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In conjunctions with Allegro distributor Ambar Cascom, we present a technology primer and a **free** samples service for the first 500 readers replying with the special card bound into the journal. Discover the benefits of the latest magnetic transducer technology.
Plus a mass review of instrumentation amplifier chips.

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Right to communicate

One cannot always say that advances in electronics bring universal benefits but one development here in the UK looks more promising than most.

Last month saw the launch of a new mobile telephone service, *Orange*, run by the Hong Kong operator, Hutchison Telecom. This brings the number of mobile phone systems operators in the UK to four.

The new service is causing some consternation among telecoms industry pundits who would claim that four separate systems is at least one too many... They fear that competition will jeopardise payback on the investment required to set these services up. The old but profitable analogue mobile phone networks represent well over £1 billion each. We don't know yet how much the operators will eventually have to spend on the newer networks for acceptably complete coverage but it will be a lot more than this.

The existing analogue services can provide between the two of them room for 6.5 million users. The digital GSM services – *One-2-One* and *Orange* – will have a capacity three or four times that in subscriber numbers. This expansion represents a wireless network capacity which begins to rival the standard public service wired network. Phones and phone calls where you want and when you want, but not yet at a price which matches the public telephone system call charges.

This strikes me as odd. Buried wires are as expensive to maintain as they are inflexible to use. How much investment does the existing wired network represent? One can only guess. £50 billion does not seem like an unreasonable estimate with a maintenance cost to match. Yet land line call and rental charges are cheap in comparison to the mobile subscriber charges. Could not the mobile operators begin to compete with mass market tariffs?

Network installation is more or less a fixed charge such that increased traffic volumes do not significantly increase system operating costs. Yet the social and commercial benefits of wider mobile phone ownership are not often mentioned: more efficient manpower usage in business, easier communications for vulnerable sectors of the population, greater public involvement in the prevention of crime.

It all comes down to the fact that accessible communications makes for a better society. And there is nothing more accessible than a mobile phone.

One might argue that the Government has done enough by allowing free market competition to throw up competing telecomms operators with their own marketing agendas. But it could be equally argued that a tilt towards a mass subscriber market would not have been out of place in the licensing arrangements. After all, the Government is happy to do this when it comes to broadcasting which is why it empowers bodies such as the ITC.

Electronics has given us at little cost one-way communication to wherever we happen to be in the form of radio and tv. It seems logical that it should do the same in enabling us to talk to each other.

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Licensing shakeup heralds spectrum sell-off

The British government is rethinking its policy on licensing radio spectrum. In the future it will be easier to get licences to use radio technology, but they will cost more, especially if the technology does not make efficient use of the spectrum. The licences will probably be auctioned to the highest bidder. By the DTI's own estimate, electronics will be the world's largest industrial sector in the year 2000. Much of the new technology will rely on radio links instead of wires. The DTI says it wants to stimulate innovation with the increased use of radio technology. But although the outbreak of peace has encouraged the military to relinquish some of the radio frequencies it was holding on a just-in-case basis, demand will always far exceed supply.

"There can never be any completely new spectrum" says Jim Norton, Chief Executive of the DTI's Radiocommunications Agency, which currently allocates frequencies. "But new technology can make more efficient use of what there is. We have to encourage people to use that technology".

The RA has now published a consultative document, *Future Management of the Radio Spectrum*, and is inviting comments.

When licensing frequencies the RA currently charges enough to cover its administration costs. There is thus no incentive for licence-holders to use their spectrum efficiently. The RA wants more flexibility when charging so that those who waste spectrum by using old-fashioned radio systems will have to pay more than those who use new technology which wastes less spectrum. The transition from analogue to digital technology, says the RA, would have a "profound effect" on spectrum efficiency.

The RA would like Britain to follow Australia and New Zealand, and auction frequencies to the highest bidders. Successful bidders could then sell or subcontract the frequencies. Radio allocations cannot currently be transferred in the UK.

The RA also wants to subcontract the job of allocating some frequency blocks to independent spectrum management organisations. The RA would then act as a wholesaler. The aim is to make it easier for people to get licences.

The status of licence-exempt services which pay no licence fee would also have to be resolved. These services rely on equipment which the public can buy and use without needing any licence. They already include cordless phones, remote controls, security devices, and burglar alarms which people can fit without needed to lay wires.

The consultative document is available free from the Radiocommunications Library, Waterloo Bridge House, Waterloo Road, London SE1 8UA. Comments should be sent to Laurence Green, PO Box 3989, London SE1 8YD. Cut-off date is 22 July.

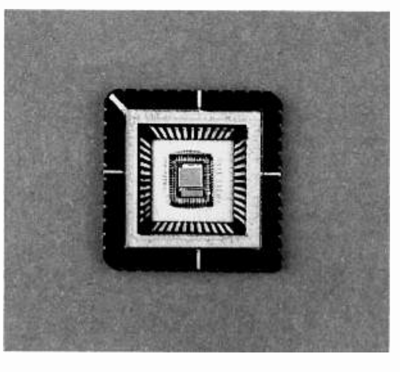
Barry Fox

Camera chip for less than $10

Edinburgh-based camera-on-a-chip designer VLSI Vision (VVL) has introduced its first high volume standard product – a single chip camera which costs less than $10. VVL already has customers lined up for the device which integrates a 160 x 160 pixel array with the A/D converters necessary to produce a digitised image. One application is expected to be in the emerging video communications market.

The device, called the *WL1070*, includes all the control circuits required for an autoexposing camera and its 30mA current consumption will allow it to be used in hand-held cameras. The device has a single element plastic lens.

VVL has been integrating image sensing arrays with A/D conversion and digital signal processing functions for the last three years.



Poor reception for new satellite

Three million homes in the UK, and 12 million more round Europe will find that their dish systems will need modification or replacement to receive signals from the new satellite which Astra plans to launch this autumn.

Luxembourg company Societe Europeenne des Satellites, launched its first satellite five years ago, and now has three, known as Astra 1A, 1B and 1C, at the same point in space, at 19° East. Each broadcasts 16 tv channels in the frequency range 10.95-11.7GHz. Thus one dish aerial and one receiver can receive a total of 48 channels.

SES will launch a fourth satellite, 1D, this autumn. Although 1D was planned as a backup, to take over transmission duties if there is any technical failure on 1A, B or C fail, broadcasters want more channels for new programmes. So Astra has said it will let them use 1D to provide 16 new channels.

The 1D satellite will have to broadcast the new channels between 10.7 and 10.95GHz. Astra's estimate is that 90% of existing dish systems will not work properly in this band.

SES says it started warning receiver-makers about the need for wider frequency design in early 1992, and published a technical specification in December 1992. But manufacturers are only now starting to sell wideband receivers, labelled "1D compatible". SES fears that some firms will now try and sell off old stock, which cannot receive from 1D, before the public understands the problem.

●British companies Pace and NTL (formerly the research laboratories of the disbanded Independent Broadcasting Authority) have joined forces to make digital receivers for home use. Nokia of Finland has teamed up with US company TV-Com to make rival models. All will use MPEG-2 digital compression. Pictures compressed to a rate of 2MBit/s give quality similar to VHS tape, and let each satellite channel carry up to 16 programmes. At rates of 4 or 8 Mbit/s, picture quality is perfect, but each satellite channel can carry only four or eight programmes.

European satellite operator Eutelsat has demonstrated the technical viability of a novel twist to the idea. Its transmitters have a frequency range of 36MHz. Eutelsat will use 27MHz to broadcast a programme in conventional analogue form, and the other 9MHz to deliver an 8MBit/s data stream, which carries 1, 2 or 4 digital tv programmes.

B.F.

Microsoft in space

Nothing grabs the headlines like big numbers, especially when they come from Bill Gates, the multi-billionaire founder of Microsoft. But even the most imaginative aerospace engineers caught their breath last month when Gates revealed a plan to build a $9bn global computer network using no fewer than 840 low Earth orbit (LEO) satellites.

The scale of what Gates and his partner mobile phone entrepreneur Craig McCaw are proposing is so much larger than any satellite project past, present or planned that it is all too easy to either dismiss it out of hand or marvel at the audacity.

But Gates and McCaw, with the help of NASA space scientists, are only part of the latest drive, which includes many of the world's largest telecommunications operators, to build mobile communications networks using extraterrestrial infrastructure.

This is virgin territory and opinion is divided on the most feasible technology. But engineers with years of experience of building and launching satellite systems have expressed doubts over the feasibility of using a network of over 800 satellites.

Teledesic, the company Gates and McCaw have set up to develop their global LAN, is proposing to create an orbiting packet-switched data network of 840 LEOs, 40 satellites in each of 21 polar orbits some 700km above the earth. Computers and hand held terminals will be connected around the world using 30GHz uplinks and signals will be routed in space around the network of satellites.

The question is, even if it were feasible to launch and maintain that many satellites, would they be able to offer an acceptable service which is proposed to include carrying and switching not only 16kbit/s data links, but also broadband video transmissions up to 2Gbit/s?

International satellite operator Inmarsat has already rejected the use of LEOs for its satellite-based mobile phone system, while Motorola's Iridium satellite phone project has already revised the number of LEOs it will use from 77 to 66.

Inmarsat decided there were too many uncertainties in launching and maintaining a network of 54 LEOs. But an important factor was system complexity with most calls needing to be switched between LEO satellites.

There is no technical precedent for the proposal. A single network of 50 LEOs – let alone 840 LEOs – has yet to be created. The nearest equivalent is the US and Russian global positioning systems. The GPS satellite system created by the US defence department uses a network of 24 LEOs. "But that took 15 years to create and many satellites were lost along the way," said a spokesman for Inmarsat. However, that may not be a fair comparison, as the US defence department had no obligation and little motivation to rapidly roll out all the satellites needed for a commercial GPS service. The proposed mobile communications services will have every commercial incentive to overcome the technical and regulatory hurdles and launch satellites quickly. Chris Elliott, a satellite specialist at Smith System Engineering, believes the only questions facing Teldesic and Iridium are financial and regulatory rather than technical: "Least of the worries is the technology, you can buy your way out of most problems," said Elliott.

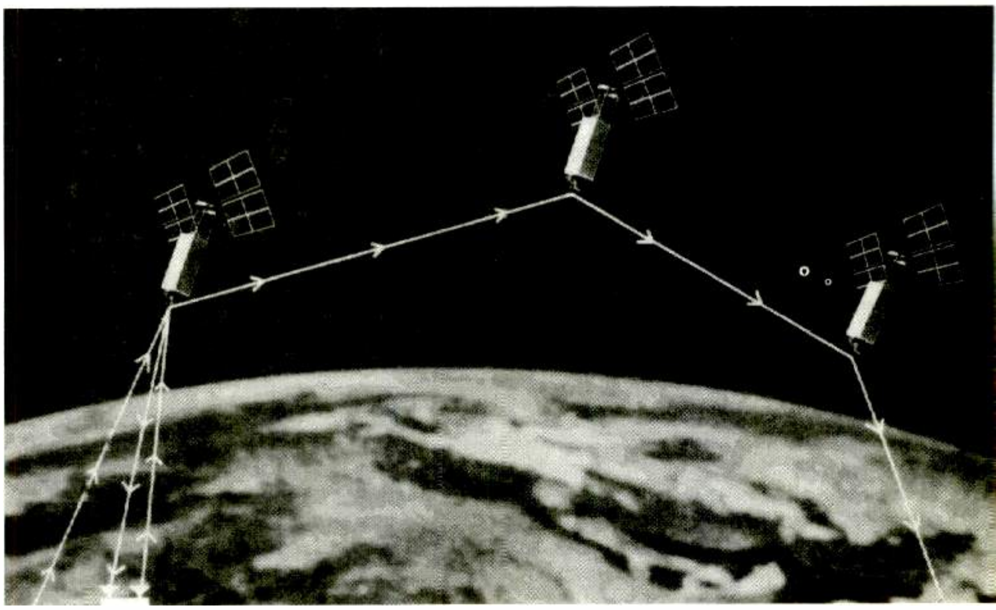
But it costs money to launch and maintain a network of LEOs, which may each cost as little as $2m. The Russian Glonast satellite positioning system has had more problems. Its LEOs have failed after as few as 18 months' operation. The scale of the Teledesic plan magnifies the difficulties associated with launching LEOs. Even a 10% failure rate could be cost crippling to the dreams of Bill Gates. Even if the launches were successful, there are still uncertainties over the feasibility of constructing a digital communications network with the latest asynchronous transfer mode (ATM) switches in space. Satellite links at 30GHz are certainly not trivial, said Elliott. Another option is to route signals back down to earth stations where they could be switched using terrestrial networks. The speech delays due to the multiple hop satellite links could be anything up to a second.

Teledesic's plan relies on being able to switch traffic in space at up to 2Gbit/s. That is state-of-the-art digital switching technology in land-based systems. It is far from certain how they can be built into a satellite payload which requires the use of specialised radiation-hardened semiconductors to withstand the space environment.

Teledesic's plans are being taken seriously by other operators including Inmarsat. The use of the 30GHz band, if Teledesic is allowed to use it worldwide, will mean the system will not be bandwidth limited as may be the case with Iridium, according to Elliott.

Secondly, the large number of satellites will ensure areas of high population will have more than one LEO overhead at anyone time. This improves coverage in built-up areas and it also makes in-building access a possibility, said Elliott.

Perhaps the most important factor is that Teledesic has two of the most successful IT pioneers behind it.



Microsoft founder Bill Gates plans to build a $9bn global network relying on 840 satellites.

Powered up PowerPC

PowerPC partners IBM and Motorola are to produce a faster version of the PowerPC microprocessor as the race to push ahead of Intel *Pentium* performance continues.

The new *PowerPC 604* will run at 100MHz with a SPEC integer rating of about 140. It will be used in new computers from IBM and Apple to be launched mid-November at the Comdex/Fall trade show.

Samples of the 100MHz chip are expected in the next quarter at a US price of around $600. Intel says that it will have a 150MHz version of the *Pentium*, code named the P6, ready by the middle of 1995 or earlier.

IBM and Motorola are expected to have completed the port of Microsoft Windows NT operating system to the PowerPC by the middle of this year which will open up a large part of the market that was formerly stuck with Intel architecture systems.

Auto radar avoids collisions

Imagine you are driving along and the car in front stops abruptly and, before you know it, your car also starts to brake thus avoiding a costly shunt.

The system which allows a constant time interval between the vehicle in front and your car is called Autonomous Intelligent Cruise Control (AICC) developed by the Philips Research Centre. Jaguar Cars, British Aerospace and Rover are also a part of this project which resulted in a prototype radar which in its trials has proved a cheap safety aid to drivers.

The AICC needs a sensor to determine the distance to the car in front at all times. A strong candidate for the task is millimetre-wave radar.

Front-end hardware comprises antenna, microwave head and associated control circuitry and a signal and data processor sub-system where the sensor data is processed. The user interface is connected to the vehicle controller via a bi-directional link.

The controller's decisions are based on information gathered from the radar and the driver and then passed onto the brake and throttle actuators. The driver can override the system's actions at any stage.

According to Dr Andrew Stove, the project leader who has worked on car radars since the late '80s, the most novel part of the whole concept is the antenna designed by the Philips Research Centre.

The basic principle of a frequency scanned antenna is not new," said Stove. "But we've , developed a 94GHz and 77GHz antenna that can be injection moulded, meaning low cost."

The antenna, 160mm in length, is an array of 28 smaller phase-array antennas as radiation elements which are based on Frequency Modulated Continuous Wave (FMCW).

FMCW works on the principle of mixing the transmitted and received frequencies in the GHz range to generate a beat frequency in the kHz range making it easy to process.

By measuring the beat frequency, the range to the target can be calculated by following a simple formula involving the beat frequency and the velocity of the signal propagation.

Using FMCW modulation according to Dr Stove proves a good choice for an automotive radar. Because its transmission is continuous, the radar can use a transmitter with a much lower peak output power (approx. 10mW) and allows a simple Gunn-effect device to be used for the transmitter and the receiver oscillators, making the overall transmitter design simple and cheap.

The electronically scanned antenna scans the beam to map out the area ahead of the car. Previously, for automotive applications mechanically scanned antennas were used, but they proved expensive and somewhat unreliable.

As there is usually more than one target present in the scanned scene, the spectrum of the IF signal is analysed, to find the position of possible targets, by a *56001* DSP chip which performs a FFT (Fast Fourier Transform) of 256 to 1024 points, depending on the system used. The prototype system operates at a typical range of 150m.

The project is part of Prometheus and started two years ago under the ARIADNE (Application of a Realtime Intelligent Aid for Driving and Navigation Enhancement), part of the DRIVE 2 initiative. The system was developed a year ago and tested on a Rover, and more recently on a Jaguar Sovereign. It is expected to go in upper range model cars with a volume price of around £10.

Traffic news in a flash: one of Metro Traffic Control's fleet of four traffic spotting aircraft equipped with air to ground broadcast link equipment. Each aircraft station comprises a wide audio bandwidth 400 to 500MHz transmitter for the broadcast downlink and two monitor receivers. One receiver would normally be tuned to the radio station's output while the other supplies the uplink studio cue. The airborne station also includes some simple mixing equipment. The complete setup is powered by an acid gel battery.
Metro uses the Wood&Douglas supplied equipment to service 24 radio stations in and around London including LBC, Invicta and Spectrum. Metro plans to expand the service to a further six metropolitan areas.

Channel $2^1/_2$

Channel 5 Broadcasting Ltd, the consortium of MAI, Pearson and Time Warner that wants to launch a fifth terrestrial tv channel in Britain, has come up with a new answer for critics who argue that its plans to start another old-fashioned analogue service will block the UK's future plans for a completely new digital service.

Believing that the government is on the point of making a final decision, the consortium says it will use only one of the two frequencies earmarked by the government for Channel 5 (UHF tv channel 37), leave the other (35) free for digital use and make up the difference by using completely different frequencies which it claims are available.

This, argues the consortium, will also have the happy effect of "significantly reducing" the interference which a fifth channel will cause to many millions of existing VCRs, satellite tuners and video games. The consortium estimates that retuning costs due to the new service will be £20 million.

