadcasting in the studio.

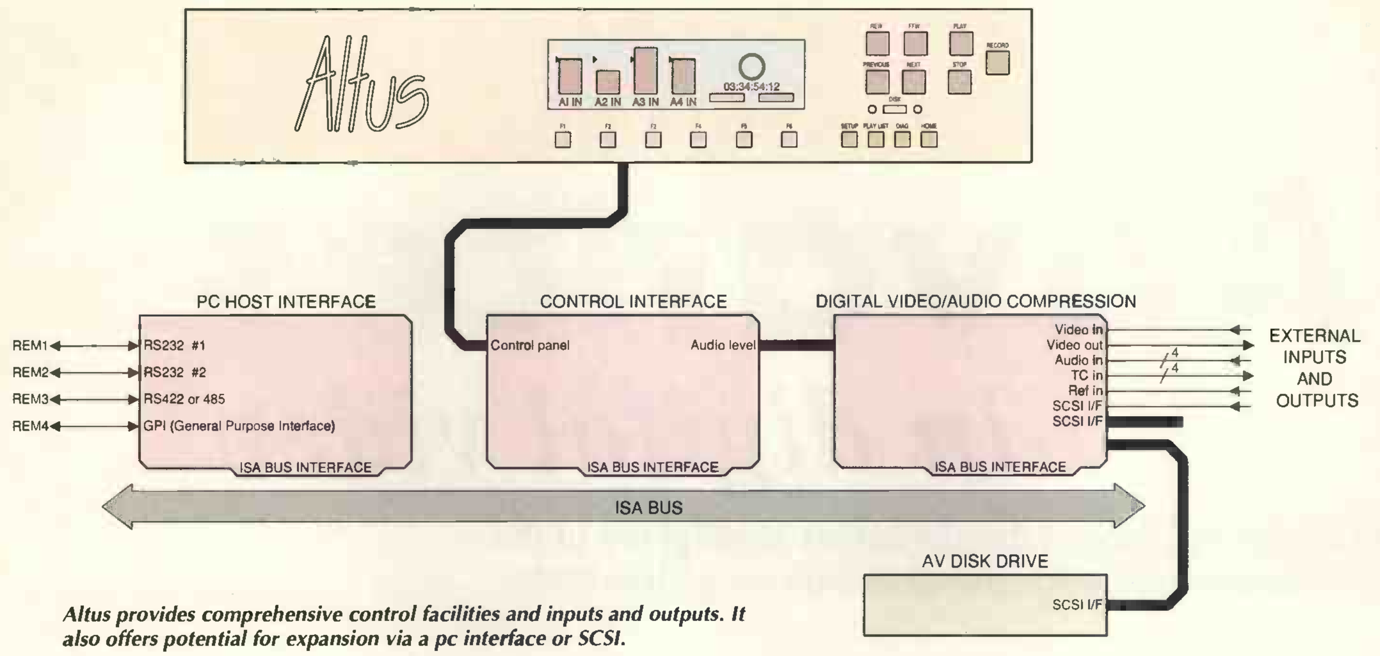
Digital editor with pc interface

Hi Tech Systems' *Altus Digital Disk Recorder* is a rack mount/desk system, designed to be a cost effective alternative to the video tape recorder. As with *CamCutter* there is random access to the recorded material so that frames and clips are instantly available, allowing a variety of playback functions. In addition, there are all usual functions. Configuration and set-up is simply achieved using 'soft' keys – including adjustment of the compression ratio.

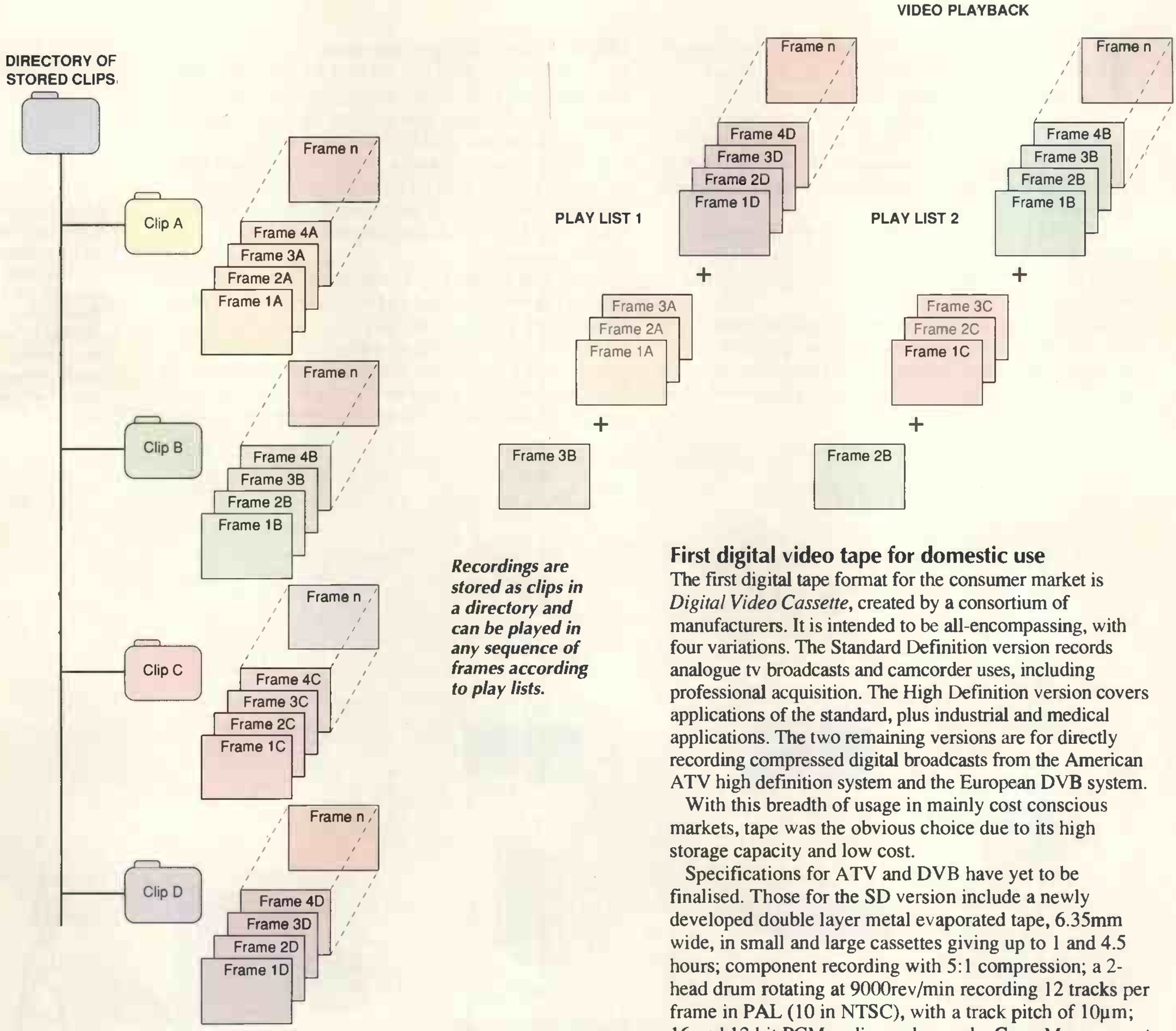
Connecting the recorder to a pc provides further controls over monitoring, playback and recording, using a suite of Windows programs. There is a choice of 2.1, 4.3 or 9.1Gbyte disks, recording approximately 10, 20 or 40 minutes depending on file size and image complexity. An SCSI bus connector is provided to connect an external RAID (redundant array of inexpensive disk drives) system.

Material from CamCutter can be broadcast immediately, brought back to the studio for editing, or stored on a central media server for enterprise-wide access.





Altus provides comprehensive control facilities and inputs and outputs. It also offers potential for expansion via a pc interface or SCSI.



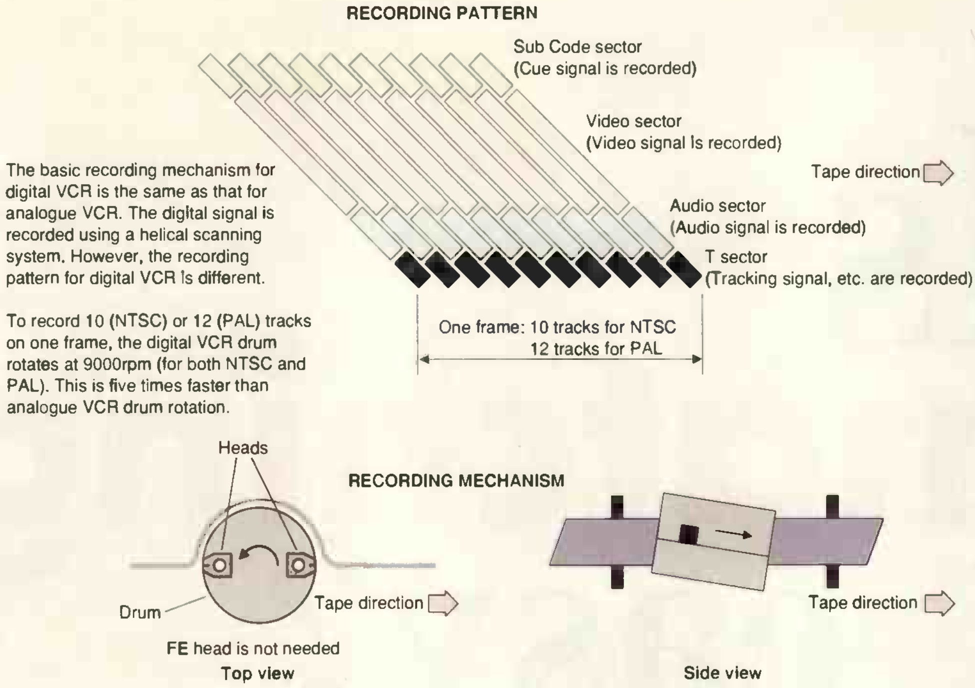
Recordings are stored as clips in a directory and can be played in any sequence of frames according to play lists.

First digital video tape for domestic use

The first digital tape format for the consumer market is *Digital Video Cassette*, created by a consortium of manufacturers. It is intended to be all-encompassing, with four variations. The Standard Definition version records analogue tv broadcasts and camcorder uses, including professional acquisition. The High Definition version covers applications of the standard, plus industrial and medical applications. The two remaining versions are for directly recording compressed digital broadcasts from the American ATV high definition system and the European DVB system.

With this breadth of usage in mainly cost conscious markets, tape was the obvious choice due to its high storage capacity and low cost.

Specifications for ATV and DVB have yet to be finalised. Those for the SD version include a newly developed double layer metal evaporated tape, 6.35mm wide, in small and large cassettes giving up to 1 and 4.5 hours; component recording with 5:1 compression; a 2-head drum rotating at 9000rev/min recording 12 tracks per frame in PAL (10 in NTSC), with a track pitch of 10μm; 16 and 12 bit PCM audio modes; and a Copy Management System.



Recording arrangement of the standard definition Digital Video Cassette format, showing how the tracks are laid down in sectors by the two-head drum.

DVCPRO differs in using more robust metal particle tape, which is run at double speed to increase track pitch to 18μm, and the addition of a control track and cue channel. The *HD* version uses metal-evaporated ME tape run at double speed to record twenty 10μm tracks per frame. Horizontal resolution is about 500 lines for SD – 25% greater than Hi8/S-VHS – and 600 lines for HD, with double the vertical resolution.

Developments in video editing

Tape based video editing is a laborious process requiring scenes to be copied from player(s) to a recorder in real-time. Each scene has to be found by winding the tapes. This is linear editing. Recently non-linear editing has been developed to exploit the random access facility of hard disks.

In non-linear editing, the video and audio is digitised, usually compressed, and transferred to disk. Each scene is represented on-screen by a still, and these scenes can be instantly accessed in any order, rearranged, trimmed, the audio edited, and effects and captions added. If it is not right it can be returned to a previous 'undo' level – instead of starting again as with tape. And systems can be networked for several editors to work at once.