The new plan halves the capacity of any future digital service, from 8 new programmes to 4, or 4 to 2, depending on which of the consortium's figures you pick. Viewers will be less likely to buy a digital decoder if there is little extra choice. **B. F.**

TV decoder in double standards

General Instruments has demonstrated a dual-mode set top video decoder which is compatible with both MPEG-2 and its own DigiCipher digital TV compression algorithms. Products based on the new Digicipher II technology were demonstrated at the recent Cable and Satellite exhibition in London.

US MPEG-2 chip maker C-Cube Microsystems and Motorola have agreed with GI to licence the DigiCipher II technology.

Digicipher II is a follow-on from GI's DigiCipher technology and gives the option of two compression modes (DigiCipher and MPEG-2), which ultimately allows interoperability between the two.

DigiCipher provides high quality video transmission with fast channel acquisition (300ms) and low system end-to-end delay. It is significantly cheaper to implement than MPEG-2 as it has eliminated the use of B-frames (bi-directional frames).

Although DigiCipher provides desirable features for consumer television, interactive video and video-on-demand, it has retained its open architecture so it can provide future services such as computer generated images and still images sent down low bit rate communication channels.

DigiCipher was introduced in the US last year and is already used by US broadcast operators such as Primestar, a digital service is capable of delivering 77 channels. ■

RESEARCH NOTES

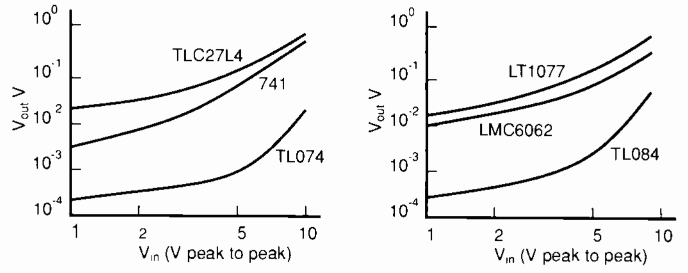
Unwanted signals that sap op-amp performance

Ever-growing concern about electromagnetic compatibility has led to numerous studies of the effects of rf signals on the performance of most common electronic components. Bipolar transistors, fets and common op-amps have been extensively investigated. Most investigations have looked into the tendency of such components to rectify amplitude-modulated rf signals and turn them into unwanted radio programmes or requests for taxis.

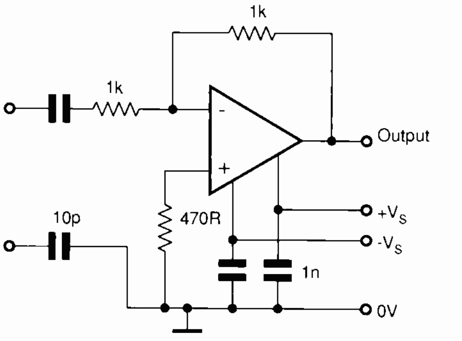
But A. S. Poulton (*Electronics Letters*, Vol 30, No 4) has taken things a stage further by examining the dc performance of various common op-amps with out-of-band-interference present at the input terminals. Poulton says that while some forms of induced noise may be acceptable when integrated over a long period, electromagnetic interference can have more subtle nuisance effects that do not cancel out, particularly a change in the dc offset voltage.

Poulton's experimental set-up simulates the sort of situation that might occur in mixed analogue and digital circuitry, where high frequency square waves are loosely coupled to the op-amp's input. The 10MHz frequency chosen is well outside the op-amp's normal pass band and is coupled via 10pF capacitors from a balanced signal generator. The third piece of experimental kit is a dc millivoltmeter connected to the op-amp's output to monitor the effects of the simulated interference. All three are independently powered to avoid ground or supply line loops.

The standard test circuit employs the inverting amplifier arrangement shown in the figure, although Poulton also tests a non-inverting buffer circuit which produces broadly similar results.



Rectification increases rapidly with increasing input voltage.



Amplifiers were connected in an inverting circuit and tested with 10MHz square wave input to assess their emi susceptibility.

Balanced square waves at 10MHz, varying from 1V to 10V pk-pk, were fed to the various op-amps – including *741, TL074, TL084, TLC271, LT1077* and *LMC6062*. In each case, with the exception of the *741*, the output voltage increased sharply and in a positive direction, with increasing rf input signal.

Poulton concludes that all op-amps appear to rectify rf to some extent, though jfet and cmos types appear to be less susceptible that those with bipolar input circuitry.

A variety of other factors, including bias levels, interference pulse length and parasitic capacitances also play a part. Poulton observes that feedback capacitance does very little to reduce susceptibility, probably because open-loop ac gain is already low at the frequencies in question.

The measured changes in dc performance noted may not be huge, but Poulton says that they can be enough to compromise the low offset that is normally taken for granted in these devices. An effective cure in many cases, is to fit a capacitor between the input pins. It must be of a sufficiently high value to bypass the rf, but still be small at frequencies

NRPB picks up amateur gauntlet

Amateur astronomer Anthony Hopwood's observations on the power line/leukaemia link (Natural radiation focused by power lines: new evidence, *EW + WW*, November, pp.912-915) have prompted the professionals at the National Radiological Protection Board to test his hypothesis – with a negative result.

At first sight this may appear a triumph of the professional scientist over the amateur. In reality, Hopwood has made a worthwhile contribution and prompted the professionals into carrying out the detailed measurements for which they alone are equipped.

In purely scientific terms, the debate still lacks some plausible theory to underpin the present rather weak epidemiological evidence. Research is progressing on two basic fronts: improvement of statistical data from population studies, and exploration of possible mechanisms that might link power cables with biological changes that would lead to cancer.

Based on measurements made with a simple Geiger-Muller instrument, Hopwood theorised that power lines might somehow focus cosmic rays, concentrating them in two parallel strips, either side of the line. Since cosmic rays are high energy particles that can – like x-rays – cause ionisation in biological material, here was what appeared to be a very plausible hypothesis. But one of many theoretical difficulties is the fact that 11kV overhead lines – the type under which Mr Hopwood made his measurements – radiate only a fraction of the energy necessary to deflect a cosmic ray.

The National Radiological Protection Board (NRPB) was one of the agencies stimulated by Hopwood's work. After an initial negative result – using professional instruments but under the medium voltage lines where Hopwood made his first measurements – the NRPB team sought the cooperation of National Grid Company to repeat the experiment under a major 440kV transmission power line at Walland Marsh. The geology was uniform and the experimenters had access to records of the actual power flowing in the cable.

In the end, the NRPB scientists still found no significant variation in cosmic ray count at any position under the power lines. The NRPB concludes, with minor reservations, that there is no way in which a focusing process could account for any suggested elevation in the risk of cancer under power lines.

Replying, Hopwood says that the failure of the professional scientists (and himself) to find any focusing of cosmic rays in recent experiments could be due to the decline in solar activity since the last solar maximum.

Undoubtedly, even within the world of professional science, there are many research roads that lead nowhere... or can occasionally produce unexpected and spectacular results.

But we should never underestimate the creativity of the amateur scientist. Long live kitchen table experimenters!

Could safety net make air travel safer?

Neural networks being developed at the Georgia Institute of Technology in Atlanta could help make the world a little safer – particularly for airline passengers. Georgia networks are currently being used to help identify the risk factors that lead to a fire when an aircraft crashes.

The study, led by Dr James Mahaffey, began with a commission to help make the transport of weapons less dangerous. But in principle the technique could be used to predict the outcome of hazardous waste spills, disease epidemics or indeed any situation in which there are a large number of complex inter-relating factors.

A problem is that when risk factors are as complex and numerous as they are in aircraft operations, traditional probabilistic risk assessments are limited in what they can tell you. Mahaffey explains that you start multiplying rare probability by rare probability and end up with very small numbers that can be meaningless.

Neural networks, on the other hand, can spot weak correlations between thousands of different types of data – compared with the dozen or so that even the best human mind can visualise.

To learn enough about what happens in an aircraft crash, Mahaffey's neural network, running on a Microway computer, had to be fed with a vast amount of data. Records of 17,000 crashes, covering different aircraft types, weather conditions, pilot experience and thousands of other likely factors proved barely adequate to train the network properly.

Nevertheless, when put through its paces using data from 2000 real-life crashes, the system correctly predicted the extent of fire in around 70% of cases. This compares with a figure of around 10% using standard probabilistic techniques.

Mahaffey's neural network shows, perhaps surprisingly, that almost every measurable factor is correlated to some extent with whether or not an aircraft will burn when it crashes. Obviously, some factors are more closely and more strongly associated than others.

Fire could be less of a risk in aircraft crashes following the work of James Mahaffey at Georgia Institute of Technology.

What the neural network doesn't seem to have learned yet are the sheer practicalities of air transport. When asked to predict the safest type of aircraft with which to transport weapons it came up with two rather intriguing suggestions: a glider and a helium balloon. Next on the list of fireproof possibilities was any aircraft flying over water with empty fuel tanks.

As Mahaffey readily admits, the last example might not burn, but it would certainly crash.

Supercomputer previews Jupiter blockbuster

The impact of ShoemakerLevy 9 on Jupiter is expected to be one of the most spectacular celestial events in the Solar System. There's only one snag as far as we on Earth are concerned; the collision will take place on the side of Jupiter facing away from us. But scientists at Sandia National Laboratories are hoping that by riding on the back of the fastest supercomputer in the world, they could, in effect, be transported into the heart of the action.

As the impact will happen on Jupiter's dark side, no-one will see the collision directly and its evidence will reach us second-hand, either from reflections off Jupiter's moons or else by surveying the damage to the planet's surface as the impact site rotates to face us.

So researchers at Sandia are using their 1840-CPU Intel Paragon to improve their predictions of what is going to happen. The team, from the Computational Sciences and Mathematics Center have been concerned mainly with the processes that are expected to occur as the cometary fragments hit Jupiter's atmosphere and disintegrate explosively. The latest update, presented at a Lunar and Planetary Science Conference at the Johnson Space Center, suggests that as the cometary fragments break up in the hydrogen/helium atmosphere they will deposit their energy in a column of hot gas at high pressure. The column will then expand explosively into the surrounding atmosphere, creating a fireball that grows upwards supersonically, at speeds approaching that of the incoming comet, only in the opposite direction.

The researchers are also seeking the answer everyone wants to know; what, if anything, will be visible from the Earth?

Earlier estimates have suggested that astronomers would have to rely on reflections or observations from spacecraft. Updated orbits calculated from new images taken from the recently repaired Hubble Space Telescope now put the impacts tantalisingly close to the limb of Jupiter, the edge of the planet as seen from Earth. In fact, according to the new calculations, some of the cometary fragments will not disappear behind the planet until they have already entered the thin atmosphere 400km above the denser clouds. Less than a minute later, the hot fireball is expected to rise back into view.

Whether it is bright enough to be visible from Earth is still an open question, though it is expected to emit detectable radiation at other wavelengths – enough to validate the latest calculations.

Mains hum – or ear drum

Hum, such as that generated by the mains, can be a nuisance. But for some, it could be much more of a headache than they realise.

There can be few engineers brought up before the days of monolithic stabiliser chips who are not conversant with hum. Even in professional recording or broadcasting studios, the usual way of telling whether the equipment was switched on was to put an ear to the speaker and listen for the ever-present 50 or 100Hz. But though mains hum is now a rarity in good audio equipment, it has not disappeared from our lives entirely.

An editorial in the *British Medical Journal* (Vol 308, No 6925) by professor C G Rice of the Institute of Sound and Vibration Research at Southampton University says that hum – from whatever source – still causes much annoyance and ill-health. Moreover, many reported cases of low frequency hum remain unresolved, even after extensive investigation.

Rice points out that mysterious hums have been the subject of scientific investigation for more than 20 years, in everything from universities to research establishments, companies and charities.

The *BMJ* editorial says that some 90% of all reported cases of noises or vibrations in the 1-150Hz region can be identified and dealt with by acoustic consultants or environmental health officers. Causes include supply utilities, industrial plant, electromagnetic radiation equipment and furnaces. But there remains a mysterious 10% of reported cases of hum where no external source can be found.

Once the obvious environmental causes have been eliminated, says Rice, an investigator should take a look at the complainant. Some variants of tinnitus, he says, can manifest themselves as a continuous or throbbing low frequency hum.

But if the victim resorts to extreme measures – such as moving house – to escape the noise, then a brain tumour is a very likely cause.

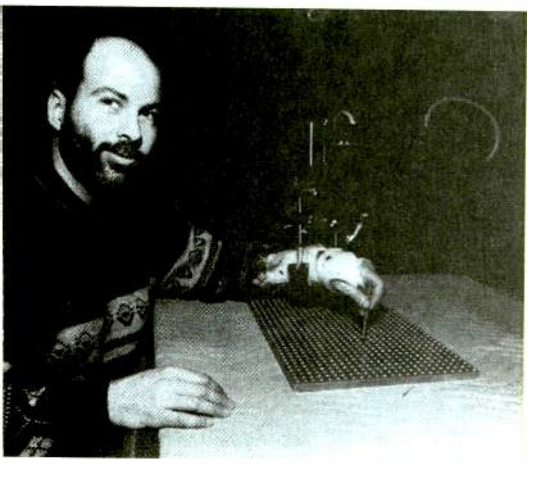
A less dramatic possibility is that some people may have increased hearing sensitivity at certain specific low frequencies, though laboratory studies have so far failed to demonstrate this unequivocally. For such people a hum which is tolerable or even unnoticed by normal people may well cause annoyance.

More research into hum is clearly needed, though as professor Rice points out, one almost insuperable difficulty is the highly individual nature of each and every case.

Mystery hums, it seems, will continue to be a feature of life for a minority of the population.

MIT robot gets a healing hand

Giving your limbs a good work out could soon take on a new meaning. Because engineers at the Massachusetts Institute of Technology (MIT) have developed a robot that can learn exercise routines from a physiotherapist and then go on to guide a patient through those same exercises. Uniquely, according to its developers, the robot can also record all the resulting biomechanical data and use it to make objective measurements of the patient's condition, together with his or her progress.



Arm twisting: MIT's robot can guide patients through physical therapy and assess their progress.

Professor Neville Hogan, leader of the team that developed the robot, says it will be used in clinical trials this Autumn on people who suffer mechanical deficits arising from strokes.

At present there is a great deal of debate over whether manual manipulation of stroke victims' limbs really helps them recover. The new robot has two properties that should help decide the matter once and for all. In the first place it can take the patient through a completely standardised exercise routine; secondly, it can monitor precisely the patient's response to that exercise.

The system can record exactly the amount of force a patient applies, the velocity of any movements and the absolute position of the limb at any moment in time. Normally, such movements can only be assessed qualitatively by touch and feel.

The robot, called 'Manus' (Latin for 'hand'), is decidedly user-friendly. As well as performing the basic mechanical functions, it also provides video games to hold a patient's attention during what can often be very monotonous exercises. The games correspond to a patient's performance at any given exercise and provide a visual clue that he or she is not pushing hard enough or pushing too hard. Four different games have been designed for patients at different levels of recovery.

As well as operating in its therapeutic mode, Manus could also become an important teaching tool. Professor Hogan suggests that it could be used to record the hand movements of a skilled doctor, surgeon or physiotherapist as that person manipulated a dummy (or even a real patient). A trainee could then repeat the exercise while the machine recorded every difference in detail. With a second slave machine, the "expert" could even feel directly what the trainee was doing – or not doing. And of course the slave machine could be thousands of miles away on the end of data link.

Research Notes is written by John Wilson of the BBC World Service.

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Applied i/o for the PC

Controlling i/o from a PC via a free RS232 port is simple, convenient and fast enough for most applications. Bill Teliki discusses the topic – from sending and receiving data to scanning photgraphs on a dot-matrix printer.

There are three main ways of controlling input/output devices from a PC, via the Centronics printer port, via a dedicated i/o card or via the serial port. Although faster than the serial alternative, the Centronics port is difficult to read from and write to. It also has the drawback that it prohibits use of the printer.

A dedicated i/o card can be fast and inexpensive. However it takes up a slot and ties the interface card to one machine. Additionally, laptops have no card slots so the interface cannot easily be made portable. Input/output cards can also be inflexible, often requiring one card for each function.

Most PCs have two serial ports and one is usually free. The interface is well documented and programmable in a variety of languages. If correctly buffered, a serial interface also provides some immunity against power surges. In the event of an overload, the buffer chips will usually fail, leaving the PC intact.

Circuit description

Data to and from the serial line is buffered by IC_3 of Fig. 1. Using a *MAX232* minimises power supply requirements. From a single +5V rail, it produces the ±10V levels needed for correct RS232 transmission.

Serial data is converted to parallel bytes and *vice versa* by the universal asynchronous receiver/transmitter, IC_4. Data bytes are buffered by $IC_{5,6}$. To send data presented at the inputs to IC_6 down the serial line, a negative pulse needs to be applied to the transmitter-buffer register load input, /TBRL, on pin 23. When a low level is applied to the pin, data present at $TBR_{1\text{-}8}$ is transfered into the transmitter-buffer register. When the /TBRL goes high, data is transfered to the transmitter register from which it is transmitted serially.

Conversion of data from serial to parallel is equally straightforward. Whenever a byte of data arrives down the serial line, the data-received output, DR, goes high on pin 19. If this output is then inverted and fed into the data received reset, /DRR, on pin 18, the data-received output will be reset. In this case the inversion is carried out by a *BC109*.

This arrangement means that multiple byte data can be sent down the serial line and the data-received output used to drive latches, etc. It also avoids wasting a data bit line.

Because all data in/out lines are terminated with 10kΩ pullup resistors, the length of ribbon cable that can be driven is approximately seven feet. Bit rate limits of the board are 600 and 76,800, although for 76,800 a special software communications library unit will be needed together with a faster uart in the host PC. One such device is the *NS16550*.

Serial input/output

Standard PCs have a set of bios routines that can be accessed by a high level language such as Pascal. These routines allow a maximum bit rate of 9600baud, or 19,200 for the PS2. This means that you can output data in bytes at 960Hz, assuming one start bit, eight data bits, no parity, and 1 stop bit.

Digitisation is limited to half the rate of data output. This is because any a-to-d converter has to be triggered before it will convert, and the trigger signal usually comes down the serial line.