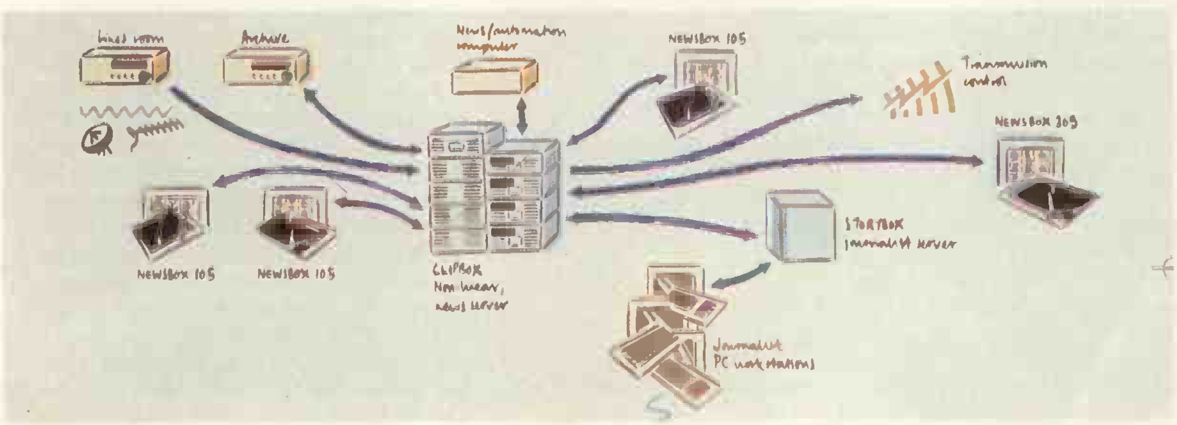
The completed programme can then be output to tape, disk or to air. Alternatively, the computer can generate an edit decision list which tape machines can use for automatic linear editing. Some systems are linear/non-linear hybrids, capable of editing with video recorders and hard disk.

Non-linear editors are available as either plug-in boards and software or a complete computer system. A new trend is systems packaged for specific uses. One of the main focuses is on news, with systems from Avid, Lightworks and Quantel.

Sony's new system is intended for live broadcasts: two channels of hard disk storage provide quick replay and editing of highlight scenes while the live video continues to be recorded. An endless recording capability enables recording until the disk is full and then re-recording from the beginning. It can store up to one hour of video with 160kB per frame, 6:1 JPEG compression. The JPEG file size are selectable from 60:1-6:1. Additional drives can be installed to expand capacity to five hours.

JPEG is a popular method of compression, but it does rely on large, fast hard disks to function at speed. Eidos' new system takes a different approach, employing removable 1.3Gbyte magneto-optical discs, as well as hard disks. This change is made possible by Eidos' proprietary compression engine, Optimizer, which is more efficient than JPEG and thus compensates for the lower performance of MO discs. This can be swapped as easily as tape cassettes.

Non-linear editing began in the broadcast field and is working its way down through professional applications to consumers – with their lower quality requirement. ■



Typical use of non-linear editing systems for news productions, with Quantel Newsbox edit suites networked with a clipbox server to all other facilities.

Is matching easy?

On the surface, matching a circuit's output to the next one's input appears straightforward, but as Ian Hickman shows, this important technique can be a source of loads of problems.

Matching a load to a source or *vice versa* ensures that the maximum possible power will be transferred, as stated by the well known Maximum Power Theorem.

In electronic signal processing, the matched condition is usually preferred, but this is not necessarily so in other applications. For instance, internal resistance – 'source resistance' – of a new 1.5V cell is around the 1Ω level, whereas the resistance of a 1.5V 300mA lens-end flashlight bulb is 5Ω when lit. This ensures that five sixths of energy supplied by the cell finishes up where it is wanted, producing light. In the matched case, a 1Ω bulb might produce more light, but 50% of the energy would be wasted simply warming up the battery. If you want extra light, it makes more sense to use more cells in series and a higher bulb voltage, which still only draws 300mA.

In other cases, a source is, by design, simply incapable of supplying a matched load, a good example being a 660MW turbo-alternator. With this, the design minimum value of the load is about 30 or more times the internal resistance – overload protection devices would trip long before the matched load condition were met.

Matching standards

In the design phase of electronic modules where matching is important, such as tv camera signal processing chains, telephony cable repeaters, radio receivers and a variety of transmitters, extensive use is made of test equipment such as signal generators, spectrum analysers etc.

Sources are designed to produce an accurately known output level, such as –l0dBm, into a matched load. A –l0dBm level is a level of –l0dB relative to a power of lmW delivered to a matched load, or 100μW. In telephony, where a 600Ω impedance system is common, lmW corresponds to 0.775Vrms.

Telephone engineers often define OdBm as meaning lmW in 600Ω. But the more common usage is to define it as lmW in whatever the system design impedance is. This corresponds to 225mV in a 50Ω system, common in rf equipment, 273mV in 75Ω, common in tv baseband signal working, or 387mV in 150Ω, in twisted pairs in underground cables.

In radio-frequency testing, a module's input port is commonly driven by a signal generator with a purely resistive output impedance of 50Ω, and its output port is terminated in the 50Ω resistive input impedance of a spectrum analyser. In the case of the module's input, power delivered to it will be very close to that delivered to a 50Ω resistor. This occurs even if the module's input impedance departs fairly markedly from 50Ω resistive.

This is illustrated in **Fig. 1a)** and **b)**, where things have been normalised to unity, i.e. a generator with a 1Ω source delivering lW to a

Binomial theorem

If R_L in Fig. 1a) increases by 1% to 1.01Ω, then total circuit resistance – R_S being 1Ω – increases by 0.5%. Thus current *decreases* by 0.5%. Power P dissipated in R_L is given by $P=i^2R_L$. If i decreases by 0.5%, then i^2 decreases by 1%. But R_L has increased by 1%, so the product is – virtually to a first order – unchanged. This is a result of the Binomial Theorem, but can equally be verified by working out $i^2R_L=(2/(1+1.01))^2\times1.01$ on a pocket calculator.

Some results from the Binomial theorem:

$(1+\delta)^2=1+2\delta$	$(1-\delta)^2=1-2\delta$
$(1+\delta)^{-1}=1-\delta$	$(1-\delta)^{-1}=1+\delta$
$(1+\delta)^n=1+n\delta$	$(1-\delta)^n=1-n\delta$
$(1+\delta)^{-n}=1-n\delta$	$(1-\delta)^{-n}=1+n\delta$

Note that these results only apply if $\delta \ll 1$ and n is smallish, so that second order and higher terms are insignificant.

nominal 1Ω load. For reasons explained in the panel, provided R_S=1Ω, the power in the load is close to 1W even if R_L varies. If R_L is one third of an ohm, or 3Ω, power in the load is 750mW, or only −1.25dB for a voltage standing wave ratio of 3:1. However, although power in the load is not very sensitive to variations in R_L, power dissipated internally in the source, and hence total power supplied by the 2V ideal generator, varies markedly. Figure 1 b) shows that total power supplied by the 2V generator varies from 4W for a short circuited load, down to zero when R_L equals infinity.

Figures 1a) and b) show the situation at 0Hz, or dc, where the effect of any incidental reactive terms in R_S and R_L can be ignored. The maximum power theorem applies equally at ac, but with the added complication that in general we are dealing with impedances rather than pure resistances. This is shown in Fig. 1c), where inductive and capacitive components are shown in R_S and R_L respectively. However, it could equally well be the other way round, or both reactances could be of the

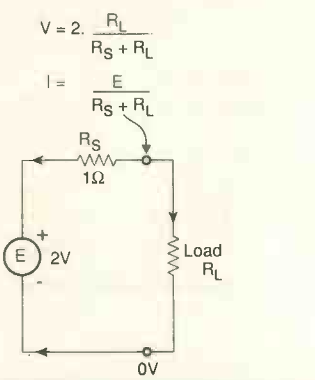


Fig. 1a). Normalised source and load, illustrating the maximum power theorem.

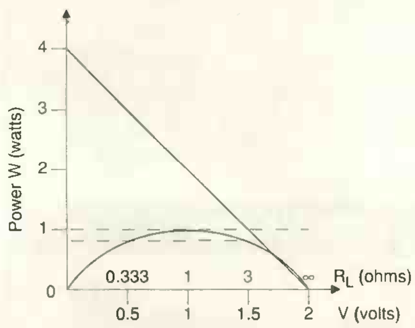


Fig. 1b). Power in the load resistor R_l as a function of its value, when R_s equals the design system impedance of 1Ω (curve) and total power supplied by the generator (sloping line).

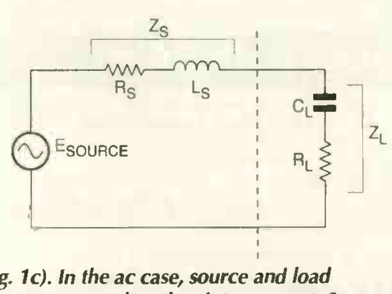


Fig. 1c). In the ac case, source and load reactance must be taken into account. Source to the left of the dotted line, load to the right.

same sign – positive for inductances and negative for reactances.

Conjugate matching

Using Fig. 1c), assume that reactance of inductive component L_S of the source equals that of the capacitive component C_L of the load at the frequency of the sinewave source E_{SOURCE}. In this case the two cancel each other out, power in the load being determined purely by the values of R_S and R_L. This is known as the conjugate matched condition and can only occur at the one frequency – conjugate matching is inherently narrow band. For this reason, signal generators are designed such that Z_S is as nearly as possible purely resistive, any L_S or C_S being ideally zero. A similar comment applies to the input impedance of measuring instruments, such as power meters and rf spectrum analysers.

If Z_S in Fig 1c) is purely resistive and equal to the system design impedance, power in the load is relatively independent of its exact value. But what if the dotted line in Fig. 1c) had been drawn horizontally? This makes what is now Z_S the load (exactly 1Ω, say) and variable Z_L is now the source resistance.

Considering the dc case, power in a fixed load of 1Ω as R_S varies from zero to infinity is now given by the vertical distance between the curve and the sloping line of Fig. 1b). If R_S varies, power in the load varies wildly, even if the load is a pure resistance equal to the system design impedance. Is this important? The following cautionary tale shows that it is.

Is matching input to output enough?

Some years ago, the company I worked for was developing modules for an advanced all-band surveillance receiver. One engineer designed the front-end half-octave filter module, another the rf amplifier and first mixer module, another the first IF and so on through the second and third mixers an IFs.

The design aim was that during servicing, replacing any or all of the modules should leave the overall performance within specification. To this end, a voltage standing wave ratio tolerance was placed on the input and output impedance of each module.

As development progressed, insertion loss or gain of each module was checked using a signal generator and spectrum analyser, or a network analyser, as available. The analyser was also used to check port impedances. Thus module inputs were driven from a respectable 50Ω source, and their output checked with a faultless 50Ω load. Nevertheless, complete receivers exhibited a range of performance which was outside the specification limits.

Unlike the situation on test where each module port is connected to a 50Ω interface – at the interface between modules in use – both port impedances could be different from 50Ω.

Graphs documenting this effect have appeared in an issue of the Marconi house magazine but I have failed to unearth it. So I worked it out again for the simplified case where both R_S and R_L vary, but both are resistive. **Figure 2** shows how power in the load R_L varies with the value of R_L for a series of different values of R_S from 0.25-2Ω. Source emf behind R_S is 2V as in Fig. 1a) – call this Case I.

When R_S is 1Ω, maximum power naturally results in a matched 1Ω load. This curve is the same as in Fig. 1b). In accordance with the maximum power theorem, when R_S equals 0.25Ω – top curve – maximum power of 4W occurs in the load when its value is also 0.25Ω. To a nominal 1Ω load, an R_S of 0.25Ω delivers 2.56W or +4.1dB. Significantly, the power changes rapidly for small deviations of R_L from unity. Likewise, when R_S is 2Ω, maximum power of 0.5W occurs in a 2Ω load, while when R_L is 1Ω, output is −3.5dBW.

Case I applies where source emf is what it should be – twice the voltage across a matched load – but source resistance R_S is incorrect. This corresponds to the case of a signal generator where the output impedance defining resistor is damaged. The instrument is otherwise unchanged from new, perhaps a rather unusual case, or perhaps due to the inadvertent application of rf power to the signal generator's output.

Where output impedance is poor

A different set of circumstances arises in Case II. A module with a poor output impedance has been set up to deliver its rated output to a resistive load equal to the system design impedance, for example 50Ω. In this case, internal emf E_{SOURCE} will have been effectively adjusted – in terms of the normalised circuit of Fig. 1 – to something other than 2V, so as to deliver 1W into a 1Ω load. Thus if R_S is lower than 1Ω, E_{SOURCE} will have been set to less than 2V, and to more than 2V if higher. Power delivered to the load R_L, as a function of the value of R_L for various values of R_S, is shown in **Fig. 3**.

Since E_{SOURCE} has been adjusted to deliver 1W into 1Ω, whatever the value of R_S happens to be, all curves pass through 1W when R_L is 1Ω. But only in the case where R_S equals 1Ω is the curve horizontal for R_L of 1Ω. This gives the relative independence of load power *versus* R_L that obtains when the value of R_S is correctly set at the nominal system impedance. Nevertheless, the variations of power delivered with variation of both R_S and R_L are much less in this case than Case I, permitting the results for Case II to be plotted in Fig. 3 with a vertical scale twice that of Fig. 2.

Matching in filter designs

Correct matching is particularly important where filters are concerned, as the following illustrates. Some time in the '80s, my then boss had a crisis with the company's new hf receiver – which was already over budget and overdue. The 20-30MHz sub-octave filter was far too narrow, and excessively lossy to boot. "It's got to be fixed, but don't spend more than a week on it, even if the conclusion is that we have to relax the specs."

Having spent the rest of the morning getting the necessary test equipment together, I was

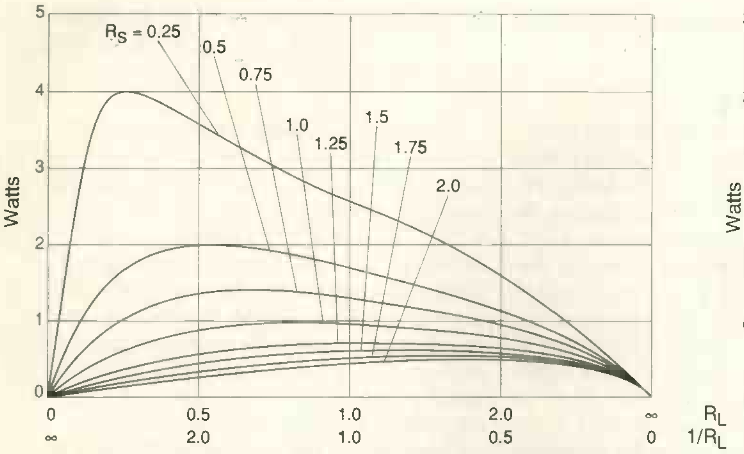


Fig. 2. Power in load resistor R_L as a function of its value, showing curves for several different values of R_s. Source emf E_{SOURCE} is as in Fig. 1a).

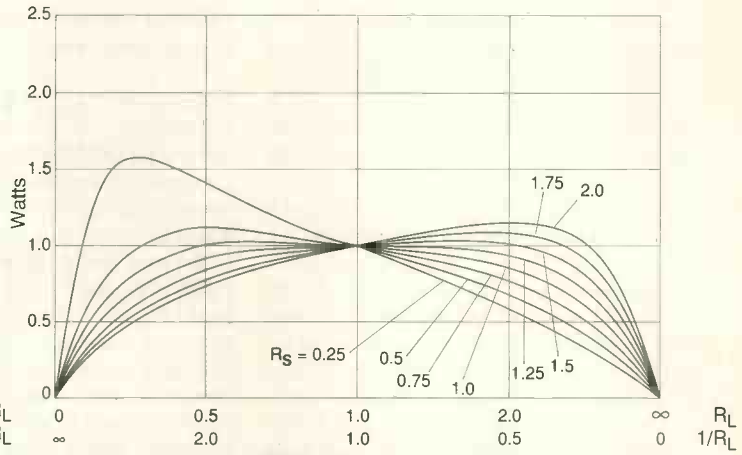


Fig. 3. Power in the load resistor R_l as a function of its value, showing curves for several different values of R_s. Source emf E_{SOURCE} adjusted to give 1W into a 1Ω load resistor R_L, for each value of R_s.

able to report half way through the afternoon that the filter was now working fine. So it should, for it was a seven pole elliptic design straight out of the reference.

Checking the values on the circuit diagram confirmed that the engineer who had designed it had done his denormalising sums right. Capacitors on the board were also all correct, and the coils all capable of being tuned by means of their slugs to the correct inductance.

Output of the filter module was normally connected to the rf/first mixer module, which presented a good low voltage-standing-wave ratio input. But before getting there, output of the 20-30MHz filter, when selected, had to pass through a number of band-select relays and board tracking – which looked distinctly capacitive. As far as the filter was concerned, this capacitance was part of the load, which should have been purely resistive but wasn't.

Reducing the value of the final shunt capacitor in the filter effected an improvement, and further reduction made it better still. In the end, it turned out that no capacitor was needed at all, the circuit strays equalling the design value of the filter's final shunt capacitor. But in the end I settled for 1.8pF, to avoid C_{99} on the circuit diagram being shown as 0pF.

The final capacitor in the 15-20MHz was also reduced in value somewhat, the lower frequency filters being in spec. due to the much larger values of their final capacitors. ■

Reference

Geffe, P R., 'Simplified Modern Filter Design' Iliffe Books Ltd, London 1964.

NEW PRODUCTS CLASSIFIED

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ACTIVE

Arrays

5Mgate asics. Up to five million gates and 10Mbyte of memory, or nearly 50 million transistors can be shoe-horned into the *G10* series of asics from LSI Logic; to point these figures up, LSI say that such a density is equivalent to more than eight Intel *P6*s. Gate delays are down to 50ps. There is a library of complex elements such as three-transistor memory structures and mixed-signal circuitry providing 12-bit and 270MHz data conversion. LSI Logic Europe plc. Tel., 01344 426544; fax, 01344 481039.

Fpgas. Xilinx offers the *XC8100* family of field-programmable cmos gate arrays using the company's *MicroVia* antifuse technique and sea-of-gates architecture to provide virtually 100% of 1000-9000 usable gates. Logic cells are configurable to carry out synchronous, combinational and three-state logic functions. Supply is 5V and 3.3V. Xilinx Ltd. Tel., 01932 349401; fax, 01932 349499.

Microprocessors and controllers

PICs with voltage comparators. Microchip's *PIC16C622* eprom-based PIC has two analogue comparators, a programmable voltage reference and the first 4V low-voltage protection to be offered. It uses the risc-like architecture common to the series, although here, the program and data are in different memories, so that separate buses are used to fetch and execute in one cycle. There is 2K/14 on-chip programmable memory. Polar Electronics. Tel., 01525 377093; fax, 01525 378367.

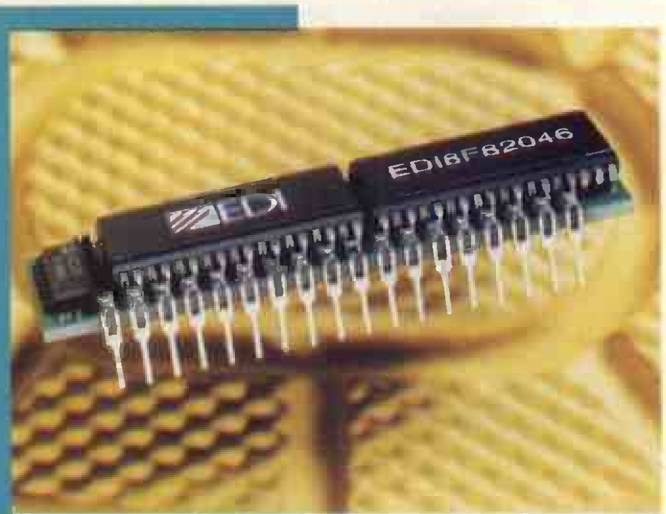
Audio processor. Philips offers the *TDA1548T* single-chip audio processor, which includes Bitstream filters for digital de-emphasis, volume and tone control, Bitstream d-to-a converters and headphone amplifiers, providing all audio functions needed for CD, MD and DC personal stereo players. It is compatible with I^{2}S or LSB-justified serial input data and provides 1.7Vpk-pk audio into 32Ω. Low voltage and the avoidance of the processing needs of purely digital volume and tone controls reduces power requirements from the 3V battery pack. Two external components – capacitors for analogue post filtering – a further decoupling capacitor and three potentiometers are needed for a full stereo output. Philips Semiconductors (Eindhoven). Tel., 00 31 40 722091; fax, 00 31 40 724825.

Graphics accelerator. Advance Logic introduces the *ALG2046* dram-based, 64-bit PCI graphics accelerator, which contains graphics engine, d-to-a converter, clocks and BIOS in the one 208-pin package; with two 256K/16 drams, the accelerator forms a complete graphics sub-system that can be up-graded by simply increasing memory to 2Mbyte. Silicon Concepts Ltd. Tel., 01428 751617; fax, 01428 751603.

Graphics controller. *CL-GD5436* is the fastest of Cirrus's dram-based graphics controllers, having a 64-bit graphics engine and enhanced bit-BLT capability with larger, double-buffered registers for faster performance with Windows, NT and OS/2, in 24-bit colour. Features include a direct, glueless interface to the PCIbus and the device supports extended burst cycles on the bus. It can be used with *PowerPC* and other PCI-based systems as well as PCs. Graphics modes up to 1024/768 in true colour and 1280/1024 in 256 colours. Cirrus Logic Inc. Tel., 01727 872424; fax, 01727 875919.

486 embedded processor. A family of 486 processors intended for use in equipment other than pcs is announced by National Semiconductor. The *NS486* has a three-stage pipeline, against the pc version with five, and has no co-processor. It does have extra peripherals, including a dram controller, programmable interrupt controller, lcd controller, PCMCIA controller, etc, and the device gives a compatible 486 instruction set, a 32-bit core, 25MHz working and 12mips performance. Power-saving modes are provided. National Semiconductor GmbH. Tel., 0049 1805 32 7832

16-bit microcontrollers. Hitachi has the *H8S* series of 16-bit devices, which achieve 10Dhrystone mips at 5V with a 20MHz clock, a multiply or multiply-and-accumulate instruction taking 200ns, and provides 67mips/W at 2.7V (48 at 5V). Power savings have been obtained by making some of the internal clock-driven functions into event-driven logic circuits, not needing a constant clock input. Benefits are lower switching frequencies, giving lower power consumption, and lower emi; power dissipation is 75mW at 10MHz. Peripherals provided include a 1MHz 10-bit a-to-d converter, a 32-bit data transfer controller needing no cpu intervention or dma controller and an enhanced memory interface capable of accessing fast page-mode dram and burst-mode rom. Hitachi Europe Ltd. Tel., 01628 585163; fax, 01628 585160.



Memory chips

2Mb/8 memory modules. EDI has introduced a 2Mb by 8 sram having an access time of 20-45ns. The *EDI8F82046C* uses four 412/8 srams and is in a 36-pin dip with a Jedec-approved pinout designed for noise immunity and easier up-grading. EDI (UK). Tel., 01276 472637; fax, 01276 473748.

C-programmed controller. Greening Technology's *Micro Genius* microcontroller is smaller than a credit card and costs £65 for one off. Programmed in C, it has 14 configurable i/o lines, an RS232 port, 32K of sram, up to 512K of eprom or 256K flash eprom, a watchdog timer, psu and an analogue input. It can hold up to 20000 lines of C code, including multitasking and access to a C library. In development, the controller is connected to a pc and code compiled to flash eprom on the board. Greening Technology Ltd. Tel., 0116 2796500; fax, 0116 2796501.

Optical devices

Led strips. Board-mounting strips of five, six or nine 3mm leds on 0.25in centres are introduced by Lumek. They are flexible, can be divided into smaller sections and are available in any combination of red, super red, yellow, green or amber, light intensity or lens type, with bi-colours as an option. Lumek Opto/Components Inc. Tel., 001 708 359-2970; fax, 001 708 359-8904.