One way of increasing the data rate for a-to-d conversion is to send one trigger signal and then collect a predetermined number of data bytes. In this way data can be gathered in bursts interspersed with minimal gaps.

Programming

The kernel software **Listing 1**, shows the essential procedures that can be used to com-

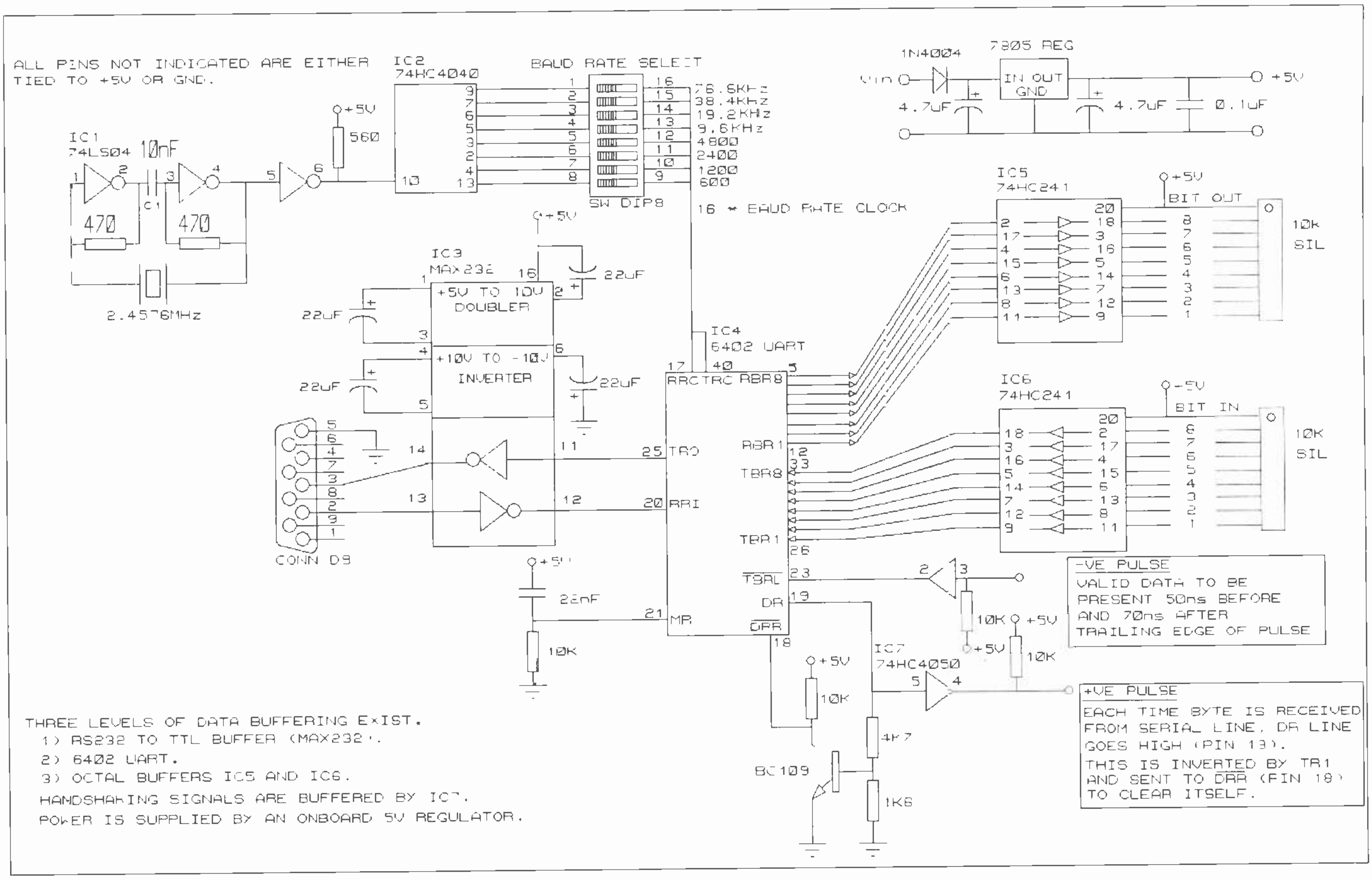


Fig. 1. Advantages of interfacing a PC to i/o via the RS232 interface are numerous. Being external, a serially fed i/o card can be plugged easily from one PC to another. It provides some i/o fault immunity to the PC and software writing is easy. Achieving a high operating speed however involves very high serial communication rates.

This image was scanned into a PC using the head traversing and paper feed mechanisms of a dot matrix printer. An optical fibre fixed to the print head scanned the image. Light output from the fibre was sensed via the circuit of Fig. 2 on page 455 and resulting digital data was fed to the PC via the serial interface above.

municate with the board. The functions of bits in the MOV AL,11100011B statement are,

```
Bits        Function
76543210
000         110 baud
001         150 baud
010         300 baud
011         600 baud
100         1200 baud
101         2400 baud
110         4800 baud
111         9600 baud
   00       no parity
   01       odd parity
   10       no parity
   11       even parity
     0      1 stop bit
     1      2 stop bits
      10    7 bit word length
      11    8 bit word length
```

A-to-d conversion

Figure 2 shows the circuit diagram of an a-to-d converter with associated circuitry for digitising light intensity. Whenever a byte of data from the host computer is sent down the serial line, a brief positive pulse is sent out from the RS232 board. This is used to trigger the a-to-d converter.

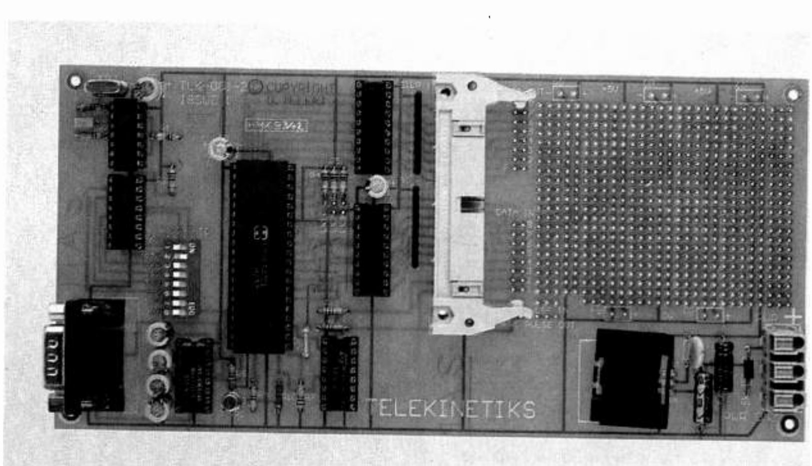
When conversion is complete, the INT line output of the a-to-d fires the *74121* monostable. This in turn provides the negative pulse to the board which then sends the data to the host.

The op-amps are single supply rail types. If a photodarlington transistor is used then problems will occur due to the high sensitivity of the device. This will make it difficult to obtain a linear response from dark to light.

I have tried the circuit with a few different types of photo-transistor and without problems. Even a *BC109* with its top cut off produced acceptable results, although the 10kΩ potentiometer in Fig. 2 had to be replaced a 1MΩ type. **Listing 2** shows a program fragment using the routines of the kernel listing to communicate with the board and take data from the a-to-d.

Listing 1. Kernel routine in Pascal for communicating with the PC i/o card over an RS 232 link.

```
PROGRAM SERIAL (INPUT,OUTPUT);
USES
DOS,CRT;
VAR
   BYTE_RCVD : BYTE;
  DATA,RT_CD : BYTE;
PROCEDURE INIT_COM;
BEGIN
  ASM
    MOV AL,11100011B;    {9600 BAUD, 8 BITS, NO PARITY, 1 STOP BIT.}
    MOV DX,1;            {0=COM1, 1=COM2, 2=COM3, 3=COM4           }
    MOV AH,0;            {SPECIFY SERVICE 0                        }
    INT 14H;             {BIOS COMM, INTERRUPT                     }
  END;
END;
PROCEDURE SEND_BYTE(BYTE_TO_SEND:BYTE);
BEGIN
   ASM
     MOV AL,BYTE_TO_SEND;{SPEC THE BYTE                            }
     MOV DX,1;           {SET COM2                                 }
     MOV AH,1;           {SPECIFY SERVICE 1                        }
     INT 14H;            {BIOS COMM, INTERRUPT                     }
   END;
END;
PROCEDURE GET_BYTE;
BEGIN
  ASM
    MOV AL,0;            {CLEAR AL                                 }
    MOV DX,1;            {SET COM2                                 }
    MOV AH,2;            {SPECIFY SERVICE 2                        }
    INT 14H;             {BIOS COMM, INTERRUPT                     }
    MOV BYTE_RCVD,AL;    {PUT THE BYTE INTO BYTE_RCVD              }
  END;
END;
PROCEDURE QUIT_COM;
BEGIN
  ASM
    MOV RT_CD,AL;        {RETURN CODE :: LOOK AT IF SOMETHING WRONG}
    MOV AH,4CH;          {CLEAN UP AFTER FINISHED                  }
    INT 21H;             {DOS SERVICES INTERRUPT                   }
  END;
END;
BEGIN
  INIT_COM;
  {SEND DATA AND/OR GET DATA USING SEND_BYTE AND GET_BYTE}
  QUIT_COM;
END.
```

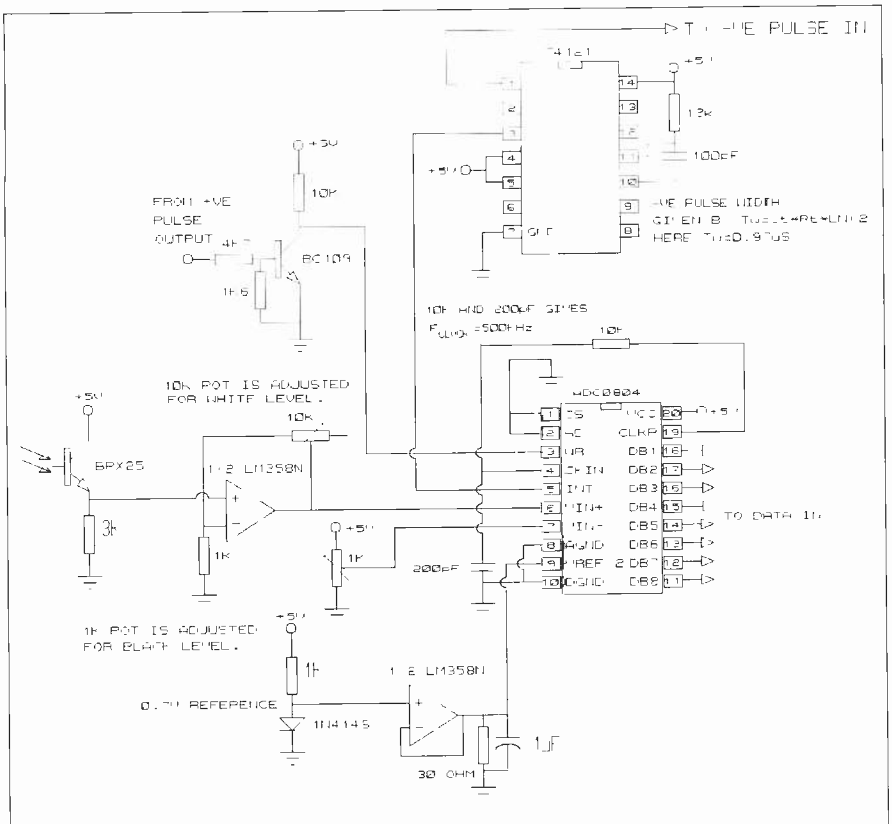


Fig. 2. Measuring light intensity via a PC. A-to-d conversion of light intensity to 8 bits provides a theoretical dynamic range of 48dB. More elaborate biasing around the opto transistor could improve this.

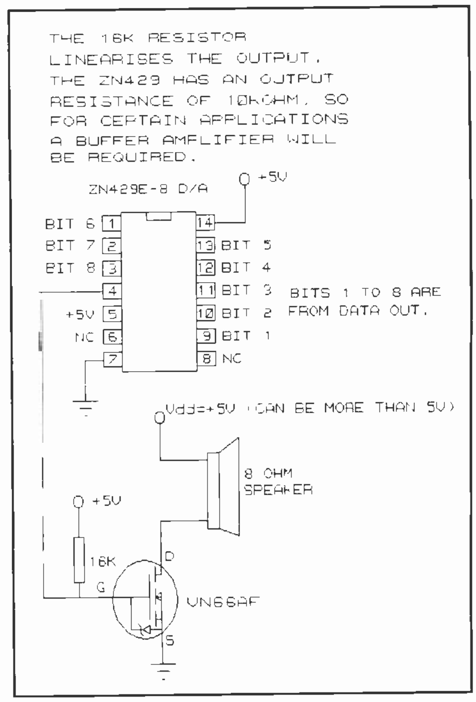


Fig. 3. This circuit uses a ZN429E d-to-a converter driving a VFET for high power. Depending on the output data rate and application, an interpolation filter may be required.

Figure 3 shows a *ZN429E* d-to-a converter driving a v-fet for power amplification. Many loads could be connected to the fet drain, but in this case a loudspeaker has been chosen. **Listing 3(a)** is a simple example of how to send data to this circuit to provide waveform synthesis. Depending on the output data rate and application, an interpolation filter may be required.

Replacing the loudspeaker with a dc motor needs a diode in parallel with the motor to prevent back emf damaging the driving circuitry. A *1N4001* with its cathode connected to the positive rail should suffice. **Listing 3(b)** shows how the motor can be accelerated up to maximum speed, held there a moment, and then decelerated.

Controlling a stepper motor

Figure 4 shows a high-power stepping motor interface and **Listing 4** its driver software. Since the motor in this example is a four-phase type requiring a ring counter sequence, binary values of each step have to be converted to decimal numbers before transmission. A value of 1100_2 for example represents 12, value 0110_2 is 6, 0011_2 is 3 and 1001_2 is 9.

Pascal background

Pascal is a procedural language, which means that all functions are first declared as procedures and then called up at the end of the program.

For example:

```
PROGRAM TEST;
USES DOS,CRT,PRINTER,GRAPH;
VAR
  A,B,C,D : REAL;

PROCEDURE BLAH(E:REAL;VAR F:REAL);
BEGIN
  F := E/2;
END;

PROCEDURE STUFF(ZARQ,BLARQ:REAL;VAR RESULT:REAL);
BEGIN
  RESULT := SQRT(SQR(ZARQ)+SQR(BLARQ));
END;

BEGIN
  WRITE('VALUE OF A,B = ?');
  READLN(A,B);
  BLAH(A,C);
  STUFF(B,C,D);
  WRITELN('VALUE OF PARAMETER = ',D);
END.
```

The USES statement calls up various inbuilt library units to do standard operations. DOS is a unit that allows you to invoke dos calls such as GETTIME, SETTIME, DISKSIZE, etc. Unit CRT manipulates text screens and controls sound while PRINTER enables you to output to the printer.

GRAPH is for Hercules to VGA graphics functions.

Declaration VAR defines the global variables you are to use and their types. The PROCEDURE statement allows you to group a sequence of operations into one self contained sub program. When declaring a procedure, local incoming variables are declared first, together with their types and then the outgoing variables, preceded by the word VAR.

Pascal also has the capability of being extended. For example if your version doesn't have a complex maths library unit, it is fairly easy to write the unit that performs the complex maths. When it is compiled, you have all the types and operations available in your program that were declared in the unit. You simply add COMPLEXMATH to the USES clause at the top of the program.

It is also possible to put assember statements into a procedure, as in the kernel listing, so that hardware can be directly controlled.

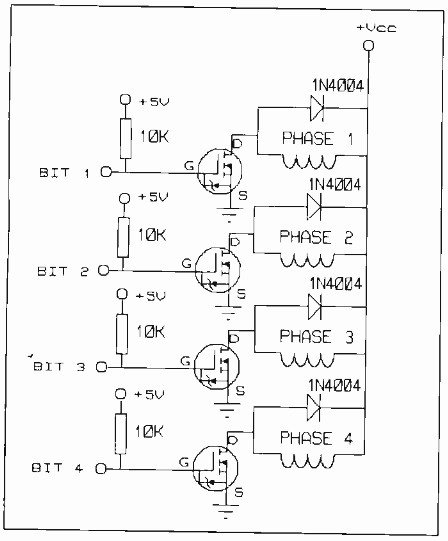


Fig. 4. High-power stepper motor controller incorporating a mosfet driver. Supply voltage to the motor windings can be more than 5V but not more than 60V using, say, a VN66AF power mosfet.

Listing 2. Very little software needs to be linked into the kernel routine in order to read an a-to-d converter.

```
BEGIN
  INIT_COM;
  FOR K := 1 TO 1000 DO              {K and DATA previously defined}
  BEGIN                              {as integer and byte types     }
SEND_BYTE(0);
    GET_BYTE;
    DATA := BYTE_RCVD;
    {DO SOMETHING TO DATA}
  END;
  QUIT_COM;
END.
```

Listing 3(a). Sending information to a d-to-a converter provides waveform synthesis for driving, say, the loudspeaker of Fig. 3.

```
BEGIN
  INIT_COM
  REPEAT
    DATA := 0   ; SEND_BYTE(DATA); {To transmit complex waveforms just}
    DATA := 255 ; SEND_BYTE(DATA); {send a different sequence of data.}
  UNTIL KEYPRESSED = TRUE;         {But the more samples that are sent}
  QUIT_COM;                        {will mean a lower base frequency  }
END.
```

Listing 3(b). This routine runs a motor up to full speed, holds it there a while then decelerates it. It can be used with the d-to-a converter of Fig. 3 by replacing the loudspeaker with a small dc motor.

```
BEGIN
  INIT_COM
  FOR K := 0 TO 255 DO
  BEGIN
    SEND_BYTE(K);
    DELAY(300);                    {DELAY FOR 300ms}
  END;
  DELAY(2000);                     {HOLD AT MAX SPEED FOR 2 SECONDS}
  FOR K := 255 DOWNTO 0 DO
  BEGIN
    SEND_BYTE(K);
    DELAY(300);
  END;
  QUIT_COM;
END.
```

Listing 4. Driving a four-phase stepping motor involves a ring-counter sequence.

```
BEGIN
  A := 12;                         {A,B,C,D previously defined as
BYTE}
  B := 6;
  C := 3;
  D := 9;
  REPEAT
    SEND_BYTE(A); DELAY(300);      {Delay for 300ms between each step}
    SEND_BYTE(B); DELAY(300);
    SEND_BYTE(C); DELAY(300);
    SEND_BYTE(D); DELAY(300);
  UNTIL KEYPRESSED = TRUE;
END.
```

Image acquisition

With a dot-matrix printer that can emulate the standard IBM printer graphics mode, it is possible to make the printhead scan across any hard-copy image in the printer's paper path. Most dot matrix printers can emulate the IBM graphics mode.