PASSIVE

Passive components

Silver mica capacitors. ACL's *D-15* range of high-voltage, precision silver mica capacitors are in resinous and/or clear epoxy and come in 100Vdc, 300Vdc and 500Vdc ratings, in values in the 1-500pF, 1-820pF and 1-510pF ranges respectively. Temperature coefficient is 'low' and tolerance is 0.25pF and 0.5%. Four operating classes are available, from –40°C to 85°C to –55°C to 150°C. Europa Components & Equipment plc. Tel., 0181-953 2379; fax, 0181-207 6646.

Connectors and cabling

Optical-fibre distribution. Molex pre-terminated distribution panels for optical fibres provide high-density connections and versatility and avoid the need for splicing on site. Each panel takes 144 fibres and is mounted in a 19in or 23in relay rack; a selection of adaptor interfaces provides flexibility in configuring the system. Molex Electronics Ltd. Tel., 01420 477070; fax, 01420 478185.

Displays

Better lcd. Crystaloid has improved the readability of liquid-crystal displays. In earlier types, the indium/tin oxide electrodes carrying current in the display have a tendency to reflect light, making inactive segments of the display appear active. Crystaloid has a new index-matching technique in which a coating is applied over all the areas to match the reflections, so that contrast in the reflections is virtually zero. The technique also eliminates short circuits caused by 'dc plating', since the coating avoids migration between

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electrodes. Ginsbury (UK) Ltd. Tel., 01634 290903; fax, 01634 290904.

Touch screens. Lucas Control announce a range of *Duralith* touch screens, in six sizes to fit most lcds. A reduced layer allows a high degree of light transmission and the screens can be easily modified to take overlays, bezels, shielding and filters for emi/rfi and esd protection. Layers of both analogue or matrix form are available, in matt or semi-gloss surface finish. Lucas Control Systems Products. Tel., 01535 661144; fax, 01535 661174.

5.5in colour lcd. NEC's 5.5in colour lcd operates over the –30°C to 85°C temperature range and is meant for use in vehicles and in industry for navigation displays and process monitors. Drive circuitry for the lcd and for rgb processing is built-in and an edge light gives luminance of 300cd/m^2. Resolution is 320/240 pixels. NEC Electronics (UK) Ltd. Tel., 01908 691133; fax, 01908 670290.

Hardware

Rack with keyboard tray. To avoid trailing leads and the difficulty of seeing the screen on an enclosed computer when the keyboard is remote, Optima has modified its *600* sloped rack to take a keyboard tray. It has a sunken area to take the keyboard, whose leads go through a hole at the back, the operator having a clear view of the screen through the panel of the 10U monitor section, which is sloped at 20°. Optima Enclosures Ltd. Tel., 01875 610747; fax, 01875 612486.

Instrumentation

Power analyser. Voltech's *PM100* power analyser measures V, I, power, VA, reactive power, power factor, peak current, crest factor current and voltage in waveforms from 0Hz to 250kHz. It also handles frequency and inrush current and harmonic analysis to the 50th. A feature is the fact that V and I are directly connected to floating inputs, eliminating external current transformers. Thurlby Thandar Instruments Ltd. Tel., 01480 412451; fax, 01480 450409.

Coating thickness meter. Tor Technologies offers the *PosiTector 100*, which is a hand-held coating thickness gauge that uses ultrasonics to measure coatings on concrete, lacquer, varnish, paint, ceramic glazing, etc. Operation is by two buttons, since there is no need to select coating or substrate and the instrument is auto-calibrating, also providing audible and visual alarms and switched Imperial/metric readings. Optional statistical and memory capabilities allow the storage of up to 1500 readings in 99 groups, with maxima, minima, tolerances, average and standard deviation appearing on the display. There is an RS232 port and infrared connection to the HP-IR printer. Tor Applied Technologies Ltd. Tel., 01455 844114; fax, 01455 844116.

Analogue/digital oscilloscope. *OX8620/8627* from Metrix are combined 100MHz analogue and 40Msample/s digital instruments. Front panels are familiar, since the microprocessor interface allows the use of rotary knobs and touch buttons with leds to confirm selection. Auto-setup is provided, setups being stored at switch-off. They are both dual-channel, dual-timebase instruments, capturing and recording up to four waveforms , each having 8000 points; *OX8627* also has glitch capture down to 50ns. Stored data can be processed in various ways in both analogue and digital modes, and there is RS232 (*OX8620*) and IEEE 488-2 (*OX8627*) communication for control or output. Metrix UK Ltd. Tel., 01256 311877; fax, 01256 23659.

Tv aerial measuring receiver. *ME 900* is an addition to Grundig's range of measuring receivers and is provided with a 5.8in colour display. It is designed for use in the 44.75-859.25MHz range, handling all well-known television standards and having a 200-programme memory. Operation is by soft key, or the analyser function can be cursor-driven, when frequency-related values from the range analyser are read and compared from the colour display. A clock and video decoder are provided. A satellite section covers 910-2050MHz. Grundig AG. Tel., 0049 911/703 86 29; fax, 0049 911/70 96 87.

Waveform analyser. Nicolet's new *2580* waveform analyser is a complete system for multichannel transient recording and analysis; it has a 486 processor, a large colour display and Windows-based software. It takes the form of a mainframe with plug-ins for up to 24 channels with a number of input capabilities. Sampling rates are up to 10Msample/s and memory length up to 8Msample/channel. In the X direction, there is split timebase and several trigger modes to cope with complicated waveforms and a high-speed data transfer system allows the display of waveforms within fractions of a second after acquisition. Complete experiments, including the use of external programs, are supported. Nicolet Technologies Ltd. Tel., 01908 679903; fax, 01908 677331.

Interfaces

PCMCIA Interface. CardWize announce the *Card Genie*, an interface unit connected by cable to a serial or parallel PC port to allow the use of flash or sram PCMCIA cards with a desk-top PC. It behaves as a standard dos drive, accepting the normal commands, and provides simple data swapping between portables and desk-tops, or data logging, in which it can replace printers in some applications. Supplied software allows direct access to the card from the dos prompt, which allows non-dos data to be read, a range of C library files eases the incorporation of the interface into embedded applications. CardWize Data Solutions. Tel., 01635 524477; fax, 01635 524488.

Eight-channel measurement. For connecting up to eight sensors of various types to a computer, IMS produces the *ADAM-4017* interface module, which can be supervised by the host computer via an RS-232/485 link. It has an on-board processor to monitor sensors and either report readings or interrupt under alarm conditions; the eight differential analogue inputs go to a 16-bit a-to-d converter. Integrated Measurement Systems Ltd. Tel., 01703 771143; fax, 01703 704301.

Literature

Network testing. Bisset offers the *Electrodata* catalogue of test and support equipment for data networks, which describes hand-held instruments such as line monitors, digital transmission testers, TI and EI analysers and interface test sets. Prominent is the E1-Watcher, a low-cost tester that monitors 2Mb/s data links used for Megastream, primary rate ISDN and 12-channel pcm. Bisset Communications Ltd. Tel., 01582 792637; fax, 01582 792648.

Inertia sensors. Inertia/impact sensors using reed-switch techniques are described in a publication from Gentech. Inrush current of the sensor is 5A at 20Vdc for 15ms and sensing range is up to 15g. They are board-mounted devices and contacts are in Form A, B or C. Gentech International Ltd. Tel., 01465 713581; fax, 01465 714974.

Filters

Inlet filters. Capable of being fitted to existing equipment, BLP's *SF1020* inlet filter is a general-purpose type meeting IEC 320 and measuring 39.4mm in length. Space is further saved by placing the connectors on top of the body, so that the filters occupy about the same area as a non-filtered type. Versions available handle 1A, 3A or 6A at up to 250Vac. BLP Components Ltd. Tel., 01638 665161; fax, 01638 660718.

Electromechanical components. Rendar has published its 1995/6 catalogue of connectors, cordsets, fans, switches, etc. New this time are 10A IEC320 power inlets, an extended Stripbloc range and a chip fan for Pentium processors. Rendar Ltd. Tel., 01243 866741; fax, 01243 841486.

Materials

Degreaser. From MMCC(UK) comes *Biosane T212*, which is a fast, cold-process degreasing agent that is environmentally and biologically safe. Its surface tension is 50% lower than that of 1.1.1 Trichlorethane and it dries quickly at room temperature. A further advantage is its high settling power, which means that debris sinks to the bottom of the tank so that components do not have to go through it, eliminating the vapour rinse normally needed. It also possesses a 99% recycling capability. MMCC(UK) Ltd. Tel., 01707 336282; fax, 01707 336290.

Metal-substrate pcbs. *IMS* is a printed board on an insulated metal substrate, produced by the French company CIRE. The material dissipates heat and provides insulation between circuitry and metal, consisting of a single-sided pcb with aluminium (or copper), thermally conductive electrical insulator, circuit copper and a solder mask, the aluminium being open to the air for cooling. Double-sided boards can be produced. Insulation depends on the material used and its thickness, but it can be up to 4.8kV. CIRE/BREE. Tel., 0033 38 30 53 62; fax, 0033 38 30 21 30.

Production equipment

Solder paste inspection. Automatic solder paste inspection system, *DEK Inspector 2* by Dek Printing Machines, has a set of new features to improve its versatility without adversely affecting speed. It lifts boards off the belt-feed transport rails, other boards continuing to travel forwards while the selected one is inspected by a combination of pattern recognition and laser scanning, as many areas of the board being inspected as possible until it is needed further along the line. It then replaces the board, picks another up and continues with next Inspection area, adding to the original inspection database in each case. Programming is effected by means of a graphical/menu-driven screen and a library of patterns, the user setting the areas to be examined and tolerances to observe. Flash lighting freezes residual motion and special lighting differentiates between paste and pads, a laser measuring paste height. Dek Printing Machines Ltd. Tel., 01305 760760; fax, 01305 760123.

Air nozzles. New company *Meech-ARTX* is to manufacture air nozzles for swarf removal or cooling air, a new design saving up to 90% of the compressed air normally used and providing an air flow 25 times that given by a normal design. There is also a range of compressed-air powered vortex tubes to handle heat loads of 10000Btu/h and spot refrigeration to –40°, hot air exhausting from one end and cold air at the other. Meech-ARTX Ltd. Tel., 01993 706700; fax, 01993 776977.

Mighty mouse. Interlink Electronics announces that *DuraPoint*, a fully sealed, stainless steel, resistive pointing device for desk-top use, is now available in Europe. It is impervious to fluids, other contamination, vibration and human beings, has no moving parts or mechanical assemblies and therefore does not become clogged up with filth from the mouse mat. It uses the company's *VersaPoint* technique, in which a touch on a button or joystick is transferred to a force-sensitive resistor. Consequently, direction of cursor movement is determined by direction of the applied force on the button and speed of cursor movement by the amount of force applied. Steadlands International Marketing, tel 01670 528200, fax 528212.

Power supplies

High-voltage. Farnell Hivolt has the first of the *PSM10 Series* of high-voltage supplies for applications needing a precise source. Current and voltage monitoring is provided and there is an output inhibit. Outputs are 2, 5, 10 and 15kV and the *PSM10/202* and */153* provide positive or negative output of 10V-2kV at 5mA and 150V-15kV at 666μA. Ripple is less than 20mVpk-pk at 2kV and under 1.5Vpk-pk at 15kV, with drift at under 50ppm/°C. For a ±2V change in the 24V input, regulation is 0.1V at 2kV and 0.75V at 15kV. Farnell Hivolt Ltd. Tel., 01234 841888; fax, 01234 824698.

S-m regulator. With the aid of only four external components, the Sanken *SAI-01* surface-mounted power regulator ic forms a complete switched-mode power supply. It gives an output current of 0.5A at 5V from 7-33V, with 80mV stabilisation and 30mV regulation. Switching frequency is 60kHz. Allegro MicroSystems Inc. Tel., 01932 253355; fax, 01932 246622.

High-current switcher. Having an output rated at 1.25A and switching at 40kHz, Cherry's *CS-3972* is meant for use in both computers and car systems, in which it serves as the pulse-width control element in any of the standard forms of switching regulator. Input range is 3-60V and output up to 60V at 1.25A at a 50% duty cycle. Undervoltage, thermal and current-limit protection is included. Cherry Semiconductor. Tel., 001 401 885-3600; fax, 001 401 885-5786. Internet info@cherry-semi.com.