By attaching one end of an optical fibre to the print head and the other end to the opto-transistor shown in Fig. 2, it is possible to read a bit-map of the image passing the print head into the PC. An example of the reproduction obtainable via this method is the photograph of Bogart on the front cover of the magazine.

A light source is needed for this application, usually positioned about 300 to 600mm away and to the side of the printer. **Listing** 5 shows the procedure that controls the printer and acquires the data from the a-to-d converter.

Because the data length of the printer in the selected mode is 960 bytes and resolution of a VGA screen is 640-pixel, line-scan is not 100% straightforward. One byte of data is sent to the printer on the odd count and two bytes on the even count. After each data transmission moves the head, the image light level is sampled.

Resolution attainable is approximately 80 dots per inch. Epson and IBM type printers also have a 480-byte line mode but it produces very poor scans. The software that does the scanning redefines the VGA palette to 16 grey shades and brackets the 256 levels that come from the a-to-d to one of these shades. Images are saved in standard paintbrush (.PCX) format with palette information.

Windows quite happily accepts the image, as long as it is in a 256 colour mode or higher. If in sixteen colour mode, Windows will remap the incoming palette to its own colour set and the result will make you want to go for a quiet lie down somewhere.

Due to the large 1mm diameter of the optical fibre – the cheapest I could find – sampling problem occurs due to aliasing. To be precise, the problem is spacial aliasing caused by the relatively large diameter of the optical fibre compared to the resolution of the printer.

As the fibre scans across an image, the sample points are slightly overlapped which results in some of the data from one sample point appearing in the next. This means that dark striations appear on the scanned image. However with the use of some filtering software the interference can be removed. A line-array ccd could produce much better results.

Temperature measurement

By using a diode – *1N4148* will do – and a 1kΩ series resistor instead of the opto-transistor and 3kΩ resistor of Fig. 2, a reasonably

Continued on page 458

Listing 5. Image scanning with the aid of a dot-matrix printer involves moving the head across the image and sampling light level at each increment. In the variable declaration, stuff is declared as type TEXT; later on in the program assign it to the printer port with ASSIGN(STUFF,'LPT1').

```
PROCEDURE SCAN;          {FULL PAGE SCAN ; EQUIV TO APPROX 80 DPI}
BEGIN WRITE(STUFF,#27,#60);      {RESET PRINTER                    }
 WRITE(STUFF,#27,#85,#1);     {UNIDIRECTIONAL PRINTING ON        }
 FOR Y := 0 TO 479 DO
 BEGIN
  WRITE(STUFF,#27,#51,#03);  {MOVE PAPER BY 3/216 INCH          }
  WRITE(STUFF,#27,#76,#192,#3);{SLOW MODE = #76:960 DOTS/SCAN RES}
  FOR X := 0 TO 639 DO
  BEGIN
   IF ODD(X) THEN
   BEGIN
    WRITE(STUFF,#00);
    DATAGET;         {DATAGET IS A DATA ACQUISITION PROCEDURE  }
   END          {CONTROLLING THE A-TO-D CONVERTER AND PLOTS}
   ELSE          {THE DATA ONTO A VGA SCREEN          }
   BEGIN
    WRITE(STUFF,#00,#00);
    DATAGET;
   END;
  IF KEYPRESSED = TRUE THEN
   BEGIN
    X := 639;
    Y := 479;
   END;
  END;
 END;
END;
```

Listing 6. Routines for plotting temperature against time.

```
PROCEDURE SETUPGRAPHICS;
BEGIN
 GD := DETECT; INITGRAPH(GD,GM,''); {INITIALISE GRAPHICS SYSTEM}
 IF GRAPHRESULT <> GROK THEN      {ALL SCREEN OUTPUT ASSUMES THAT}
  HALT(1);              {A VGA ADAPTER IS PRESENT}
 SETVIEWPORT(0,0,GETMAXX,GETMAXY,CLIPON);
 SETWRITEMODE(ORPUT);
 SETCOLOR(15);
 RECTANGLE(0,0,639,240);
 LINE(0,15,639,15);
 SETTEXTJUSTIFY(LEFTTEXT,CENTERTEXT);
 SETTEXTSTYLE(DEFAULTFONT,HORIZDIR,1);
 OUTTEXTXY(240,09,'TEMPERATURE MEASUREMENT'); {PUT TITLE AT TOP OF
SCREEN}
 K := 0;
 REPEAT
  LINE(K,15,K,240);
  STR(K*30,SCREENTEXT);        {SCREENTEXT DEFINED AS TYPE STRING}
  OUTTEXTXY(K+2,235,SCREENTEXT+'ms'); {PUT TIMING TICKS ON X AXIS}
  INC(K,64);
 UNTIL (K>639);
END;

BEGIN
 SETUPGRAPHICS;
 INIT_COM2;
 FOR K := 1 TO GETMAXX DO
 BEGIN
  SEND_BYTE(0);             {START CONVERSION}
  GET_BYTE;               {GET THE DATA}
  PUTPIXEL(K,3*BYTE_RCVD-200,15);   {PLOT DATA WITH SCALING AND
OFFSET}
  DELAY(30);              {WAIT FOR 30mS}
 END;
 READLN;
 CLOSEGRAPH;
END.
```

ADC0804 a-to-d converter

This converter is suitable for both microprocessor bus based systems and stand-alone a-to-d applications. Its conversion time of 100ms limits its use to audio frequencies and below.

Low speed is offset by the device's cheapness – at around £4.25 – and ease of use. Its on-board clock generator needs only two external components. As you can see from Fig. 1, driving the device is straightforward.

With a single 5V supply and 2.5V applied to $V_{ref}/2$, the analogue input voltage range is 0 to 5V. Fig. 2 uses a forward biased diode to provide the reference voltage of 0.7V which leads to an input voltage range of 0 to 1.4V.

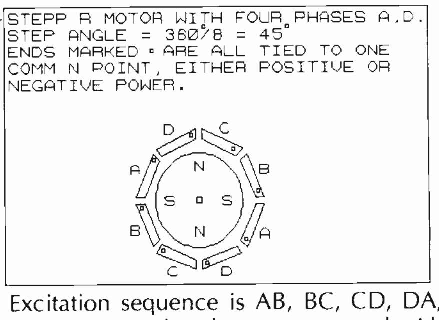
Voltage reference $V_{ref}/2$ can be any value under 2.5V. As a result, any voltage span can be converted with a full eight bits of resolution. Also, by connecting $-V_{in}$ to a voltage other than ground (note the 1kΩ pot in Fig. 2) the voltage conversion range need not start at 0V.

Stepping motor background

Consider the simplified diagram of the typical stepping motor shown below with windings A to D. The armature is a cylindrical magnet with poles N and S.

Applying a voltage to phases A and B results in the poles of the motor armature aligning with the two phases in an orientation depending on the polarity of the applied signal.

If phase A is switched off and phase C switched on while B is still on, the motor armature realigns itself with the new phase. Similarly if B is switched off and D on, the armature again aligns itself with the new phase. In this way the motor rotates.



Excitation sequence is AB, BC, CD, DA, AB. In practice the phases are wound with higher multiples of each other so that the step angle is much less. A typical example is 200 steps per revolution, which is equivalent to 1.8° per step.

It is possible to double the resolution of a stepping motor by half stepping it. Instead of exciting it as in the above sequence, if the following sequence A, AB, B, BC, C, CD, D, DA, A is fed to the motor, then 400 steps per revolution will result.

Stepping motors are found in plotters, floppy disk drives, hard disk drives, pick and place machinery and sewing machines. They are extensively used in industry, particularly for robotic manipulators.

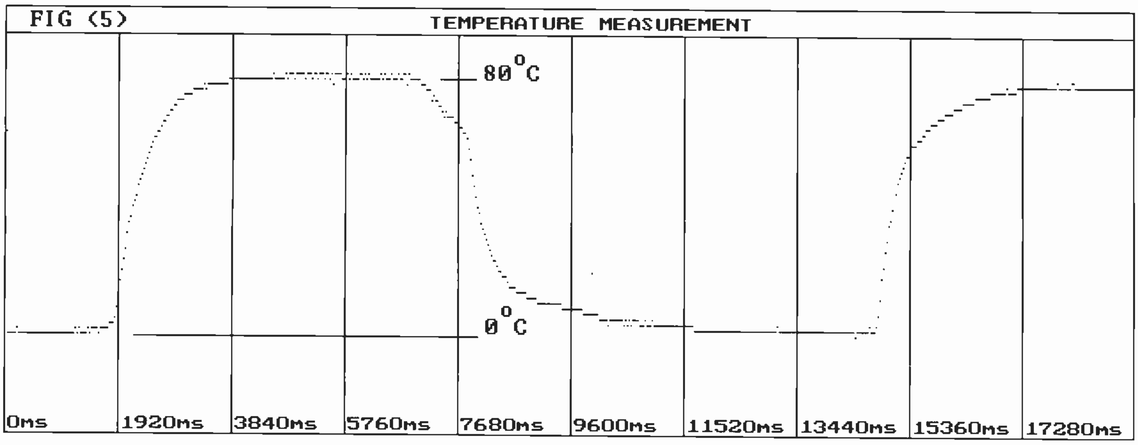


Fig. 5. Simply replacing the light sensor of Fig. 1 with a general purpose diode allows temperature to be read into the PC.

accurate measurement of temperature can be made. The configuration is the same as the diode resistor combination that provides the 0.7V reference.

This arrangement ignores the non-linear temperature dependance of the pn junction but at 5p each as opposed to £1.20 for the *LM335Z* or £3.45 for the *LM35DZ*, a degree of non-linearity is a small penalty. Note that the temperature sensing diode is followed by a gain block to amplify its approximate –2mV/°C to a level that the a-to-d converter can handle.

You may have noticed that the circuit uses a diode to provide a reference voltage to the a-to-d converter. This will always be at the local ambient temperature, so its voltage variation is going to be relatively small compared to the amplified voltage of the temperature sensing diode.

Listing 6 shows routines for plotting temperature against time. **Figure 5** is a typical temperature plot against time. The 0° point was measured by putting the diode into melting ice, while the 80° point was obtained by putting the diode into a convenient cup of hot tea.

For a faster data acquisition rate, the DELAY(30) (30ms delay) command can be removed from the listing. ■

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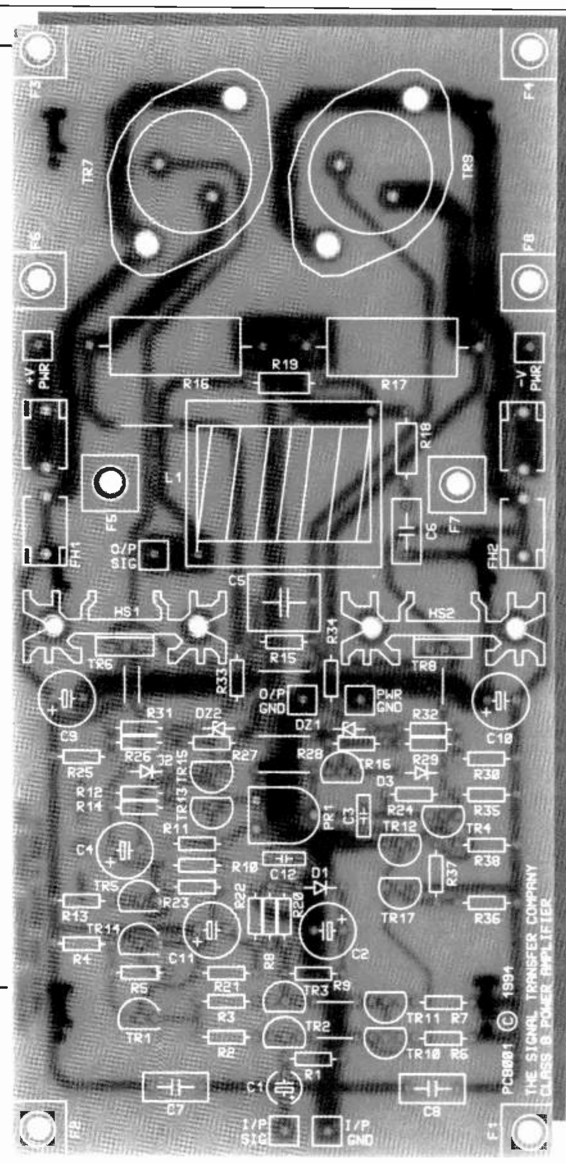

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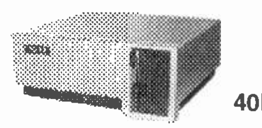

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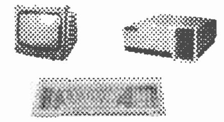

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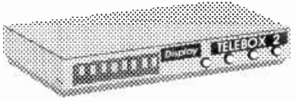

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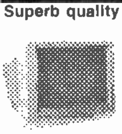

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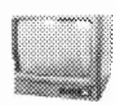

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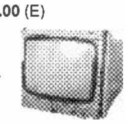

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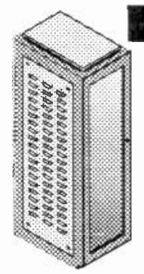

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Racal/Dana signal generator 9082H – 1.5-520Mc/s – £600.
Tektronix – 7S14 – 7T11 – 7S11 – 7S12 – S1 – S2 – S39 – S47 – S51 – S52 – S53 – 7M11.
Marconi mod meters type TF2304 – £250.
HP 5065A rubidrum vapour FX standard – £2.5k.
Systron Donner counter type 6054B – 20Mc/s – 24GHz – LED readout – £1k.
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Marconi TF2091 noise generator. A, B or C plus filters – £100-£350.
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Wiltron 610D Sweep Generator + 61084D PI – 1Mc/s – 1500Mc/s – £500.
Time Electronics 9814 Voltage calibrator – £750.
Time Electronics 9811 Programmable resistance – £600.
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PC EVOLUTION

From humble beginnings, the PC has grown into one of the world's most widely used office machines. David Guest encapsulates the PC's history and takes a brief look at its architecture.

Early computers – mainframes – interfaced to numerous terminals comprising a keyboard and monitor. Combined with a multitasking operating system, this allowed a number of users to access the same computer simultaneously. Only large companies and research establishments could justify the high capital and subsequent maintenance costs.

In 1971 Intel released its first microprocessor – the 4-bit *4004* – developed for a calculator company called Busicom.

Further development resulted in the first 8-bit processor, the *8008*, which ultimately led to the manufacture of the well known *8080* processor in 1973.

Initially, the *8080* was used for video games. It eventually devolved into a computer available in kit form only – the Altair – with a hex keypad and LED display. It was not until two further breakthroughs that the *8080* microprocessor found its way into business applications. These were development of the CP/M operating system by Gary Kildall, founder of Digital Research, and invention of the floppy disk for data storage by John Torode.

Computing machines adapted quickly, as did spreadsheet and word processing type software. By now, other microprocessor manufacturers had introduced their own devices. The most common of these was the Motorola 6800, adopted for Apple's range of computers.

From 1976, a vast variety of computers were introduced, the main core being aimed primarily at video games for the domestic market. Due to their lack of serious software however, CP/M based systems continued to dominate the business sector.

In 1979, Intel brought out the first 16-bit processor – the *8086*. This device could handle 16-bit data but, more significantly, it could address up to 1Mbyte of memory. This overcame the serious restrictions of the 64Kbyte limit imposed by existing 8-bit alternatives. Due to manufacturing problems and cost considerations, the *8086* was modified to support an 8-bit external data bus. Called the *8088*, the new device maintained a true 16-bit core and 1Mbyte addressing capability.

At this time Motorola had not yet developed its 16-bit *68000* series microprocessors. Intel, as a result, had a market edge, which ultimately led to IBM selecting the Intel processor and associated support chips for the IBM PC, launched in 1981.

The disk operating system associated with the PC – originally PC-DOS – rapidly became an industry standard. In the same way that CP/M systems had brought about the standardisation of 8in floppy-disk drives, DOS caused the standardisation of their 5.25in successors. The overnight success of the PC surprised everyone, even IBM, spurring them to release the PC-XT at the end of 1982. As standard, the XT had 64Kbyte of memory and a 10Mbyte hard disk drive, providing rapid access to frequently used software.

Subsequent release of the PC-jr, more commonly known as the Peanut, saw IBM fail in an attempt to gain access to the small business and domestic market. Poor quality keyboard and disk drive problems, together with a high cost compared to other machines in the sector, caused the Peanut's withdrawal from the market in 1985.

Released in 1984, the high end PC-AT was more successful. It performed between five and six times faster than the PC-XT, and had 512Kbyte of memory and a 20Mbyte hard disk drive as standard. Contrary to IBM's belief that the PC-AT would be used to run multitasking UNIX based operating systems, it was bought mainly as a means of running DOS software more quickly.

Popularity of the PC attracted other manufacturers. They responded by producing cheaper PCs – or clones – with improved specifications. The first of these to gain to wide acceptance was the Olivetti M24.