Frugal regulator. *ZR78L05* is a 5V voltage regulator having a quiescent current of 350μA, which is about one-fifth that of standard 78L types, also providing regulation and stabilisation to 200mA, an increase of 100% over standard. No external components are needed and the device is packaged in SOT23 or low-profile TO92. Thermal overload shuts it down, operation continuing after it has cooled. Zetex plc. Tel., 0161-627 5105; fax, 0161-627 5467.

Radio communications products

Radio-controlled clock. *Model RM913* alarm clock from Oregon Scientific is radio controlled by the 60kHz signal from MSF Rugby, which itself is referred to a caesium standard at the National Physical Laboratory. Data carried by the transmission automatically sets time and date and copes with 30/31-day months, summer time and leap years. The alarm brings on a backlight to the lcd and a separate alarm is useful for reminders during the day and there is a second time zone. The clock is battery-powered. Oregon Scientific. Tel., 0181 903 2886; fax, 0181 903 2328.

Satellite terminal. Magnavox has added to its range of satellite systems the *MX4042*, a desk-top Inmarsat terminal providing voice, 2400b/s fax and data. It consists of a remote antenna (up to 250ft of cable) and two-wire dtmf desk set with handset, signal strength bar graph, keypad and large lcd. Installed software helps to find the satellite by means of a map display on the lcd and synthesised voice prompts operation in English and six other languages. Magnavox Electronic Systems Co. Tel., 001 310 618 1200.

Transducers and sensors

Annunciators. *Appello* programmable voice annunciators have been improved to provide better sound reproduction, more facilities and easier message creation and storage. Three versions give 16s messages, two of 8s with alarm tones and four of 64s in total. Options such as pause length and number of repetitions are defined by the user and an external microphone input is provided for improved recording of speech. Analogue inputs are stored directly in non-volatile memory, with no a-to-d conversion, a technique that provides better quality and wider bandwidth than often found in conventional equipment, as well as being cheaper and quicker. European Safety Systems Ltd. Tel., 0181 743 8880; fax, 0181 740 4200.

Optical sensor. Omron claims its *EE-SX1101* slotted optosensor to be the smallest transmissive type available, being contained in a housing measuring 4.3/4/5mm, with a 2mm wide slot. For increased sensitivity and reliability, a Fresnel lens focuses more light at the detector, which is either a phototransistor or photo-ic for faster switching. The ic type has a photo diode, amplifier, voltage regulator, Schmitt trigger and n-p-n output transistor. Both through-hole and surface-mounted versions are produced. Omron Electronics Ltd. Tel., 0181 450 4646; fax, 0181 450 8087.

Switches and relays

Modular switches. Designed for mounting on pc boards, the Secme *Cosmos* switches measure 12.3mm square, leads being on a 2.54mm pitch for side-by-side mounting. Covers are removable and come in a range of colours and with numerous types of legend. These switches withstand wave soldering and immersion cleaning, and the bases are fully sealed. EAO-Highland Electronics Ltd. Tel., 01444 236000; fax, 01444 236641.

Current sensor. The Lem-Heme *LA125-P* current sensor is for nominal currents of 125mA and is pcb-mounted, providing 3kV isolation while measuring direct, alternating or complex currents in the 0 to ±200A range to within 0.6% of nominal. Bandwidth is 0-100kHz and the device will follow waveforms changing at 200A/μs. The primary current-carrying conductor is passed through a 17mm by 11mm hole in the housing. Lem Heme Ltd. Tel., 01695 20535; fax, 01695 50279.

Tilt sensors. Absolute-encoded tilt inclinometers in the *A-ID* by Control Transducers give an output of 4096mV in 1mV steps over a 360° arc while turning continuously. Encoders are of the optical absolute type, needing no zero reference or resetting and use a pendulum for gravity reference. Control Transducers. Tel., 01234 217704; fax, 01234 217083.

Vision systems

Video conferencing. Nokia offers an integrated monitor, camera, microphone and speakers for video communication. *MediaStation* includes the *NVC-100* software codec for video and audio telephony and is designed for both local and wide-area

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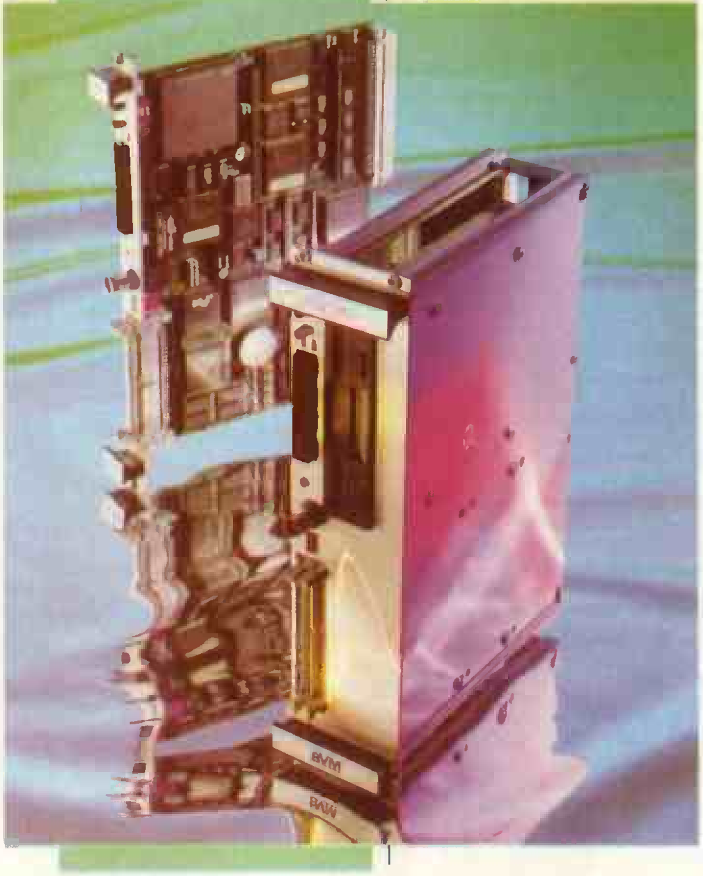
networks. The codec is a pc board with a coding system to ITU-T H.320 standard. Nokia Consumer Electronics AB. Tel., 0046 87938430; fax, 0046 87938441.

COMPUTER

Computer board-level products

110MHz single-board computer. *SE5-110* from Sun SPARC is a new member of the SPARCengine 5 family, this one having the 100MHz version of the MicroSPARC II processor, running at 140Dhrystone 1.1 mips for networking, telecomms and imaging. Interfaces for fast SCSI, floppies and 10MB/s Ethernet, serial and parallel data are built-in. A local 64-bit bus allows fast data transfer from memory to video for graphics, effectively a fourfold increase in data transfer speeds. Sun Microsystems Ltd. Tel., 01276 451440; fax, 01276 451287.

68060/40 modules. BVM announces the *BVME6000/4000* self-contained modules based on 68060/40 single-board computers. These combine the computer boards with up to 16Mbyte of flash and 32Mbyte of dram, a 3.5in 2 or 4Mbyte floppy drive and SCSI drives from 260Mbyte to 4Gbyte, and 512K of 32-bit non-volatile sram – all in a 6U by 12HP module, only power being needed from the VMEbus backplane. Two IndustryPack sites provide i/o and two serial ports with field-changeable buffer modules support RS-232, RS-422 or RS-485, and 8-bit Centronics parallel port also being included. BVM Ltd. Tel., 01489 783589; fax, 01489 780144.



Computers

Industrial computer. AMC's *ErgoTouch Computer HL* is certified to Class 1 Division 2 and rated to NEMA 4X on all surfaces and needs no air supply or purging/pressurisation enclosure; it is a complete PC contained in a 4.5in-deep housing, has a touch panel and can be mounted anywhere – wall, panel, boom or desk. Advanced Modular Computers Ltd. Tel., 01753 580660; fax, 01753 580653.

Data communications

Data/fax modems. Rockwell has the *SMV288ACW, V.34*, a new member of the *SocketModem* family of oem units. Performance is increased to a data handling throughput of 115kb/s for data and 14.4kb/s for fax send and receive. Voice mode options include adpcm companding and efficient voice and sound-bite storage; the units fit into sockets. Features include automatic line-speed selection in V.34 and V.32/V.32bis, tone, pulse and adaptive dialling and full diagnostics. Typical power taken is 885mW from 5V. Telecom Design Communications Ltd. Tel., 01256 332800; fax, 01256 332810.

Radio link. *Link-up* is a synthesised uhf transceiver module approved to the licence-free MPT1329, delivering 0.5W in the 458.5-458.95MHz band, carrying data at 9600baud. Its control processor allows simple interfacing to third-party equipment and to a more powerful asynchronous serial digital comms port to handle channel changing, status reporting and mode changes. It measures 93/60/17mm and is suitable for mounting on mother boards in oem applications. A&R Electronic Developments Ltd. Tel., 01889 574980; fax, 01889 574975.

Data logging

Quality-control. *C9000* by the French company Ati-Topquali is a data logger for work in quality control, working with manufacturers' own machines. It acquires all types of data and can measure dimensions, electrical variables and physical quantities such as weight, speed, movement, etc., as well as providing good/bad Indication. The loggers, which consist of the logger and monitor units, are modular, having between 4 and 24 channels connected to inductive sensors and can be equipped with digital sensors to give up to 64 data channels. *Gesmes VA04* software is provided for management and statistical data processing. Ati-Topquali. Tel., 0033 35804199; fax, 0033 35804591.

Mass storage systems

Flash eproms. Flash eprom 'disks' from M-Systems come in a wide variety, from 1 to 900Mbyte and in formats including PBbus and PC/104 cards, PCMCIA modules and SCSI-II drives. All have 'plug-and-go' software and file-management compatibility with the pc to emulate hard disks and other media. Allbite Technology Ltd. Tel., 01604 491717; fax, 01604 491323.

32Mbyte dram cards. Mitsubishi has 4, 8, 16 and 32Mbye dram *MelCard* memory card for upgrades in Toshiba, IBM, Compaq and other notebook computers. They use 930mW, work from 3.3V or 5V supplies, have an access time of 70 or 80ns and are JEIDA/JEDEC-compatible. There is protection against incorrect supply voltage and incorrect insertion. Mitsubishi Electric UK Ltd. Tel., 01707 276100; fax, 01707 278692.

Programming hardware

Gang programming simms. From MQP Electronics announces an increase in the number of devices supported by its *S2200* and *S2400* range of production programmers, a full range of microcontrollers from Motorola and SGS-Thomson now being handled, typically eight at a time. Now, for the first time, gang programming of flash simm and PCMCIA modules, up to 128Mb, is offered. Provision of a universal programming module in the S2000 means that this one programmer will meet all requirements. MQP Electronics. tel., 01666 825146; fax, 01666 825141.

Philips's XA programming. Data I/O announces programming support for the *XA* 16-bit controller by Philips, including the *2900* and *3900* programming systems – *UniSite* universal programmer and *ProMaster* automated handling system. The programming systems and universal programmer have sockets to take virtually every device package currently on the market. Philips Semiconductors (Eindhoven). Tel., 00 31 40 722091; fax, 00 31 40 724825.

Software

Raster-to-vector translator. *I/Vector 3.2* is a fast 32-bit translator that converts scanned raster images to vector .DXF files with no need for a cad operator, recognising circles, arcs, symbols, ocr text at any angle, line styles, hatching, arrows and area outlines, all functions being user-controlled. Translated results can be viewed overlaid on the original raster display for verification. Included is an editor to clean, up-date, de-skew and de-speckle the image and a region editor assigns different parameters to different areas of drawings, so that redundant features such as borders may be omitted. A layer manager sorts features into appropriate layers. Raster formats accepted Include RLC, TIFF Group 4, RLE and PCX and the program runs under Windows 3.3, 95, NT or UNIX/Motif. Ideal Scanners. Tel., 001 301 468 0123; fax, 001 301 2300813; Internet http:/www.ideal.com.

Visual design. Integrated Systems announces that *Matrix*$_X$, a family of visual design and development tools, is to be available for Windows 95. *Matrix*$_X$ is an integrated hardware/software system allowing engineers to design dynamic systems, test them and bring them to prototype stage, code and documentation being generated automatically. Included in this family is *Xmath*, the object-oriented mathematical analysis and visualisation tool and *AutoCode*, an automatic code generator. There are also *DocumentIt* for automatic documentation and prototyping and *RealSim* prototyping computers. Integrated Systems Inc. Ltd. Tel., 01438 751651; fax, 01438 312311.