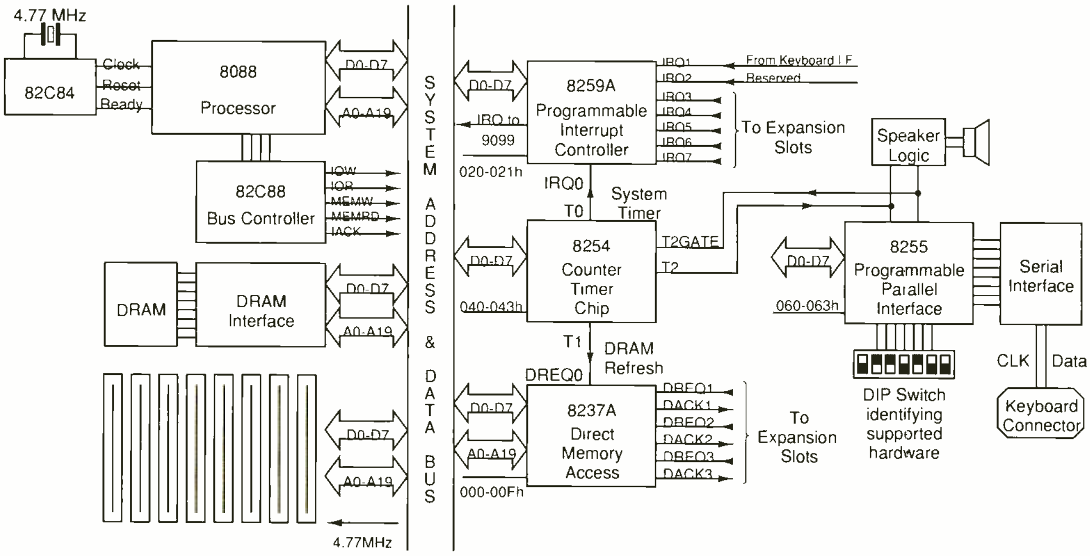


Fig. 1. In the original PC XT the 8088 processor was 16-bit internally but had an 8-bit data external bus. It could however address a megabyte of memory directly.

In 1986, Intel launched the 32-bit *80386DX* featuring a 4Gbyte memory addressing capability. It succeeded where the *80286* failed in that is supported a virtual real mode of operation. This meant that dos applications could be run in any 1Mbyte window of the system memory. Multitasking software, such as Windows, could now run dos based programs, and became a popular choice for users of the *80386* based systems.

A cheaper version of the *80386* was soon released, namely the *80386SX*. This had a 16-bit external data bus while supporting the full 32-bit core of the *80386DX*. Due to the growing popularity of software and the falling price of hardware, *386SX-based* computers quickly took the lead in the early 1990s.

Intel continued to improve its technology and developed an optimised *386DX* processor. Called the *80486DX*, it combined the main processor, memory cacheing and a maths coprocessor on a single chip. This device has found its way into the high end of the market, used for CAD and similar systems.

Building on the performance of the *80486*, Intel recently released the Pentium. This chip has received mixed reviews in the press. It requires recompilation of existing PC software source code to take full advantage of its enhanced capabilities.

Inside the PC

As illustrated in **Fig. 1**, the original PC-XT architecture was built around Intel's support chips developed for *8086/8088* based processor systems. Most of these devices are programmable, each supporting a specific role within the PC architecture.

Clocking for the processor is derived from the *82C84*, which also handles system resetting and synchronisation of the processor's handshake signal. All bus-control signals are handled by the *82C88*.

There were no facilities on the processor for performing tasks at regular intervals, such as refreshing the systems dram and updating the system clock. As a result, designers used Intel's *8254* counter-timer chip. This device contains three independently programmable timers, each capable of producing an output signal at predefined intervals.

Timer zero of the *8254* generates eight system timer interrupts via an *8259A* programmable interrupt controller. These maintain the system clock, keyboard and i/o slot cards. Direct memory access is under control of timer one. An additional three interrupts are produced by the DMA chip for use by i/o

Table 1. I/o map for the XT and AT.

I/o address (hex)	Usage
000-00F	8237 DMA #1
020-021	8259 PIC #1
040-043	8253 CTC
060-063	8255 PIO (XT)
060-067	8742 Keyboard I/F (AT)
070-071	CMOS RAM & NMI Mask (AT)
080-08F	DMA Page Registers
0A0-0A1	8259 PIC #2 (AT)
0A0-0AF	NMI Mask Register (XT)
0C0-0DF	8237 DMA #2 (AT)
0F0-0FF	80287 Coprocessor (AT)
1F0-1FF	Hard disk (AT)
200-20F	Game/Control
210-21F	Expansion Unit (XT)
238-23B	Bus Mouse
23C-23F	Alt Bus Mouse
278-27F	Parallel Printer (LPT3)
2B0-2BF	EGA
2C0-2CF	EGA
2D0-2DF	EGA
2E0-2E7	GPIB (AT)
2E8-2EF	Serial Port (COM4)
2F8-2FF	Serial Port (COM2)
300-30F	Prototype Card
310-31F	Prototype Card
320-32F	Hard Disk (XT)
378-37F	Parallel Printer (LPT2)
380-38F	SDLC
3A0-3AF	SDLC
3B0-3BB	MDA
3BC-3BF	Parallel Printer (LPT1)
3C0-3CF	EGA
3D0-3DF	CGA
3E8-3EF	Serial Port (COM3)
3F0-3F7	Floppy Disk
3F8-3FF	Serial Port (COM1)

Extended memory, XMS

Launch of the AT coincided with the introduction of processors capable of accessing memory above 1Mbyte. However, the so-called 'extended' memory could only be accessed with the processor running in protected mode.

Extended memory area was not maintained by DOS, resulting in the introduction of the extended memory driver HIMEM.SYS. This driver gave a well defined interface to DOS based software, providing access to memory above the 1Mbyte boundary.

In addition, HIMEM.SYS controls gating of the A_{20} address line, providing DOS based software with access to the first 64Kbyte of extended memory as part of the real-mode memory map. As a result, there is access to 1098Kbyte even though1Mbyte is only 1024Kbyte.

The extra memory space is achieved by a quirk of the *8086* family address mechanism whereby addresses are offset by a segment register. If the register is loaded with values above $FFFF_{16}$, the resulting memory address exceeds the 1Mbyte limit.

External hardware is needed to gate the processor A_{20} signal, controlled by HIMEM.SYS. For *80386* and later processors, the expanded memory driver uses the HIMEM.SYS driver to allocate system memory above 1Kbyte for use as expanded memory.

slots. Interrupts for the *8255A* programmable parallel interface are derived from timer two. This chip adapts counter two output to produce audio output from the speaker. The parallel interface chip also connects to the keyboard and gives the processor access to a range of status and control lines in the system.

Interfaces to communications ports and storage media, such as floppy and hard disk drives, had well defined positions within the architecture. They had specific i/o addresses, **Table 1**, and interrupt or DMA channels **Table 2**. Since only the lower ten address lines on the original peripheral cards were decoded, the address map has become overcrowded. This forces i/o devices to reside exclusively in the first 1Kbyte of i/o space, instead of in the full 64Kbyte supported by the processor.

Interfacing to the graphics display has evolved alongside display adapters. Features and resolutions have been extended while maintaining compatibility with the original monochrome display adapter, MDA, add-on card released with the original PC.

The first graphics card was the Hercules. It added 720 by 350 pixel graphics on to the 80-column by 25-line text only potential of the MDA card. Soon after came the colour-graphics adapter, or CGA standard, which supported two colours with a resolution of 640 by 200 pixels, or four colours at a resolution of 320 by 200.

Graphics resolution was further extended with the enhanced graphics adapter, or EGA, which supported 16 colours at resolutions up to 640 by 350. It is now superseded by the video graphics array standard, VGA, offering up to 720 by 400 pixels in text mode or 16 colours at 640 by 480 for graphics. Super VGA extends this capability even further and state-of-the-art PC graphics cards are now capable of 1280 by 1024 pixels with 16.7 million colours.

AT architecture

Architecture of the AT contained features of

Continued on page 465

History of dos

Written by Tim Patterson for Seattle Computer Products in 1980, dos has become the most widely used primary operating system for microcomputers. Originally called 86_DOS, it was intended to simplify porting of applications from CP/M, produced by Digital Research, which was then the leading microcomputer operating system.

Although only a few people knew of its existence, they were all impressed with it, and SCP acquired several dozen customers. Some months later Microsoft, acting secretly for IBM, commandeered SCP for a customised version. To effect this, Tim Patterson left SCP to join Microsoft in April 1981.

Consequently, in August 1981, IBM brought out its Personal Computer with Microsoft following suit with MS DOS 1.0. Microsoft went from strength to strength thanks to its established and continued reputation for producing well-supported business programmes, establishing DOS as the industry standard operating system.

Table 2. Hardware interrupts and dma channels.

IRQ Level (XT and AT)	Use
NMI	Parity
IRQ0	System Timer
IRQ1	Keyboard
IRQ2	Cascade for IRQ8-15
IRQ3	COM2
IRQ4	COM1
IRQ5	Hard disk(XT)/LPT2(AT)
IRQ6	Floppy Disk
IRQ7	LPT1
(AT only)	
IRQ8	Real Time Clock
IRQ9	Re-directed to IRQ2
IRQ10	Unassigned
IRQ11	Unassigned
IRQ12	Unassigned
IRQ13	80287 Co-processor
IRQ14	Hard Disk
IRQ15	Unassigned
DMA channel (XT and AT)	**Use, 8-bit transfers**
DMA0	Memory Refresh
DMA1	SDLC
DMA2	Floppy Disk
DMA3	Unassigned
DMA channel (AT only)	**Use, 16-bit transfers**
DMA4	Unassigned
DMA5	Unassigned
DMA6	Unassigned
DMA7	Unassigned

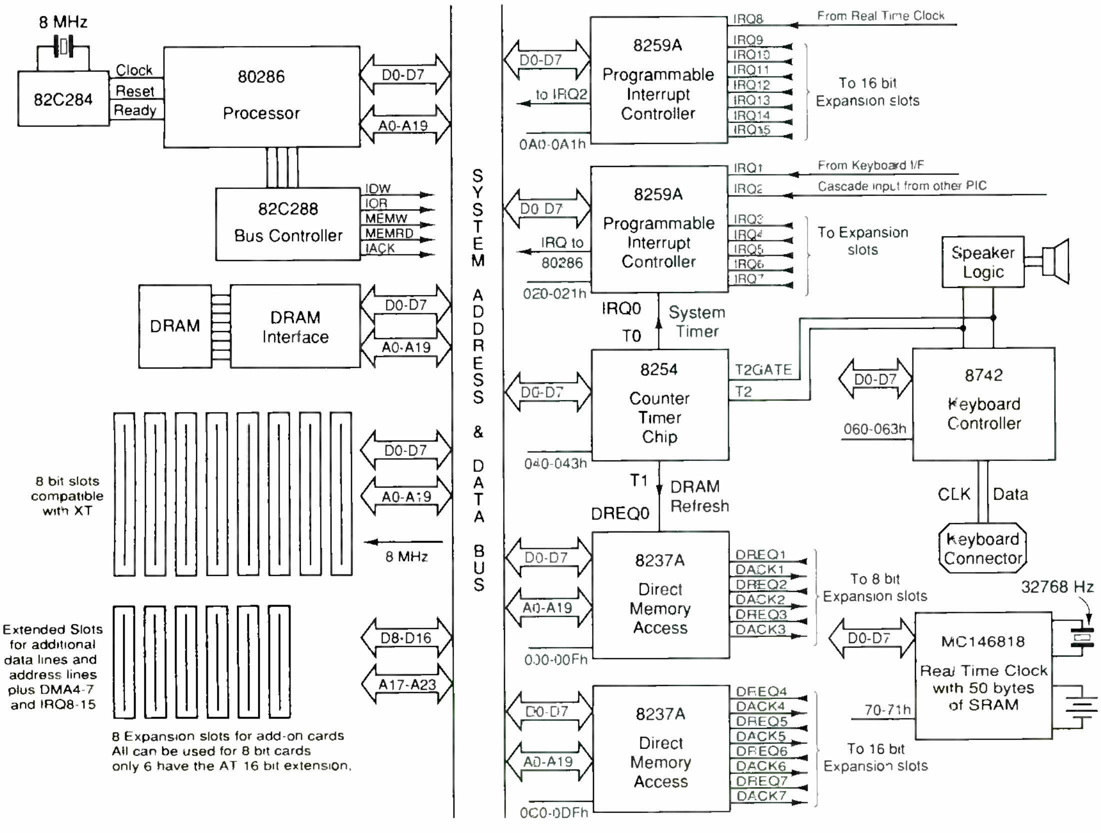


Fig. 2. A significant advance on the XT, the PC AT had a fully 16-bit processor and additional DMA capability. Unlike the XT, the AT also had the ability to date-stamp files from real-time clock information.

PC memory map

To avoid the dependency of software on hardware configurations, a standardised basic i/o system, or bios, was developed. In attempts to speed up their programs however, many writers produced software that communicated directly with i/o, bypassing the bios. As a result, future PC designs had to maintain hardware compatibility to maximise the usefulness of existing software.

In the original PC, 640Kbyte of memory was reserved for use by the operating system and programs while the area between 640Kbyte and 1Mbyte catered for the video adapter and other add-on cards. Most *80386* based computers support a minimum of 1Mbyte of memory in the area above 640K. This memory occupies the same address space as the bios roms and video memory.

Originally the bios and video memory area was only used for shadowing the bios roms to improve system performance. Software, like QEMM by Quarterdeck, and more recently DOS 5, makes this area of memory available to DOS programs, freeing up the valuable 640K of DOS memory for use by software applications.

Using a technique that pages out and relocates memory interactively, QEMM goes a step further than DOS. Even with DOS taking advantage of High Memory and the Upper Memory Area, it still takes memory in the lower 640Kbyte of memory space, **Table 3**.

The first kilobyte of memory stores the interrupt vector table for redirecting hardware interrupts to the necessary software routines. The following 256 bytes is the BIOS data area that provides information on the system architecture and parameters related to BIOS calls, **Table 4**. DOS reserves the next 512 bytes, called the DOS communication area, for its own use. Immediately following this is the DOS kernel, i.e. the hidden IO.SYS and MSDOS.SYS files, and the internal DOS drivers. Then come the drivers that relate to entries in the CONFIG.SYS file. Device drivers appear first, followed by memory areas for file handles, disk buffers, stacks and so on. Remaining memory is used by programs, the first always being the part of COMMAND.COM, known as the resident portion. Remaining space is filled by terminate and stay resident programs, or TSRs, that always remain in memory and by any programme being executed.

Extended memory	Top of memory to
High memory area (64Kbytes)	100000h
System bios	FFFFFh to F0000h
Rom expansion (XT)	EFFFFh to E0000h
Expanded memory page frame	DFFFFh to D0000h
XT hard disc bios	CFFFFh to C8000h
EGA/VGA bios	C7FFFh to C0000h
Video memory	BFFFFh to A0000h
Available memory	9FFFFh
Resident portion of command.com	to
DOS config.sys entries	
DOS kernel and internal device drivers	00600h
DOS communication area	005FFh
Rom-bios data area (256 bytes)	
Bios stack/user vectors (256 bytes)	to
User vectors (288 bytes)	
Interrupt vectors (480 bytes)	0000h

Table 3. The computer's memory map.

Expanded memory, EMS

When the PC became available, memory access capability rose from the 64Kbyte possible with 8-bit microprocessors to a relatively massive 1Mbyte. However, this rapidly became inadequate due to the increasing complexity of software and its associated data.

In response, Lotus, Intel and Microsoft initiated a joint venture to overcome the 1Mbyte limit imposed by PC-XT 8088/8086 processors. They introduced the expanded memory specification called LIM EMS 3.0.

This standard provided DOS with up to 8Mbyte of additional memory, resident on a plug-in i/o card. The memory was organised as banks of 16Kbyte pages. Four of these could reside in the DOS memory map at a time, typically from D0000 to DFFFF. Banks could be selected via hardware registers, accessed via an expanded memory driver.

Development of the AT computer with its 16Mbyte of memory was expected to see the demise of the requirement for expanded memory. Loyalty to DOS by most PC users however meant that this extra memory could not be accessed by programs.

Software running under DOS cannot take advantage of the memory management unit made available in the later processors. They only execute in real mode, preventing access to memory beyond the1Mbyte boundary. For this reason EMS was still extensively used; however, introduction of the *80386* processor with its sophisticated internal memory management unit removed the need for external hardware for bank switching.

In revision 4.0, the EMM system driver was redefined to take advantage of bank switching. It made use of system memory above 1Mbyte, allowing DOS based programs access to extended memory. The driver's implementation of extended memory makes it dependent on the extended memory driver, HIMEM.SYS. This driver must always be loaded prior to the expanded memory driver, EMM386.EXE.

Version 5 of DOS also needs the expanded memory driver in order to make use of the upper memory area between 640Kbyte and 1Mbyte. This area is normally associated with the system bios and the video driver.

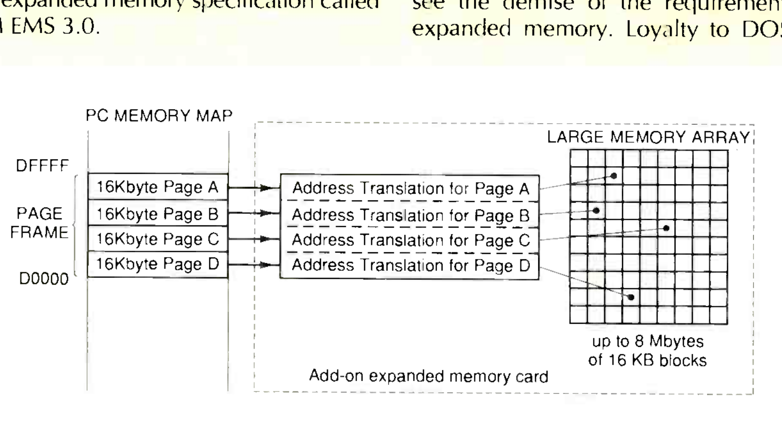


Fig. 3. Allowing up to 8Mbyte of paged memory, expanded memory concept was developed to counter the 1Mbyte addressing limitation of earlier PCs.

Table 4. PC interrupt vectors and their uses.

Interrupt	No Purpose	Interrupt	No Purpose
00H	Hardware divide by zero	1DH	Video parameter table1EH
01H	Hardware single step strap	1EH	Disk parameter table
02H	Nonmaskable interrupt	1FH	Graphics character table
03H	Debugger breakpoint set	20H	DOS terminate program
04H	Arithmetic overflow	21H	DOS system services
05H	BIOS print screen	22H	Program terminate
08H	IRQ0 clock tick	23H	DOS CTRL-BREAK
09H	IRQ1 keyboard	24H	DOS critical-error handler
0AH	IRQ2	25H	DOS disk read
OBH	IRQ3 COM2	26H	DOS disk write
0CH	IRQ4 COM1	26H	DOS terminate resident
0DH	IRQ5 PC/XT hard disk or PC/AT	27H	DOS idle
	LPT2	28H	DOS fast putchar
OEH	IRQ6 Diskette	29H	MS-NET services
0FH	IRQ7 LPT1	2AH	DOS primary shell program loader
10H	BIOS video services	2EH	DOS multiplex interrupt
11H	BIOS equipment list service	2FH	Mouse services
12H	BIOS memory size service	33H	Diskette vector if hard disk present
13H	BIOS disk services	40H	Hard disk parameter table
14H	BIOS communication services	41H	EGA BIOS redirection
15H	BIOS miscellaneous services	42H	EGA parameter table
16H	BIOS keyboard services	43H	EGA character table
17H	BIOS printer services	44H	PC/AT Int 70H alarm
18H	Invoke ROM-BASIC	5CH	NetBIOS services
19H	System reboot	67H	EMS services
1AH	BIOS time-of-day services	70H	IRQ8 PC/AT real-time clock
1BH	CTRL-BREAK handler	71H	IRQ9 PC/AT redirect of Int 0AH
1CH	Called by Int 08H handler	75H	IRQ13 PC/AT match coprocessor

the original PC-XT and supported most of its software, **Fig 2**. Additionally, the AT had a further DMA controller and an extra *8259A* interrupt controller for supporting more i/o slots.

File date stamping, introduced with DOS V1.1, brought about the need for the computer to maintain clock and calendar information even after power down. A Motorola real time clock, the *MC146818*, carried out this task and provided an additional 50byte battery-backed ram. This memory was used by the AT basic input/output system, or BIOS, to store the hardware parameters, circumventing the need for programming DIP switches.

Also introduced in the AT was Intel's *8742* microcontroller, used to interface to an upgraded 101-key keyboard. This chip could transmit data to the keyboard to control its new functions and status indicator LEDs. Additional i/o lines on the *8742* controller replaced the remaining functions provided by the *8255* in the XT, removing it from the AT architecture. ■

Further reading

Dettman, T & Kyle, J, *DOS programmer's reference* (2nd edn), Que Corporation and Precision Printing, Carmel, 1989.

Jamsa, K, *DOS programming : The complete reference*, Osborne McGraw-Hill, California, 1991.

Norton, P, *Peter Norton inside the IBM PC*, Simon and Schuster Inc, New York, 1986.

ONE-TO-ONE RADIOCOMMS

To contact someone on the telephone network, you dial their telephone number. With mobile radio, that initial contact is not so easy. Signalling protocols have evolved which allow selective calling. Communications engineer James Vincent* explains the design considerations.

**James Vincent works for a major aerospace company*

The simplest way of trying to contact a specific radio user is to make the initial call on either the channel in use or on a predetermined calling channel. This procedure is fine if the frequency is relatively clear and the users well disciplined. However on a crowded radio channel even radio discipline cannot ensure that the initial call will get through.

With the increasing demand for radio spectrum, techniques such as digital modulation, spread spectrum and amplitude companded single sideband are being employed to increase channel capacity. Unfortunately the majority of these techniques are not backward compatible with the hardware already in use, such as conventional narrowband am and fm systems. Often the considerable investment in the system infrastructure limits the options available. The simplest approach to increasing the efficiency of a radio communication system is to use special signals to call selectively a particular user or user group. Such techniques are called 'selective calling' or selcall.

Selcall techniques

Single toneburst. The simplest selective call system is based around a single audio tone which, when transmitted, causes a tone detector in the receiver to unmute and enable the audio output (this can be as simple as connecting the loudspeaker). Although simple, it carries significant disadvantages: Only tones within the audio passband of the radio system can be used. This limits the number of tones which can be used and hence the number of stations that can be called; Most tone decoders can only reliably decode tones which have an adequate frequency separation. This results in the creation of guard bands around each tone frequency. This further limits the number of tones which can be used and reduces the number of stations (addresses) which can be called.

A typical example of a single toneburst system is the 1750Hz, 600ms toneburst used to access amateur VHF and UHF repeaters in the UK and parts of Europe.

Sequential single tones. An obvious way of increasing the number of addresses which can be called is by sending tones sequentially. Each tone is sent for a short duration (typically 33 to 100ms) and the number of tones sent is predetermined (frequently five tones). This sequential approach increases the number of addresses and hence stations that can be called to theoretically 100,000 for a 5-tone system.

In comparison the single toneburst systems can usually only support up to 40 or so tones within the audio passband of the radio. The CCIR, EEA and ZVEI 5 tone (see **Table 1**) selcall systems are the commonest examples of sequential tone systems.

Sequential dual tone systems. Dual tone multi-frequency (DTMF) tones are the signalling tones as used on the public telephone system. With DTMF two tone frequencies are transmitted simultaneously for typically 40ms, each tone pair representing a digit or a character in the extended DTMF tone group; 0 to 9, * , #, and A to D.

These tone pairs may be sent sequentially just as

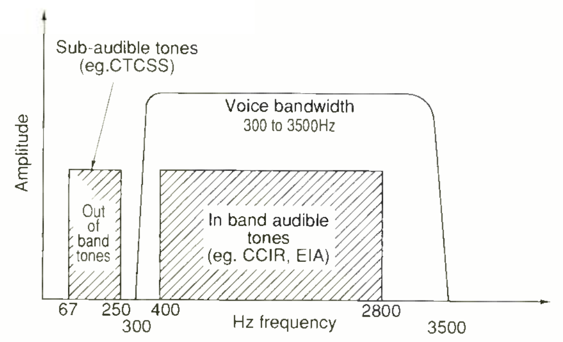


Fig. 1. Typical radio system audio spectrum showing the difference between audible in-band signalling (eg CCIR 5-tone selcall) and sub-audible out of band signalling (eg CTCSS).

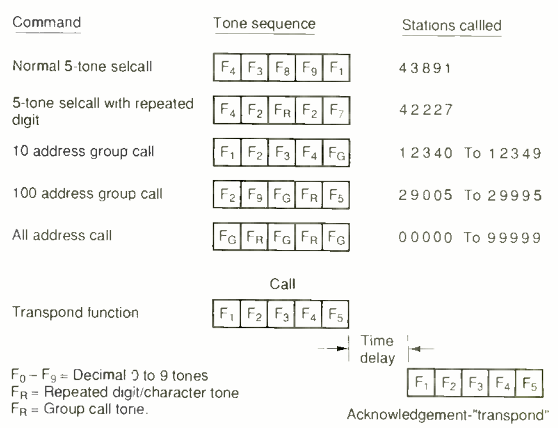


Fig. 2. Typical 5-tone selcall format indicating its flexibility and power.

in a telephone number to call a station and can provide a versatile approach to selcall. And of course the DTMF tone pairs may be used on the public telephone network during phone-patches from the radio system.

Sub-audible tone signalling. An alternative to the in voice-band signalling uses signalling frequencies outside of the radio's nominal 300-3500Hz audio passband. By using frequencies below 300Hz, a tone may be continuously transmitted allowing receiver audio output only when the sub-audible tone is present. This is in contrast to the in-band signalling systems, where the loudspeaker remains unmuted after a valid selcall tone or tone sequence allowing unwanted traffic to get through.

The sub-audible squelch system is known as continuous tone coded sub-audible squelch (CTCSS). Only receivers set to the same CTCSS tone frequency as that assigned to the transmitter will unmute on reception of a transmission. Other receivers set to different CTCSS tones will remain muted and the transmissions will not be heard. This privacy feature prompted Motorola – the inventors of the technique – to call it *Private Line* or PL.

Obviously if the user bypasses the CTCSS circuit then he will be able to hear all transmissions, even if they are not intended for the station. This by-pass feature allows monitoring for a free channel before making a transmission.

Digital selective calling. Although tone based selcall techniques were designed for analogue radio systems, digital selcall can be used on both digital and analogue systems. Digital techniques provide more addresses, higher transmission rates and error detection and correction protocols.

One such scheme is the digital selective calling (DSC), part of the Global Maritime Distress and Safety System (GMDSS). The DSC system is integral to the GMDSS concept (which becomes law by the 1st February 1999 for all ships at sea) as it provides the means of transmission and reception of distress messages between ships and coast radio stations. It also enables use as a conventional selcall system for normal voice traffic.

The DSC system is designed so that, by varying the data rate and modulation format, it can be configured for either HF or VHF marine communication systems.

Selcall in detail

With the advent of multi-tone and digital techniques, single tone in-band selcall systems have almost disappeared except for the access of repeater systems, etc. The toneburst is sent at the beginning of the initial transmission and is produced by a high stability audio oscillator using a crystal or ceramic timebase.

The usual tone detector circuit block is a phase locked loop although vibrating reeds were used on early systems. Two tone sequential (two single tones sent one after each other) selcall systems are still used in some basic communication systems such as the Citizen Band radio system in the Scandinavian countries[1].

Multiple sequential tone selcall

The sequential tone selcall systems and in particular the five tone variants are perhaps the most common selcall systems encountered in Private Mobile Radio (PMR). There are six major national and international tone-sets which are listed in Table 1.

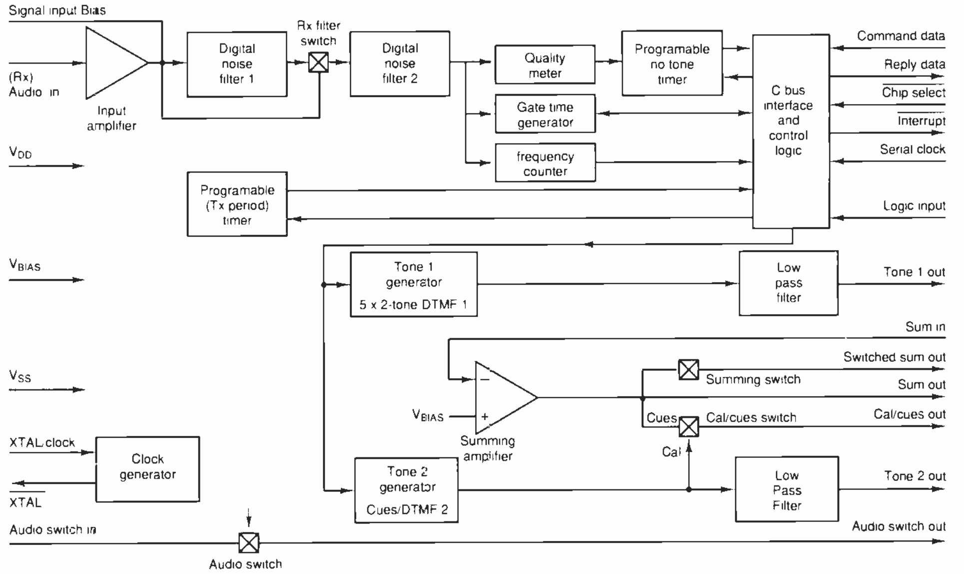
The selcall sequence usually consists of a sequence of five single tones each tone representing a specific digit of the called station's Selcall address or identity. In addition to the 0 to 9 tones there are additional tones:

Repeat tone – The repeat tone is provided to

FX803 audio signalling processor

Increased usage of mixed analogue and digital functions in asics have revolutionised approaches to selcall system design. The *FX803* is an audio signalling processor for 'in-band' tone signalling for radio systems. The *FX803* uses a non-predictive decoder and a programmable tone encoder to allow the encoding and decoding of standard or custom tone sets.

For example the device can support selcall in the following formats: CCIR; ZVEI I, II and III; EEA. In addition the chip supports two-tone selcall, single tone encoding/decoding and dual tone (DTMF) generation using either a microcontroller or a microprocessor to program the device.

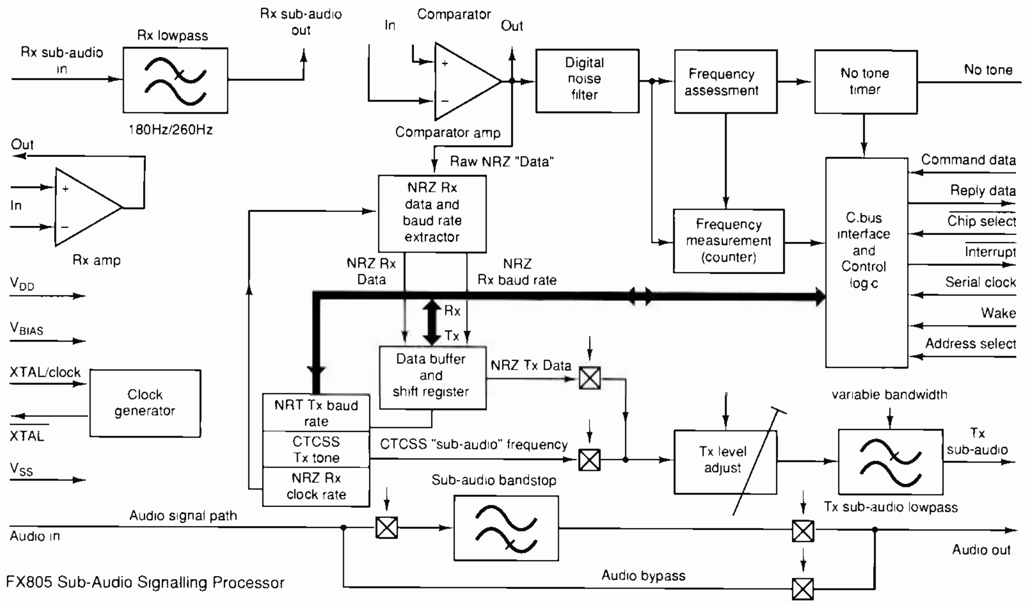


Consumer microcircuits FX803 is an advanced microprocessor controlled tone generator/decoder capable of supporting in-band selcall formats including CCIR, ZVEI, EEA, 2-tone selcall and DTMF.

Table 1

Standard	ZVEI	EIA	CCIR	EEA	CCITT	EURO
Tone Duration (ms)	70	33	100	40	100	100
Pause Duration (ms)	0	0	0	0	0	0
Code 0 (Hz)	2400	600	1981	1981	400	980
Code 1	1060	741	1124	1124	697	903
Code 2	1160	882	1197	1197	770	833
Code 3	1270	1023	1275	1275	852	767
Code 4	1400	1164	1358	1358	941	707
Code 5	1530	1305	1446	1446	1209	652
Code 6	1670	1446	1540	1540	1335	601
Code 7	1830	1587	1640	1640	1477	554
Code 8	2000	1728	1747	1747	1633	511
Code 9	2200	1869	1860	1860	1800	471
Repeat	2600	459	2110	2110	2300	1063
Alarm	970		2247	2247		
Free tone						1153
Group tone	2800	2151	2400	1055		

Note: Alarm tone frequencies can vary between manufacturers. Those shown above are for the chipsets produced by Consumer Microcircuits Ltd.
The above abbreviations stand for the following national and international standards organisations:
EEA – Electronic Engineering Association (UK).
EIA – Electronic Industries Association (USA).
CCIR – Comité Consultatif International Radio Communication
CCITT – Comité Consultatif International Télégraphique et Téléphonique
ZVEI – Zentralverband der Elektrotechnischen Industrie (German).



Consumer Microcircuits (FX805) is a microprocessor controlled integrated circuit for the encoding and non-predictive decoding of both CTCSS and CDCS (continuous continually coded squelch) formats.

FX805 sub-audio signalling processor

The *FX805* is the companion device to the *FX803* for sub-band audible and digital signalling in radio systems. This asic provides the encoding and non-predictive decoding for continuous tone coded sub-audible squelch (CTCSS) using both standard and custom tone sets. It covers digitally coded squelch (DCS/DPL). This device also requires micro control.

TABLE 2
EIA continuous tone coded sub-audible squelch tone set

67.0-XY	97.4-ZB	141.3-4A	206.5-8Z
69.3-WZ	100.0-1Z	146.2-4B	210.7-M2
71.9-XA	103.5-1A	151.4-5Z	218.1-M3
74.4-WA	107.2-1B	156.75A	225.7-M4
77.0-XB	110.9-2Z	162.2-5B	229.1-9Z
79.7-WB	114.8-2A	167.9-6Z	233.6-M5
82.5-YZ	118.8-2B	173.8-6A	241.8-M6
85.4-YA	123.0-3Z	179.9-6B	250.3-M7
88.5-YB	127.3-3A	186.2-7Z	254.1-0Z
91.5-ZZ	131.8-3B	192.8-7A	
94.8-ZA	136.5-4Z	203.5-M	

TABLE 3
Dual Tone Multi-Frequency tone set

		High Group Tones			
		H1 1209Hz	H2 1336Hz	H3 1477Hz	H4 1633Hz
Low group tones	L1 697Hz	1	2	3	A
	L2 770Hz	4	5	6	B
	L3 852Hz	7	8	9	C
	L4 941Hz	*	0	#	D

prevent a single tone transmission when sequential digits in the selcall address are identical. The utilisation of the repeat tone to indicate that a digit has been repeated avoids difficulties at the selcall decoder, particularly in determining the actual number of digits already sent.

Free tone – The free tone is used in some networks to allow determination of received signal quality and to confirm that the receiver is within the service area of the transmitter. This tone is interrupted by selcall and voice data transmissions.

Group tone – The group tone is provided for simultaneous group calling of stations in groups of 10 (four digits of selcall address common i.e. 1234X where X= 0 to 9), 100 (three digits common i.e. 456XX), 1000 (two digits common i.e. 98XXX and even 10000 (five digits common i.e. 5XXXX).

To call 10 stations whose selcall addresses begin with the same four digits, the group tone will be sent as the fifth tone i.e. 1234G will simultaneously call radios with the following selcall addresses: 12340, 12341, 12342, 12343, 12345, 12346, 12347, 12348 and 12349. This is a powerful way of calling predetermined user groups.