FFT analyser. Using any Windows-compatible sound card, *Spectra Plus 3.0* by Strategic Test provides fast Fourier analysis, Its features including FFT to 16K block size, 1/3octave measurement with flat/A/B/C weighting, thd and transfer function display and amplitude calibration and microphone compensation. Digital filtering is available as a post-processing feature. Display options include time series, phase, spectrum, spectrogram and 3D surface plot. Sampling rate supported is a maximum of 44.1kHz. Strategic Test and Measurement Systems Ltd. Tel., 01734 795950; fax, 01734 795951.

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CIRCLE NO. 144 ON REPLY CARD

Do you know *Foster-Seeley?*

Amazed at the widespread misunderstanding of how the Foster-Seeley discriminator works, Richard Brice sets the record straight.

Despite the increasing use of phase-locked-loop fm detectors, the Foster-Seeley discriminator and its cousin the ratio detector still enjoy widespread use.

Sadly, familiarity has bred contempt for the principles at work in this apparently simple circuit, top right. It is still common to see descriptions of the operation of the discriminator which refer – without qualification – to the 90° phase-shift which exists between the primary and secondary of a tuned transformer'.

Yet such a phase shift only occurs under special circumstances. As a student, I found explanations of the circuit confusing. Rather than challenge the principles given, I opted to believe that a phase shift must exist between the primary and secondary voltage of all tuned transformers.

This fallacious belief was finally laid to rest when some empirical investigations proved this was not the case. A little research has uncovered that this confusing explanation of the Foster-Seeley discriminator has a long and distinguished history.

"...operation is dependent on the 90° phase shift which occurs at resonance between primary and secondary voltages of a tuned transformer."[1]

"It will be seen that both primary and secondary windings of the IF transformer are tuned. Now, when they are both tuned to resonate to the incoming signal (i.e., the normal carrier frequency) the voltage across the secondary, V_s, lags the voltage across the primary, V_p, by 90°."[2]

"Conventional discriminators use the composition of two vectors at right angles when there is no frequency deviation and when the frequency departs from its state of rest the angle between the vectors departs from 90°. Figure 2.6 [similar to diagram top right, next page] *shows an elementary balanced discriminator. Capacitors C tune the inductances to resonate at the centre frequency of the intermediate frequency band of the receiver... At resonance the voltage V1 [V_p] will be in quadrature with the voltages in each half of the secondary winding."*[3]

Foster-Seeley discriminator analysis

$$i_{A-C} = V_o\left(\frac{1}{R} + j\omega C\right)$$

$$V_{A-B} = V_o\left(1 + j\omega L\frac{(1+k)}{R} - \omega^2 L(1+k)C\right)$$

$$i_{A-B} = V_o\left(\frac{j}{\omega kL} - \frac{1+k}{kR} - j\omega C\frac{(1+k)}{k}\right)$$

$$V_{in} = V_{A-B} + j\omega L(1+k)(i_{A-C} + i_{A-B})$$

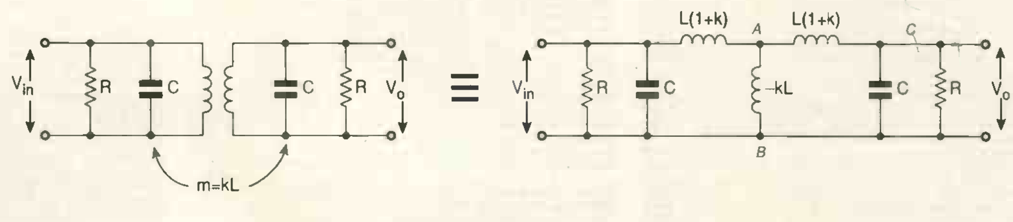
$$= V_o\left(\begin{array}{l} 1 + j\omega L\dfrac{(1+k)}{R} - \omega^2 L(1+k)C \\ + j\omega L\dfrac{1+k}{R} - \omega^2 L(1+k)C - \dfrac{1+k}{k} \\ - j\omega L\dfrac{(1+k)^2}{kR} + \omega^2 LC\dfrac{(1+k)^2}{k} \end{array}\right)$$

$$\frac{V_o}{V_{in}} = a\angle\theta$$

$$\mathrm{Tan}\theta = -\frac{\dfrac{\omega L}{R}\left(2 + 2k - \dfrac{1}{k} - k - 2\right)}{\omega^2 LC\left(\dfrac{1}{k} + k + 2 - 2 - 2k\right) - \left(\dfrac{1}{k} + 1\right)}$$

$$= \frac{Q\left(\dfrac{1}{k} - k\right)}{\omega^2 LC\left(\dfrac{1}{k} - k\right) - \left(\dfrac{1}{k} + 1\right)}, \text{ which if } \omega^2 LC = 1,$$

$$\mathrm{Tan}\theta = \frac{Q\left(k - \dfrac{1}{k}\right)}{k+1} \text{ and} \rightarrow \infty \text{ as k} \rightarrow \phi$$



Action of the discriminator relies on very loose coupling between primary and secondary windings of the IF transformer.

But, how can any phase-shift exist between the primary and secondary of a transformer if the flux closely couples both coils? The answer is – in the Foster-Seeley circuit it does not.

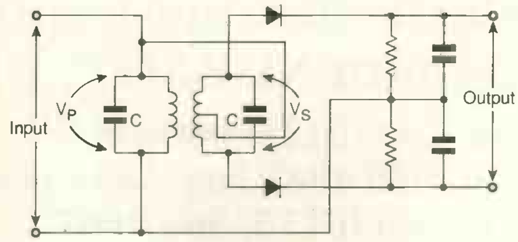
Circuit action relies on there being a very loose coupling between primary and secondary windings of the intermediate-frequency transformer. The diagram bottom left, and the associated working presents a more rigorous analysis[4]. Analysis also shows that when coupling between a transformer primary and secondary is perfect – equal to 1 – no phase shift exists between primary and secondary voltages. This is regardless of whether the windings resonate with parallel capacitances or not.

M. G. Scroggie in *Foundations of Wireless* appears to be one of the few writers who deemed it necessary to bother the reader with this crucial aspect of the discriminator's operation,

"*...because both windings are tuned exactly to the carrier wave and coupled only very loosely, voltages across them are 90° out of phase.*"

Some authors, perhaps sensing murky waters but being unsure as how to clear them, simply eschew explanation altogether. A recent, and widely set textbook[6] describes the Foster-Seeley discriminator thus,

"*The Foster-Seely (sic) detector, or its variant, the 'ratio detector' [uses] a single tuned circuit in a fiendishly clever diode arrangement to give a linear curve of amplitude output versus frequency over the IF bandpass.*"



Foster-Seeley discriminator – it is still common to see descriptions of this referring to a 90° phase-shift between the primary and secondary of the tuned transformer.

This explanation is unlikely to leave today's students crystal clear as to the operation of the circuit. Its author appears to use words as did Lewis Carroll's Humpty Dumpty who argued, "*when I use a word... it means just what I choose it to mean – neither more nor less*"[7] ■

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3. Maurice R.D.A., 'Television Engineering', Pentech Press, London, 1985.
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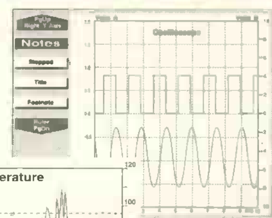
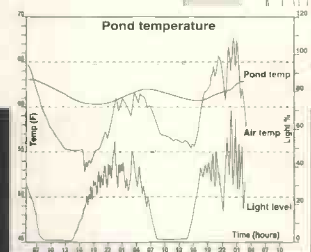
5. Scroggie, M.G., 'Foundations of Wireless', 8th edn, Butterworth & Co., London, 1971.
6. Horowitz, P. and Hill, W. 'The Art of Electronics', 1st edn, Cambridge University Press, Cambridge, 1980.
7. Carroll, L., 'Through the Looking Glass'.





Analysing AC

Owen Bishop looks at how circuit simulators handle various aspects of ac analysis – including noise and poles & zeros.

Analysis of ac is mainly concerned with the effects of frequency on the operation of a circuit. This article looks at the way that the *Micro-Cap IV* simulator handles analyses in the ac group.

Operating under dos, *Micro-Cap IV* is a well-established implementation of Spice2. It accepts netlists in conventional Spice format, but it is also able to work from a schematic using a very comprehensive schematic editor. Nodes are numbered automatically – they can also be given names if preferred – with the ground node being node zero. In *Micro-Cap IV* or *MC4*, wires that cross are taken to be connected, so jumpers are used where crossing lines are not joined.

Analysing op-amps

Figure 1 shows a low-pass active filter with Chebyshev response. Component values were calculated on paper using the conventional capacitor-ratio tables for a 10kHz Chebyshev filter with 1dB ripple in the pass-band.

The purpose of the simulation is to confirm the behaviour of the filter, and to investigate the effects of using different types of op-amp. The user has the option of allocating each component a name, or simply specifying its value.

In Fig. 1, most components are specified by their values, but the voltage source and op-amp are given names. The MODEL statements for these appear below the schematic. If necessary, these can be edited when parameters need to be changed.

Parameters in the sine-wave voltage source specify – in order – a frequency of 1MHz, amplitude of 1V, dc offset zero, and phase shift zero. Note that Spice does not distinguish cases. Both ‘m’ and ‘M’ mean ‘milli’. For ‘mega’, Spice uses ‘meg’ or ‘MEG’. Other parameters may be specified but their default values are satisfactory in this example.

Note that parameter symbols used here are not standard Spice symbols for these parameters. The operational amplifier model is specified by its model name, LF355, model type, OPA, and some relevant parameters. The term LEVEL=3 indicates that the most complex op-amp model should be used while TYPE 3 specifies n-channel junction-fet inputs.

Other parameters specified are open loop gain, 2×10^5, positive and negative slew rates, 5V/µs, and gain bandwidth, 2.5MHz.

Performing ac analysis

The netlist also issues commands for the ac analysis. The second term in the .AC command determines frequency points for which the results are calculated.

There are three options. The term DEC stands for ‘decade’ meaning the frequency range is to be divided into decades. The total range is 100Hz to 1MHz divided into four decades – 100Hz to 1kHz, 1kHz to 10kHz, 10kHz to 100kHz and 100kHz to 1MHz. The number of frequencies within each decade is specified by the number following DEC, in this case 20.

These points are automatically located on a logarithmic scale. Instead of DEC, I could have stipulated LIN to

Spice commands

These Spice statements are available for ac analysis.

.AC initiates an ac analysis, with an ac frequency sweep. Small-signal response of the circuit is calculated for each point on a specified range of frequencies, assuming that the circuit is linear.

Spice begins the analysis by computing the dc operating point – refer to ‘Analysing dc via Spice,’ *EW+WW* Oct 1995 – and uses the results of this to provide the voltage levels with reference to which all non-linear components are linearised.

.NOISE calculates noise response for input and output.

.PZ performs a pole-zero analysis – available only in Spice3.

.DISTO analyses distortion caused by non-linearities in semiconductor devices.

.PRINT AC calls for the results of the analysis to be printed as a set of tables – not in Spice3.

.PLOT AC calls for the results of the analysis to be plotted as a graph – not in Spice3.

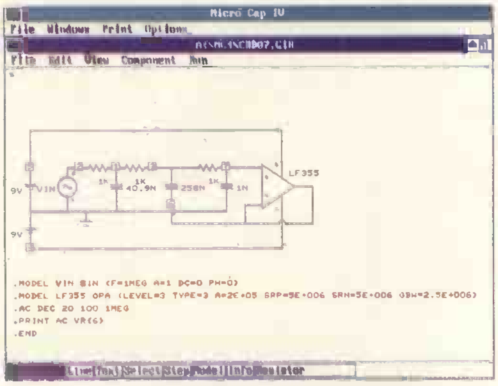


Fig. 1. Typical schematic drawn with Micro-Cap IV includes definitions of models and optional command statements for the analysis.