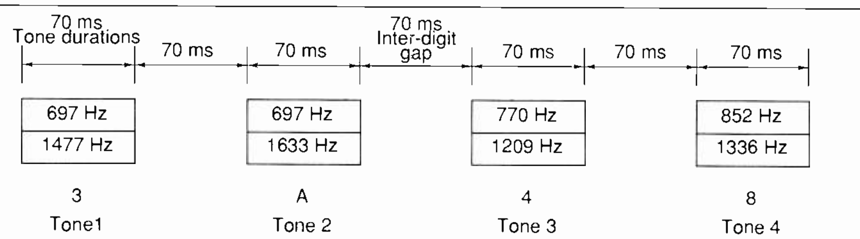


Fig. 3. A DTMF selcall sequence for address 3A48 (The sequence could be of any length dependent on selcall system.)

Alarm tone – The alarm tone may be used to trigger circuitry to indicate a predetermined alarm condition at the transmitter.

Some systems use an approach where the first tone in the selcall sequence is lengthened. This is so that a receiver may be woken from a standby or sleep mode. The lengthened first tone allows time for the receiver and selcall decoder to settle and decode the selcall sequence. The same approach is also implemented to prevent time delays caused by the operation of repeaters within the radio network from possibly corrupting the selcall sequence.

Frequently a selcall system provides a transponder function where, after a valid selcall has been received and decoded, the called station will transmit an acknowledgement signal back to the calling station. This is known as revertive signalling and provides confirmation that the selcall has been received. Selcall decoders are configured not to transpond to group calls as this could cause an overload of the channel in use and the selcall system itself.

Continuous tone coded sub-audible squelch

CTCSS is frequently used in combination with other selective calling techniques to provide an additional safeguard to prevent the reception of unwanted transmissions. It was originally developed as a means to minimise the effects of co-channel interference in systems utilising repeaters.

By requiring a radio to continuously transmit a sub-audible tone to gain access to the repeater, the accidental re-transmission of signals caused by weak mixing products of other transmissions (which will not normally have the correct CTCSS access tone present) is prevented.

The standard Electronic Industries Association (EIA) frequency codes[2], in hertz, with their Motorola alphanumeric designators, are shown in **Table 2**.

DTMF touch-tone

The DTMF *touch-tone* technique was developed by Bell Laboratories in the mid1960s to replace the conventional pulse (rotary) dialling used on the public switched telephone network. The advantages of DTMF have lead to worldwide adoption. DTMF decoders in single chip format decode the 16 possible DTMF tone pairs shown in **Table 3**.

Often the 2 of 8 tone system reduces to a 2 of 7, as signals in the 1633Hz column are not used for conventional signalling on the public telephone network. However they can, and are, often used in radio selcall systems where DTMF tones are used for signalling. The DTMF tones were selected to avoid problems that can occur on telephony channels, for example frequencies used for each tone pair are chosen to avoid false triggering by voice signals (known as falsing).

The DTMF signalling format on the public telephone system is a 40ms tone duration and a 40ms inter-digit time. It has been found that radio systems function better with at least a 50ms tone duration and a corresponding 50ms inter-digit time (tone duration and inter-digit times of 70ms are common). DTMF has been successfully used on FM radio systems[3,4,5,6] but, like many other tone based selcall systems it is unusable on single sideband radio. This is because the necessary carrier regeneration at the receiver can cause unwanted phase and

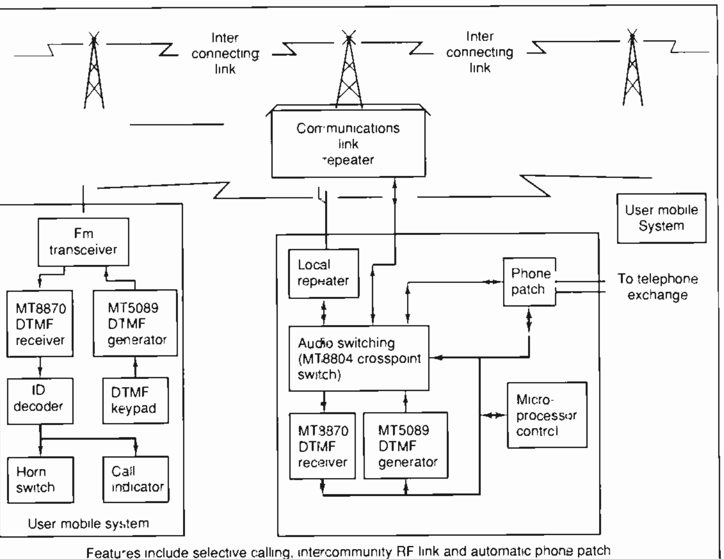
Silicon Systems DTMF

Silicon Systems range of low-cost DTMF receivers (decoders) detect and decode DTMF tones. The introduction of these single chips has considerably simplified the design of DTMF selcall and signalling system which use this standard.

Until the introduction of these devices it was common to see complete 12 tone decoders constructed from seven 567 PLL ICs plus decoding logic. Each 567 required precise frequency adjustment and frequently the system's performance would degrade over time due to drift in each tone decoding section.

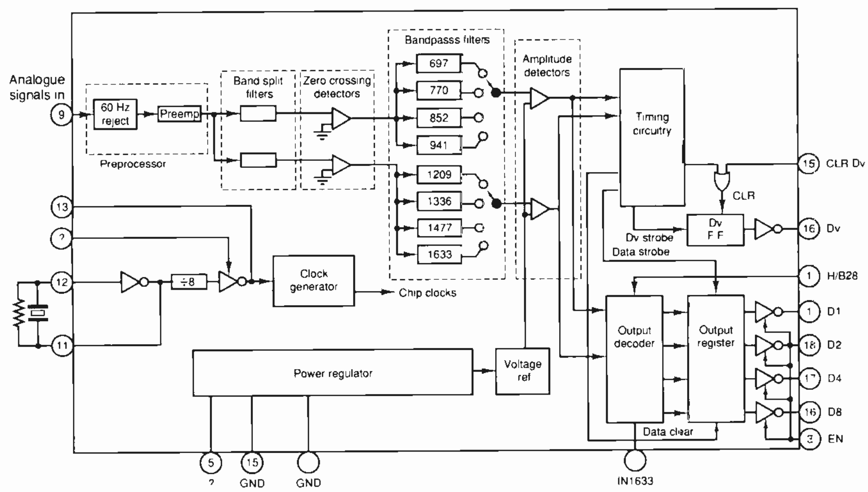
A simple DTMF selcall can be built around such a device as the *SSI204*, the circuit diagram above (from the SSI databook[14]) shows the operational simplicity. With the addition of a sequence detector (either constructed from standard logic or for example a *LS7225* custom sequence detector/keyless lock IC) after the *4514* 4-to-16 decoder IC, a selcall system can be easily realised.

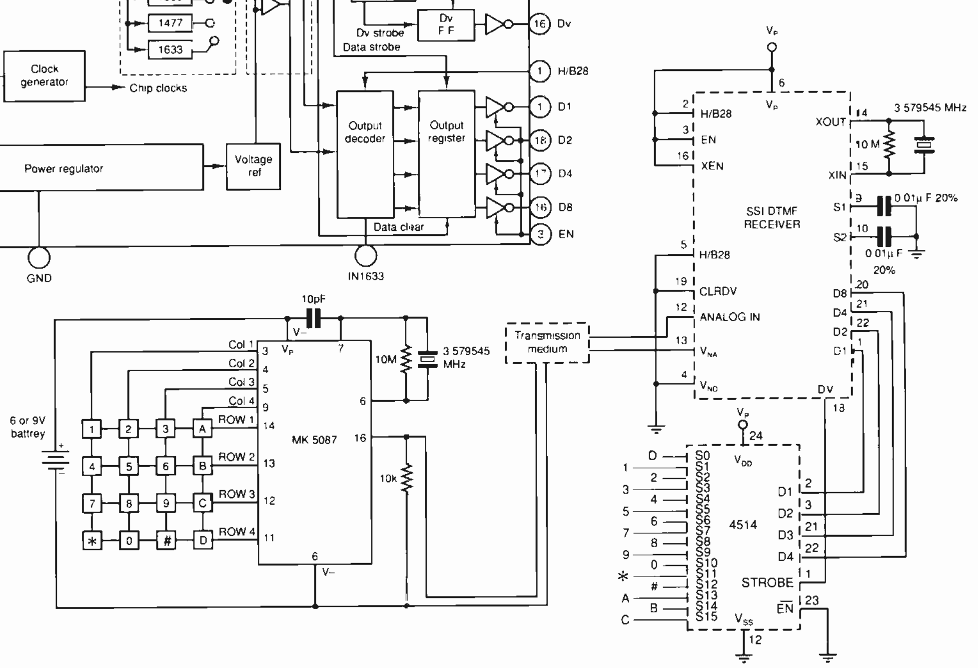
Mitel Semiconductors produce a similar range of DTMF encoders and decoders and their applications note *Applications of the MT8870 Integrated DTMF Receiver*[3] gives examples of typical radio communication systems using DTMF signalling for Selcall and control.



Features include selective calling, intercommunity RF link and automatic phone patch

Mitel Semiconductors suggested approach to a DTMF controlled radio system (MT devices are Mitel ICs).





Silicon Systems (SSI) based DTMF Tone Decoder. The detected digit from the 4514 could drive a keyless lock and sequence detector IC such as the LS7225 to produce a low cost selcall system.

The NE566 VCO and NE567 PLL tone decoder

The 566 is a general purpose voltage controlled oscillator which produces high stability square and triangular waves. The output frequency is a function of an external capacitor and resistor and, if desired, it may be linearly modulated by a control voltage. This device can used as the basis of a tone generator in a simple tone calling system.

The output frequency of a *566* (f_0) is found from the equation:

$$f_o = \frac{2.4(V^+ - V^5)}{R_1 C_1 V^+}\ \text{Hz}$$

where R_1 is between 2kΩ and 20kΩ, C_1 is in farads and V^5 is the voltage between pins 5 and 1 of the IC. The 1nF capacitor between pins 5 and 6 is to prevent parasitic oscillations that may occur during VCO switching. One should note that the output waveform does not fall to zero volts and a suitable coupling capacitor will be necessary.

The *NE567* is a phase-locked loop IC designed for general purpose tone and frequency decoding. The IC operates over a wide frequency range 0.01Hz to 500kHz and contains a logic compatible output which can sink up to 100mA. The bandwidth, centre frequency and output delay are determined using four external passive components.

The detection frequency (f_0) is determined from the following:

$$f_o = \frac{1}{R_1 C_1}\ \text{Hz}$$

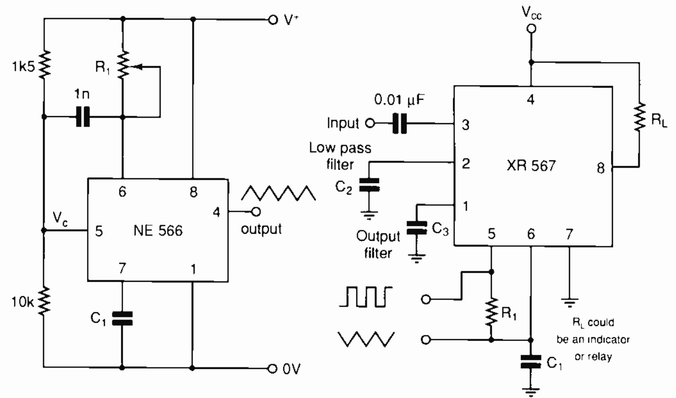
where R_1 is between 2kΩ and 20kΩ and C_1 is in farads.

C_3 forms a simple lowpass post detection filter to eliminate spurious outputs due to out-of-band signals. The value of $C_3$is not critical for most applications and, to eliminate the possibility of false triggering by spurious signals, the application sheet recommends that $C_3 >= 2C_2$ where C_2 is the loop filter capacitance.

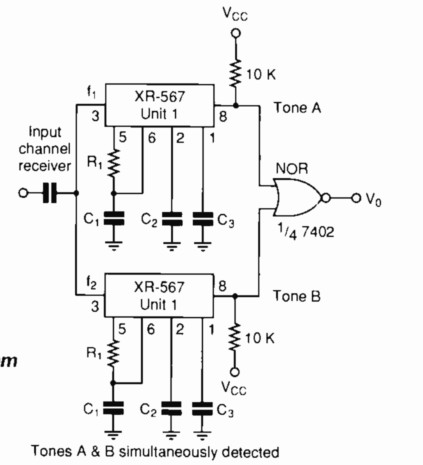
The detection bandwidth, *BW*, is the frequency range centred about f_0 for which a input signal (greater than the 20mV detection threshold voltage) will be detected by the IC. The detection bandwidth represented as a percentage of the centre frequency can be found from the approximation:

$$BW = 1070\sqrt{\frac{V_i}{f_o C_2}}\ \%$$

where V_i is the input voltage (volts rms) and C_2 is the loop filter capacitance in µF. The *567* can be used to detect single tones and with additional *567* devices and components dual tones (as in DTMF) and sequential tones. Details can be found in Exar Semiconductors' applications sheets.



Typical circuits for NE566 function (tone) generator and NE567 PLL tone decoder (for further design details see data sheets from National Semiconductor or Exar.)



A dual tone (DTMF) decoder constructed from 567 tone decoder ICs.

frequency shifts in the tones preventing reliable decoding. For SSB systems, techniques such as DSC and the approach used in aeronautical selcall, are employed.

Aeronautical selcall systems

The aeronautical radio service frequently uses selective calling techniques on HF radio to contact aircraft on transoceanic flights. An interesting aspect of the aeronautical selcall system is that only ground to air selective calling is available: the aircraft cannot selcall any other aircraft or ground station[7,8]. The purpose of the selcall system is solely to minimise the aircrew workload and remove the stress of continuously monitoring a noisy communication channel for calls.

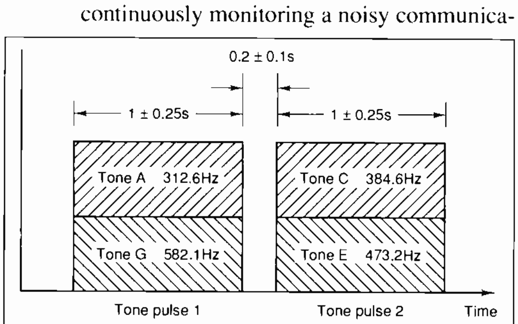


Fig. 4. Aeronautical Mark 3 Selcall structure for selcall address AG-CE.

The selcall signal sent by the ground station consists of two consecutive tone pulses of 1.0±0.25s duration, separated by an interval of 0.2±0.1s, each tone pulse consists of two simultaneously transmitted tones (similar to DTMF). The call should consist of one selcall tone sequence without repetition. A typical selcall address would be AG-CE which referring to **Table 4**, would consist of an initial burst of 312.6 and 582.1Hz tones followed by a second burst of 384.6 and 473.2Hz tones.

It is important to note that even though voice communications are predominantly in USB in the northern hemisphere, the selcall is always sent in a full carrier transmission, either full DSB or amplitude modulation equivalent (USB plus inserted carrier – H3E). This is because the selcall system cannot operate in the suppressed carrier single sideband mode. The necessary carrier insertion at the receiver can cause frequency translation errors, which would prevent satisfactory and reliable decoding of the selcall tones. As a result a dedicated AM detector is provided in the receiver solely for the demodulation and subsequent decoding of the transmitted selcall tones. This is unnecessary for aeronautical VHF radio which operates on AM.

With the tones above the number of available codes is 10,920. On detection of a valid selcall an audible warning (often a chime) and an indicator is triggered to warn the aircrew that they are being called by a ground station.

Digital selective calling

The maritime radio digital selective calling (DSC) system format is the same for both HF and VHF, only the modulation formats and signalling rates differ[9,10]. A 7-bit binary code is used with three error correcting check bits, providing a total of 128 characters. The characters 00 to 99 are decimal numbers whereas 100 to 127 are reserved for specific service commands. These 10-bit character streams are formed into data fields where the selcall message is formatted. The user is able to select various formats for his selcall message:

1. Calls to all ships in a given area.
2. Distress call.
3. Calls to certain groups of ships (such as calls to one particular fleet).
4. All-ships call.
5. Calls to an individual ship or coast station.
6. Calls to a coast station providing automatic connection to the public telephone network.

The DSC system is very powerful and in the case of a distress call allows the inclusion of

Table 4
ARINC Mark 3 airborne selcall tones

Designation	Frequency (Hz)
A	312.6
B	346.7
C	384.6
D	426.6
E	473.2
F	524.8
G	582.1
H	645.7
J	716.1
K	794.3
L	881.0
M	977.2
P	1083.9
Q	1202.3
R	1333.5
S	1479.1

data fields describing the nature of the distress, the ship's position (which can be provided automatically by the sat nav system) in addition to the time and ship identification. It is even possible to specify the preferred communication mode (radiotelephone or radio teleprinter) for subsequent radio traffic.

The digital data stream is modulated using frequency shift keying. A frequency shift of ±85Hz is used on HF, +85Hz for logic 0 and –85Hz for logic 1. For a SSB signal this equates to modulating tones of 1785Hz (logic 0) and 1615Hz (logic 1). The frequency shift keying is modified for the FM VHF service. The system modulates the carrier with tones of 2100Hz (logic 0) and 1300Hz (logic 1). HF signalling rate is 100 baud while on VHF 1200 baud is used (due to the wider channel bandwidths on VHF).

Maritime Mobile channel 70 (156.525MHz FM simplex) has been allocated exclusively for GMDSS. It is hoped that the introduction of DSC will increase the safety of all vessels and reduce the congestion of the International Calling and Distress channel 16 (156.800MHz FM simplex).

Selcall circuits

Single tone selcall decoders often consist of a PLL tone decoder such as the *LM567* and driver circuitry to mute the audio and activate call indicators. (It should be noted that this device is particularly sensitive to input drive levels for reliable tone detection).