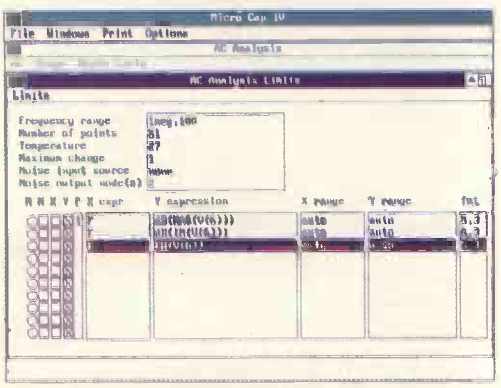


Fig. 2. The ac analysis dialogue box allows analyses to proceed interactively.

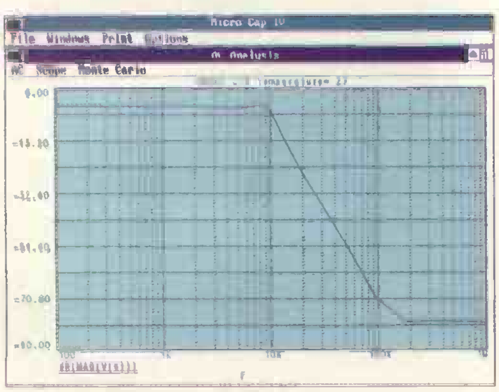


Fig. 3. AC response of the circuit of Fig. 1 shows the typical response of Chebyshev low-pass filter.

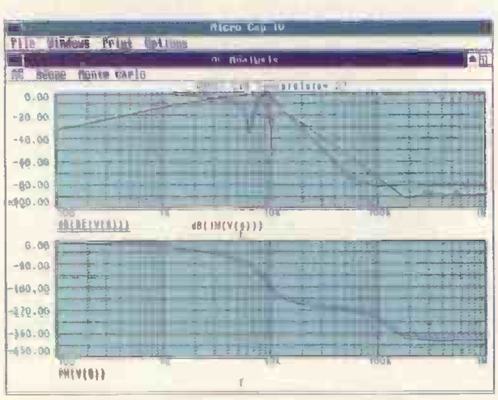


Fig. 4. The upper graph shows real and imaginary components of the output of the filter of Fig. 1. The lower graph shows phase response.

Netlists in Micro-Cap IV

Micro-Cap IV is able to use a standard Spice netlist, which may be typed on the Editing screen, then saved as a .CKT file. Alternatively, it will import a Spice netlist from another simulator or typed on a word-processor, provided that it is saved as an ascii file without special headers.

The most convenient technique is to draw the schematic, add .MODEL and other statements, then work with this combined schematic-netlist, using the interactive Limits boxes, of which there is one each for dc, ac, Fig. 2, and transient analyses.

On request, MC4 prints out a netlist but this is not a Spice netlist and can not be used to run a simulation. It is a compact statement of the circuit and parameters, excluding command statements.

Under the statement line .OPTIONS, it lists the values of a number of parameters currently being used by Spice. These include GMIN, the minimum conductance value used in calculations, and RELTOL, a tolerance level used to determine when Spice has performed sufficient iterations.

These and other parameters have default values which it is not normally necessary to alter.

have the frequency range divided on a linear scale. Spice also has a OCT option, not available in *MC4*, in which the range is divided into octaves, the end of each sub-range having twice the frequency of its beginning.

Command statements shown on the netlist are optional, as *MC4* has an 'ac analysis limits' dialogue box which allows the analysis to be set up and modified interactively. If command statements have been entered on the netlist, they appear in the dialogue box and can be used as the basis for an analysis.

Otherwise, and more usually, analyses are controlled from the box, **Fig. 2**. To reach this from the schematic screen, select the Run menu, then item 2:AC Analysis. The upper panel holds run parameters, beginning with the frequency range and listing the final and starting frequencies. The default number of steps is 51, giving 50 equal intervals throughout the range, and overrides the Spice command given on the netlist.

The lower part of the box controls graphical output. Enter details of the curve or curves to be plotted. Checking in column N gives a numeric printout. Its format – number of places before and after the decimal point – is set by the numbers in the 'fmt' column. The X and Y columns specify logarithmic scales if checked or linear scales if unchecked. Expression X is usually frequency, F, in ac analyses. Expression Y has a variety of forms.

In Fig. 2, the first line states that the y-variable is to be the magnitude, MAG, of the voltage, V, at node 6, plotted on a decibel scale. The second line calls for the imaginary component of the voltage at node 6, plotted on a decibel scale. The third line asks for the phase of the voltage at node 6. Click on 'Limits' and select 1:Default all, to enter 'auto' in the X range and Y range columns. The software now calculates suitable limits for the graphs.

To plot only the magnitude-frequency graph, select the table's first row by typing '1' in the plot column, P, Fig. 2. Click on the square in the Limit box top left corner and select 6:Close. The ac analysis screen is revealed. Click on AC and select 1:Run. This gives **Fig. 3**.

The magnitude scale is a little difficult to interpret. This is because the second grid line down is at –3.6dB, but the pass-band shows an output of 0dB from 100Hz up to about 9kHz. There is one large ripple between about 1.5kHz and 7kHz. The –3dB frequency is 10kHz, as required. Roll-off is from –3dB at 10kHz to about –66dB at 80kHz, or about –21dB/octave.

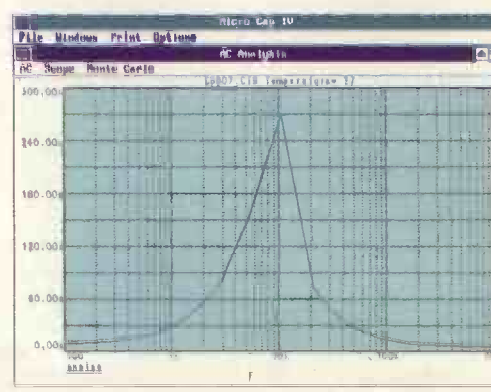


Fig. 5. Output noise analysis of the filter of Fig 1 shows a strong peak at 10kHz.

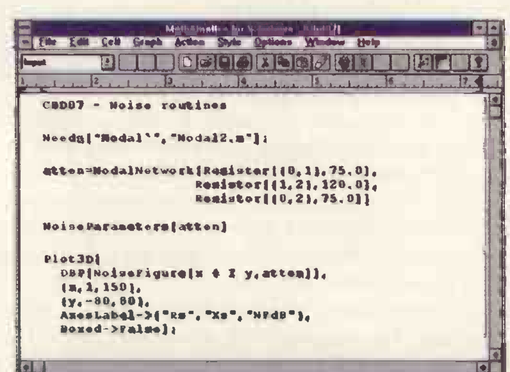


Fig. 6. Nodal routines are used for calculating noise parameters of circuits such as the attenuator network of Fig. 7.

Figure 4 shows the result of returning to the 'analysis limits box', and making a few amendments. Edit the Y expression in the first line to dB(RE(V(6))), to plot only the real component of the output voltage. Type '1' into the P column of the first two lines, so that both real and imaginary components are plotted on the same graph. Type '2' in the P column of the third line to obtain a separate graph of the phase. The result is Fig. 4.

There are some interesting changes of amplitude between 6kHz and 10kHz, when the phase lag is rapidly swinging from –90° to –180°. The phase curve follows a 'wavy' line, typical of a Chebyshev filter.

For more precise values than can be read from the graphs, return to the ac analysis Limits box and check the N column of the appropriate graphs. The graphical display is then accompanied by a numerical printout. **Table 1** shows part of the results of printout of the curve of Fig. 3, locating the –3dB point at 9.85kHz, which is reasonably close to the designed value.

The earlier part of the printout shows a local minimum (–0.863dB) at 4.57kHz, and a local maximum (+0.207dB) in the pass-band, confirming the expected 1dB ripple.

Table 1. Part of the printout of the ac analysis locates the –3dB point.

F (kHz)	dB(mag(V(6))) (V)	
9.690	–2.632	
9.770	–2.857	
9.850	–3.088	–3dB at approx 9.85kHz
9.930	–3.322	
10.010	–3.560	
10.170	–4.044	

Table 2. Results of using NoiseParameters to analyse the attenuator circuit of Fig. 7.

Fmin –> 25.
NFmin –> 13.9794
Rn –> 312
Yopt –> 0.02
Gammaopt –> 0

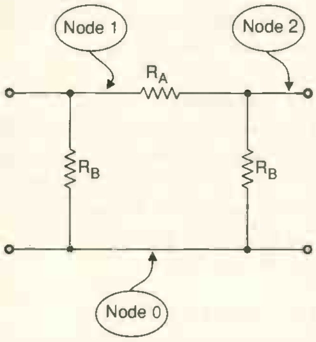


Fig. 7. A π-network is the subject of the noise analysis reported in Table 2.

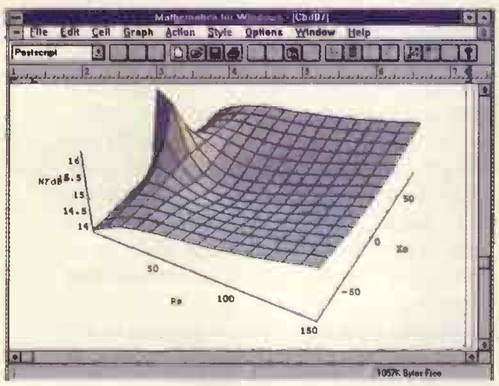


Fig. 8. A three-dimensional plot of the noise figure of the attenuator helps the engineer to see how noise is related to source resistance and impedance.

Component substitutions

Once the schematic/netlist has been prepared, it takes only a few minutes to investigate the effects of substituting different components. Replace the *LF355* with a bjt-input op-amp, the *LM301A*, which costs only half as much. On the schematic, click on 'Select' at the bottom of the screen, then double-click on the MODEL line. A text box opens near the top of the screen, containing the current model definition. Edit this to,

```
.MODEL LM301A OPA (LEVEL=3
TYPE=1 A=16E+04 SRP=5E+005
SRN=5E+005 GBW=1E+006)
```

The new op-amp is type 1, i.e. with a bipolar transistor input. It has lower open-loop gain, a much slower slew rate and a narrower, though still adequate bandwidth. Running ac analysis produces curves of much the same shape as before, but the –3dB point is now at 9.45kHz, and the roll-off above that point is about 22dB/octave. There would appear to be no advantage in using the *LF355*.

The original calculations specified capacitors to a high degree of precision. Edit the schematic to replace the capacitors with E12 values. Replace 40.9nF with 39nF, 258nF with 270nF and leave the 1nF capacitor unchanged. The analysis printout shows the –3dB point at 9.25kHz with roll-off still at 22dB/octave. There is little need to use precision capacitors.

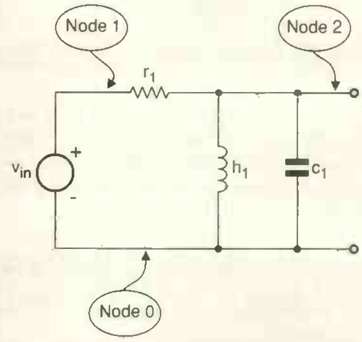


Fig. 9. Poles and zeros of this band-pass filter are analysed by a Nodal routine in Fig. 10.

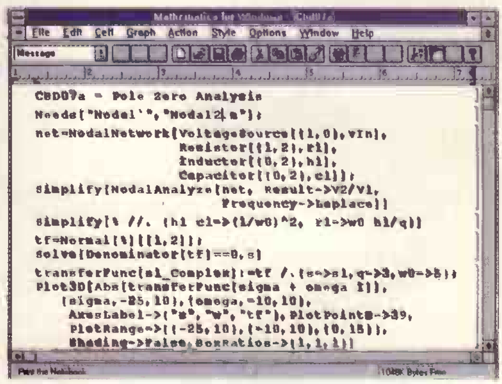


Fig. 10. NodalNetwork function is used to assign the symbol net to the circuit of Fig. 9. Then the net is analysed to locate its poles and zeros.

Spice output variables

Syntax used by *MC4* is not applicable to other Spice simulators. Standard Spice variable names for output voltages at a given node are,

VM Magnitude of complex voltage
VR Real part of complex voltage
VI Imaginary part of complex voltage
VP Phase of complex voltage
VDB Magnitude of complex voltage, in dB.

The corresponding variable for output current begins with I instead of V. Variable name of an output voltage is followed by the node number, in brackets. If two nodes are quoted in brackets, the variable refers to the difference in voltage between the two nodes.

The variable name of an output current is followed in brackets by the name of the voltage source through which the current is to be found. This may be a dummy source – refer to 'Deeper into dc analysis,' *EW+WW* Nov 1995.

Simulating noise

Spice is able to simulate noise generated by resistors and active components. Thermal, shot and flicker noise are represented by appropriate voltage sources in the component models. Their total effect at the circuit output is obtained by using the command line,

```
.ONOISE V(N) Vname n
```

Here, N is the output node number, Vname is the name of the input voltage source (this can also be Iname, for a current source), and n is the number of points in the frequency range for which a noise report is required. Noise reports for output and input noise are called up by the command statement,

```
.PRINT NOISE ONOISE INOISE
```

By default, ONOISE and INOISE are expressed as voltages. But if ONOISE and/or INOISE are followed by (DB), the values are in decibels.

At each point, the analysis lists the thermal noise contributed by each resistor. It also lists noise generated in semiconductor devices – thermal noise from resistances, shot noise from currents across junctions and flicker noise. These noise levels are expressed as mean squares (e^2/Hz). Their total is calculated, and the square root of this is also printed, to give rms values.

Finally, the transfer function of the circuit is evaluated and the equivalent noise at the input, INOISE, is obtained by dividing the rms output noise by the transfer function. All of this is repeated at each frequency point, so a complete noise analysis extends to several pages of printout. A detailed analysis such as this is useful for identifying the major sources of noise, with a view to reducing their effect. For a more overall view, the .PLOT command produces a graphical printout of rms values for each point.

A graphical display is obtained in *MC4* by completing the two 'noise' entries in the 'ac analysis limits' box of Fig. 2. Enter 'VIN' as the noise input source and '6' as the noise output node. In the curve table, enter '1' under P, 'F' under X expression, and 'ONOISE' under Y expression. Select default limits. In the ac analysis window, select Run.

Graph **Fig. 5** shows the noise level peaking at 270nV/√Hz at 10kHz. Run an analysis for INOISE in the same way. This results in a similar peaked curve, reaching its maximum of 85μV at 105kHz.

To investigate the effects of temperature on noise, return to the 'Limits' box and alter the temperature setting – in a regular Spice netlist, this is done by a .TEMP statement. Increasing the temperature from the standard 27°C to 50°C raises the peak of the ONOISE plot from 270nV to 280nV. Decreasing the temperature to 0°C reduces it to 257nV.

Reset temperature to 27°C, then substitute an *LT1028AM* op-amp, described as an ultra-low noise op-amp, for the *LF355* and run an ONOISE analysis. The peak comes at 10kHz as before, but now the noise level is only 160nV/√Hz, confirming that the description of the op-amp is appropriate.

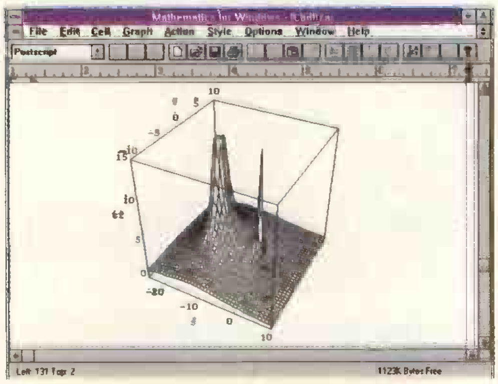


Fig. 11. In a three-dimensional plot such as this, peaks in the surface correspond to poles and depression to zero level corresponds to a zero. Plots of this type are useful for visualising the poles and zeros of a circuit.

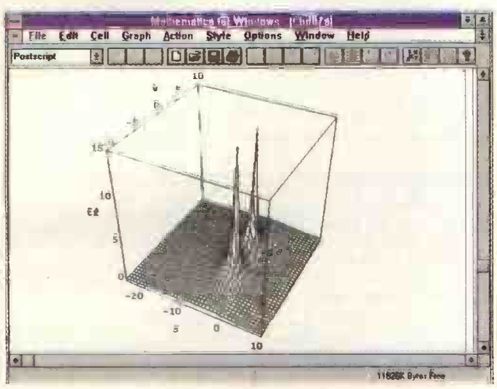


Fig. 12. Changing two parameters produces a plot with two imaginary poles instead of the two real ones of Fig. 11.

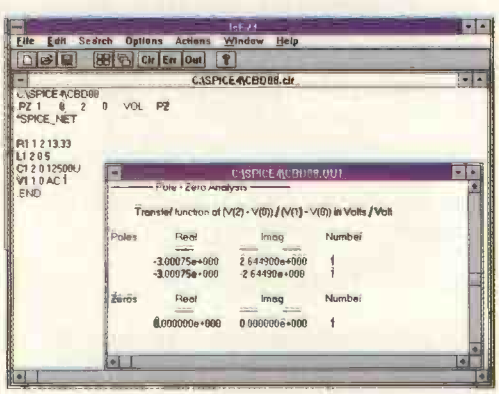


Fig. 13. The netlist of Fig. 9, with the pole-zero analysis seen in the Output file. The circuit has two complex poles, q and ω_0 having been given the values that produced Fig. 12.

Measuring generated noise

A widely used measure of the noise generated in a circuit is noise figure. Spice does not provide a direct calculation of noise figure, but it can be found via *Mathematica*. Last month I demonstrated a routine from the *Electrical Engineering Pack*.

An even more extensive collection of *Mathematica* functions and utilities is published by Macallan Consulting under the title *Nodal*. **Figure 6** shows a *Mathematica* Notebook which demonstrates two of the several Nodal functions associated with noise calculations. Run *Mathematica*, load the notebook then click on the first line followed by the Evaluate button, to load *Nodal*.

One of the most frequently used *Nodal* functions is NodalNetwork. Just as in a Spice netlist, a circuit is described in NodalNetwork by listing its components and the nodes to which it is connected. As in Spice, node 0 is the ground node.

A netlist can be quoted in full in any functions in which it is used or, to save typing and possible errors, it may be assigned to a variable name once and for all. In Fig. 6 a simple attenuator network, **Fig. 7**, with input and output impedance 50Ω and an attenuation factor of 5, is assigned to the variable 'atten'. Using this variable in *Nodal's* NoiseParameters function provides the results listed in **Table 2**.

Parameters listed are Fmin, the minimum noise figure, NFmin, the minimum noise figure in decibels, Rn the network resistance, Yopt the optimum source admittance, and Gammaopt the optimum source conductance.

One of the advantages of *Mathematica* is that it allows three-dimensional plots. This makes it is possible to visualise the effects of two parameters simultaneously. **Figure 8** shows a decibel plot of the noise figure of the attenuator against source resistance and source impedance. The minimum noise figure is found when source resistance is 50Ω and source impedance is zero.

Pole-zero analysis

Pole-zero analysis is available only in Spice3, and certain high-level Spice implementations. At present, of the simulators I am looking at, only the full version of *IsSpice* has this feature. The command statement for the band-pass filter of **Fig. 9** is,

.PZ 0 1 0 2 VOL PZ

The numerals specify input and output nodes respectively, VOL indicates that the input is a voltage – use CUR for a current – and PZ calls for both poles and zeros to be calculated. For poles only use POL, or for zeros only, use ZER.

Nodal provides a routine for visualising the poles and zeros of a transfer function. In **Fig. 10**, the circuit of Fig. 9 is described by the netlist net. In this example, components are not given specific values but symbolic names: *r1*, *c1* and *h1*. The ability of *Mathematica* and *Nodal* to work with symbols as well as, or instead of, numeric values is one of their strengths.

The network is analysed, using the 'NodalAnalyse' function, to find the transfer function, which is the ratio V_2/V_1. The 'Simplify' function simplifies the result of the analysis and it is displayed as,

$$\frac{h1 \times s}{r1 + h1 \times s + c1 \times h1 \times s^2}$$

This formula can be used for calculating the transfer function, given values of *r1*, *c1* and *h1*.

Analysis continues by eliminating the component symbols, using well-known relationships; replace *h1c1* with $1/\omega_0^2$ and *r1* with $\omega_0 h1/q$, where ω_0 is the resonant frequency and q is the quality factor of the filter. After simplifying, the transfer function is,

$$\frac{qs\omega_0}{s^2 + qs\omega_0 + \omega_0^2}$$

Simplification is required after certain calculations because *Mathematica* does not always reduce an expression to its simplest possible form. In the example above, it omits to cancel out *h1* throughout the expression when 'Simplify' is not used. Simplify tells it to try harder.

To find the zeros, look for values of the variables to make the expression equal to zero. Values that make it equal to zero are values which make the numerator equal to zero. Zeros can usually be found by inspecting the expression. Quality factor and ω_0 can not be zero, but the expression evaluates to zero when s is zero. This is one of the zeros of the function. Making s infinitely large also makes the expression equal to zero, so a second zero is at infinity.

To find the poles, look for variable values which make the expression take an infinite value. These are those which make the denominator equal to zero, and the next step in the routine is to find them.

The expression above is actually part of a list, though this is not displayed on the screen. The 'Normal' function picks out the expression and allocates it to symbol *tf*. Solve the equation for s when the denominator is put equal to zero. This gives two values for s. It is obvious that the standard formula for solving quadratic equations has been used,

$$s = \frac{-q\omega_0 \pm \sqrt{(q^2\omega_0^{\,2} - 4\omega_0^{\,2})}}{2}$$

This equation can be used directly to calculate the position of the poles on the s axis, by substituting actual values of q and ω_0. For example, if q is 3 and ω_0 is 5, then s is –1.91 or –13.09. The number of poles equals the number of zeros, as required by theory.

A clearer picture of the distribution of poles and zeros is obtained by plotting the transfer function for a range of values of s and ω in three dimensions. Define another transfer function in which $s1$ is a complex variable, and replace s in the original transfer function.

For the sake of comparison with the numerical example above, make q=3 and ω_0=5. Any other reasonable values could be used, with different results. The three dimensional plot is shown in **Fig. 11**. The shallow depression in the surface at (s=0,ω=0) is one of the zeros. The surface slopes down in all directions around the plot, indicating the other zero at ($s=\pm\infty$, $\omega=\pm\infty$).

The poles show up clearly as two peaks on the negative s-axis, where s=–1.91 and s=–13.09. The appearance of the plot is changed if different values of q and ω_0 are used. In **Fig. 12** the plot with q=1.5 and ω_0=4 shows two complex poles, the zeros being located where they were before.

Nodal calculates the transfer function in symbolic form but all Spice programs require component values to be specified. In **Fig. 13** the *IsSpice* netlist of the filter of Fig. 9, set up for q=1.5 and ω_0=4. It identifies the input between nodes 0 and 1 – the voltage source – and the output between nodes 0 and 2. Both poles and zeros are to be found. In the 'Actions' menu, click on 'Simulate'.

The analysis takes only a fraction of a second, then select 'Edit' in the 'Actions' menu, and click on the OUT button to display the output file. This shows that the circuit has complex poles at the points s=–3+j2.65 and s=–3–j2.26, as has been demonstrated by *Nodal* in Fig. 12. There is one zero, which is at the origin. ■

Reference

Riddle, Alfred, and Dick, Samuel, '*Applied Electronic Engineering with Mathematica*', Addison-Wesley Publishing Company 1995. Includes a demonstration version of Nodal on diskette.

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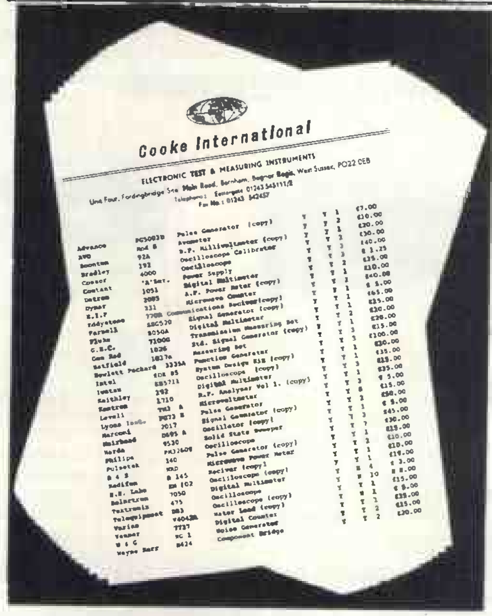

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