Multiple sequential tone selcall encoders and decoders tend to use custom asics which combine all the necessary functions onto a single IC – building a sequential tone coder out of separate building blocks is definitely doing it the hard way. The British company Consumer Microcircuits produce a wide range of audio processing and signalling devices for this sort of thing. Their databook is essential reading for anyone designing a selcall system[11]. ■

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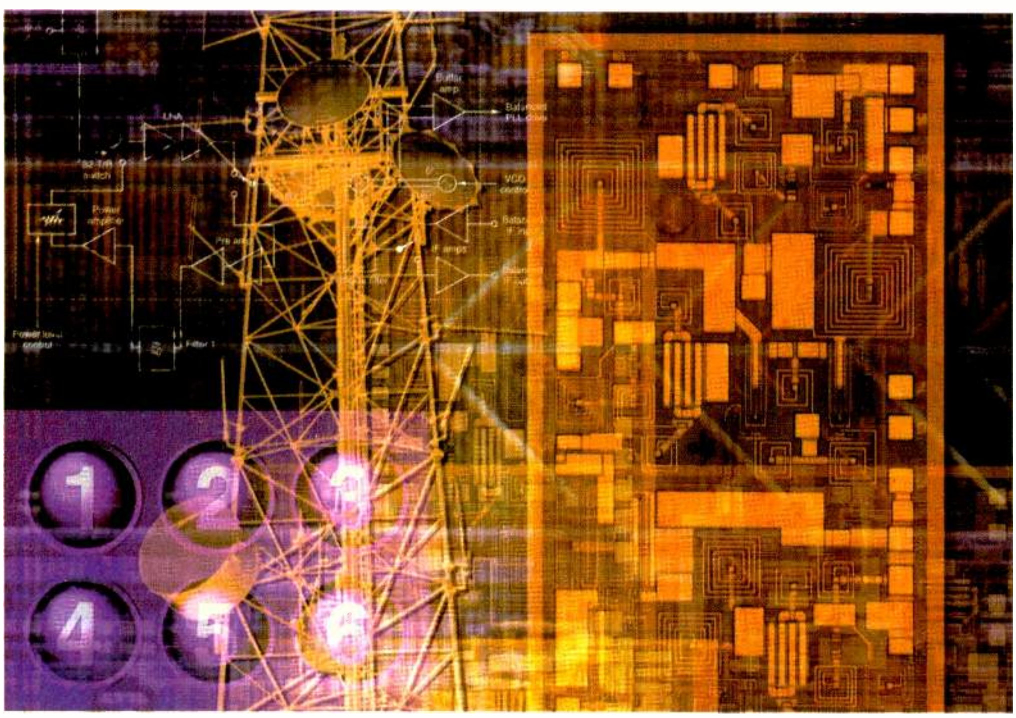
New wave MICROWAVES

3: lumped components and waveguides



Microwave hybrids and integrated circuits make use of printed features to implement reactive components. Mike Hosking* explains the design rules behind the patterns in the foil which replace components.

**Mike Hosking is a lecturer in telecommunications and microwaves at the University of Portsmouth.*

Inductive and capacitive effects can be produced from a length of transmission line as opposed to the discrete (or lumped) elements used at lower frequencies.

Take, for example, a length of 50Ω microstrip transmission line on alumina operating at a frequency of 3000MHz. This requires a width to thickness ratio of unity, so that the effective dielectric constant is about 6.7 and the transmission line wavelength is about 4cm. A distance of 1cm, therefore represents a phase shift of 90°: sufficient to change an inductive reactance into a capacitive reactance.

In a conventional electronic circuit though, operating at, say, 300kHz, the same distance of 1cm might be typical of the size of a resistor or capacitor. But the phase change across this component would be just 0.0036° and would not affect its particular value of R, C or L. A lumped element, therefore, is one whose value does not change significantly with frequency.

Generally, lumped elements measure less than about $^{1}/_{20}$th of the transmission line wavelength. Conventional components start to show distributed effects above a few hundred MHz and they also become very lossy, so that different techniques must be used to produce the circuits. Fortunately, the higher the frequency, the smaller the value of inductance and capacitance necessary to produce a given effect and, at microwave frequencies, one is usually concerned with quantities of less than a few nH and less than a few pF.

Fabrication of the lumped elements is carried out using a photo-etching process to define the circuits on low-loss substrates. The technology used is most important as, in order to be competitive with distributed circuits, high Q factors must be achieved. Inductors are formed from a length of narrow conductor track which, for inductances greater than about 1nH, can be extended into a planar spiral as in **Fig. 1a**.

Single-layer capacitances can be formed from the etched conductor ribbon by an interdigital arrangement shown in **Fig. 1b**, where use is made of the microwave fringing fields. This makes circuit processing simpler but, if a higher capacitance per unit area is required, then a three-layer metal-dielectric/metal sandwich can be used as in **Fig. 1c** for monolithic circuitry, or else an actual chip capacitor may be bonded into the circuit.

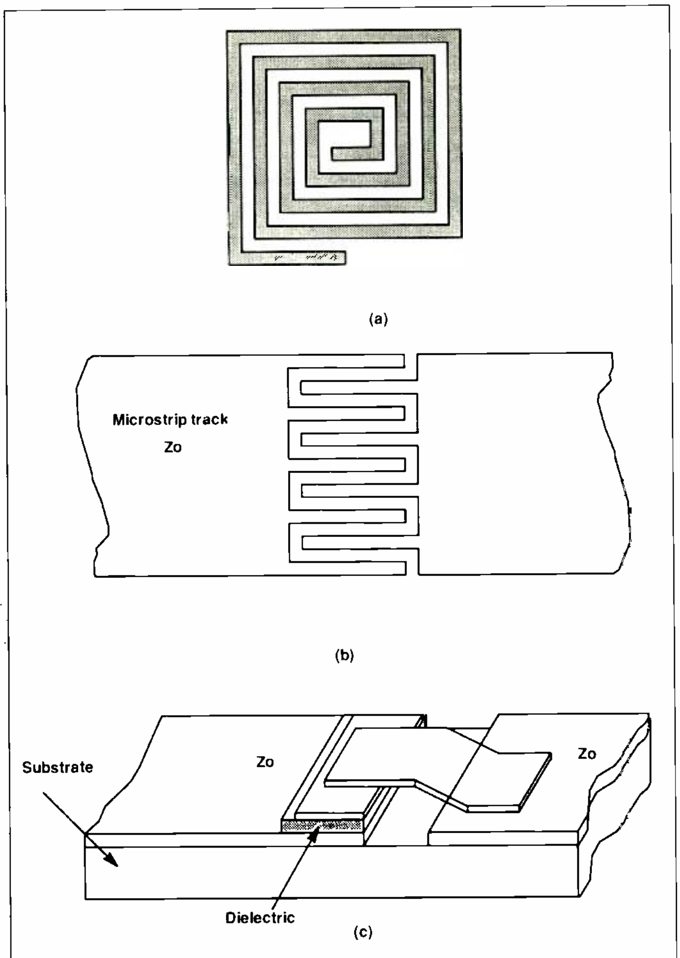


Fig. 1. Various forms of lumped-element components:
a. Square spiral inductor;
b. Interdigital series capacitor;
c. Integrated single layer capacitor. These lumped elements are usually less than about $^1/_{20}$th of a guide wavelength in size to avoid showing distributed effects.

Lumped inductors

Any conductor possesses inductance, the value depending upon its geometry. For a strip whose thickness is very much smaller than its width, the inductance per unit length is a function of the length to width ratio, l/w. If the operating wavelength were, say 10cm, and the restriction that a component must be smaller than $^1/_{20}$ of a wavelength were applied, then the permissible inductor length would be 5mm. For a width of $^1/_4$mm, the l/w ratio is 20 and the inductance would be about 4nH, neglecting the conductor thickness. By decreasing the conductor width, the inductance may be increased or, alternatively, a shorter strip could be used. However, another factor must be taken into account before reducing the size too much and that is the inductor Q. This can be represented as the ratio of the reactance to the resistance and this resistance is the high-frequency value, which takes into account the skin effect.

The skin effect increases the dc resistance and is also a function of conductor geometry, the lowest resistance being obtained with a wide strip. In addition, the conductor thickness must be several times the skin depth. Theoretically, the Q factor of a lumped element inductor increases as the square root of frequency and this relationship is achieved in practice for a strip inductor at frequencies well into S-band (2000-4000MHz). Thereafter, the rate of increase of Q slows down as, for the same l/w ratio, the inductor becomes less 'lumped'.

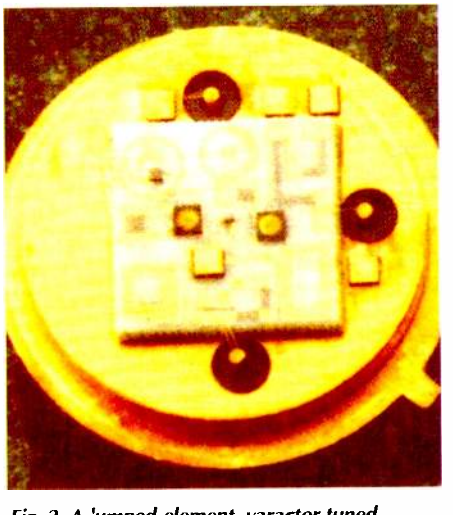


Fig. 2. A lumped-element, varactor tuned bipolar transistor oscillator on an alumina substrate and packaged within a TO-8 can. Frequency of operation is between 1000 to 2000MHz with about 20mW of output power. (ADE France)

Q factors greater than about 50 are required in order to achieve element losses of several tenths of a dB and, by taking care with substrate and conductor, such values can be obtained at 1000MHz, rising to a maximum of about 100 as the frequency increases.

So far, attention has been confined to the strip inductor but, for higher inductance values, this can be formed as a circular or square spiral such as that shown in Fig 1a. Provided that the outside dimensions meet the size requirements, such an element will still appear as lumped. Its inductance, however, will be less than that of a strip of the same unwound length, due to the mutual proximity of the turns. Once again, the highest Q factor results from wide strips which, for a given inductor size, results in small, inter-turn spacings of the order of $^1/_{20}$ to $^1/_{40}$ of a millimetre. Typically, circuits at L band (1000-2000MHz) or S band (2000-4000MHz) frequencies require inductances of less than 5nH, which can be produced by less than two turns.

However, when printed dc bias lines are also included on the substrate, it is necessary to provide an rf choke (low-pass filter) to keep the rf out of the bias circuit, and inductances of 20-40nH are commonly used. A good example of spiral inductors on a GaAs monolithic circuit can be seen in the wireless LAN circuit in Fig. 5a of Part 1 (*EW+WW* April 1994). The closely-wound, larger spirals are about 0.4mm square comprising nine turns of high impedance line and would given an inductance of about 0.4nH.

Connection to the centre of multi-turn spirals can be made either by a conductor track deposited beneath the coil and insulated from it or by a bonded wire looped directly over the turns. In this way, some circuit tuning is possible either by shorting out part of the turns or by bonding to different parts of the spiral.

Capacitors

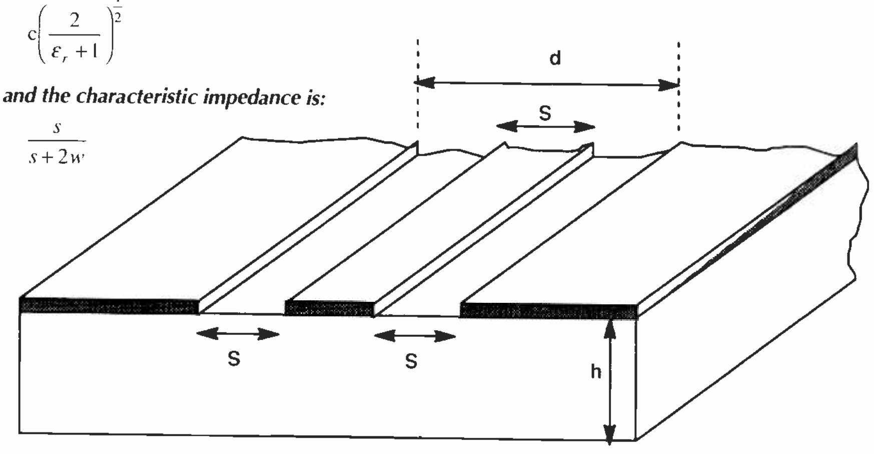
Capacitors can be produced by the thin film, vacuum deposition technique in the two forms

Fig. 3. The important dimensions of coplanar waveguide in which the approximate phase velocity is given by:

$$c\left(\frac{2}{\varepsilon_r+1}\right)^{\frac{1}{2}}$$

and the characteristic impedance is:

$$\frac{s}{s+2w}$$



shown in Fig. 1b and 1c. The interdigital capacitor is economical to construct as it involves no more processing steps than are normally used with microwave ICs, that is, one plating and one etching operation. On the other hand, the parallel plate capacitor requires the deposition and etching of both a dielectric layer and another conductor.

The interdigital capacitance is produced by the rf fringing fields across the gap between fingers: to achieve usable values, the spacing must be small. Typically the finger gaps range from 0.05 to 0.02mm and, by varying the number of fingers, produce capacitance values up to 1pF.

Using such small gaps demands attention to the substrate surface finish, conductor thickness and the photo-etching process for consistent results. Values of capacitance less than 1pF generally find circuit application at frequencies into X band (8200-12,400MHz) and this type of structure can still remain a lumped element at these frequencies.

Considerably higher values of capacitance can be produced by using the parallel plate type of Fig. 1c; it does; however, involve additional deposition and etching. For compatibility with the vacuum deposition process, a commonly used dielectric material is silicon dioxide (SiO_2), although air-bridge designs can be used where the dielectric is air and the top conductor a self-supporting bridge. The capacitance of this type of structure is directly proportional to the area and, with a 1μm SiO_2 film, 0.022μF/in^2 can be obtained. Thus, an rf bypass capacitor of, say, 40pF, is only about 1mm square.

Much effort has been spent developing chip capacitors for microwave frequencies: single-layer ceramic capacitors can operate up to 10GHz, with values of capacitance between 1pF and 1000pF depending on the frequency. These components must achieve high Q-factors through low dielectric loss and low resistance contacts. However, the contacts themselves introduce self-inductance and, hence, a potential problem of self resonance. WHen selecting chip capacitors, such self resonance must be placed outside the operating frequency range and, indeed, manufacturers now characterise microwave capacitors with a complete set of s-parameter data.

Within the range of frequencies so far mentioned, the Q factor of capacitors tend to be higher than those for inductors. In the case of the parallel plate and ground plane aided types, dielectric losses must be taken into account as well as those of the conductor. However, with good quality dielectrics, most loss can be attributed to the conductor material and the Q factor decreases with increasing frequency. At 1000MHz, Qs of several 100 or more may be achieved depending on technique, falling to 100 or less at 10,000MHz.

Lumped assemblies

The loss of lumped elements compares well with the measured loss in distributed microstrip circuits and this is important when designing such circuits as amplifiers. Higher power transistors, for example, have port impedances of a few ohms and, besides the inherent waste of power in the external circuit, the loss limits the degree of impedance matching, and hence power output, that can be achieved.

Figure 2 shows a lumped element voltage controlled oscillator covering the 1-2GHz frequency band. The central transistor is bipolar and is tuned by a varactor diode, emitter coupled LC circuit. The inductance is formed by the single turn loop in the upper left hand corner and the varactor is mounted in the centre of the loop. The circuit includes two mos capacitors (black squares) and a larger ceramic chip capacitor. Various values of circular and rectangular spiral inductors may be seen, all printed on a ceramic substrate within a TO-8 encapsulation.

Coplanar waveguide

Although microstrip circuitry offers a high degree of component integration and is a relatively mature technology in terms of the availability and accuracy of design tools, it does have some restrictions. It is not truly uni-

Faster than light

Due to the boundary constraints imposed by the enclosing walls, the wavelength of the microwave signal within the waveguide λ_g is no longer the same as its value in free space. Instead, it is given by:

$$\frac{1}{\lambda_g^2}=\frac{1}{\lambda_o^2}-\frac{1}{\lambda_c^2}$$

where λ_o is the free space wavelength and λ_c is the cut-off wavelength.

However, assuming that the denominator cannot be negative, then this relationship will make $\lambda_g>\lambda_o$, i.e. if the frequency f remains the same, then the velocity of propagation in the guide $f\lambda_g$ is apparently greater than $f\lambda_o$=c, the velocity of light.

Let us evaluate this numerically. Taking, say, a frequency of 10GHz in X-band waveguide of cut-off frequency 6.557GHz, then λ_o= 30.0mm, λ_c= 45.75mm and so λ_g=39.74mm. This would give a velocity of 3.97 x 10^8 ms^{-1} in the guide, i.e. some 32% faster than the velocity of light, of 3 x 10^8 ms^{-1}, an apparent anomaly.

However, the velocity which we have just calculated is not the velocity at which information, or energy, travels along the guide but is known as a *phase velocity* and is the velocity of a point on the sinusoidal waveform of the microwave signal. Laws of physics are not contravened because it is information which cannot travel faster than light.

One could create another apparent anomaly of this nature by considering a rotating scanning laser beam on Earth traversing the surface of the Moon. Knowing the Moon's diameter, its distance from Earth and the scan rate, we could compute the speed at which the laser beam travelled across the Moon's surface. For a scan rate exceeding about seven revolutions per minute, the rate of traverse exceeded the velocity of light. This rate could be made arbitrarily large. However, an observer on the Moon would only see the information (i.e. the light intensity) changing at a maximum rate equal to the speed of light.

Thus, inside the waveguide there is another velocity involved, which is that of information transfer by the signal and this is termed the *group velocity*. The two are related by the expression $V_{phase} \times V_{group} = c^2$, so one of the basic physical laws is not violated.

planar, due to the ground plane being on the opposite side of the substrate to the circuitry and it is thus inconvenient for designing and fabricating shunt components. In addition, lumped elements such as the spiral inductors and interdigital capacitors described previously have parasitic capacitance due to the ground plane. This reduces the self-resonant frequency of these components and, hence, their useful operating range.

CPW, **Fig. 3**, overcomes these problems and also offers a higher level of integration (about 30%). Consequently, much microwave circuit design effort, particularly for the communications industry, is being concentrated on CPW. Circuits will operate well into the millimetre wave region above 30GHz and monolithic circuits on GaAs have been produced at 60GHz.

As Fig. 3 shows, both the circuit itself and the ground plane are on the same side of the substrate, thus no holes need be made for shunt components and parasitic affects are reduced. In fact, lumped elements are used up to 30GHz. The fields propagate in a mode pattern confined to the main conductor strip of width w and the two slots of width s. The characteristic impedance is a function of the ratio:

$$\frac{s}{s+2w}$$

A typical range of values is shown in **Fig. 4** for the cases of a fused quartz substrate (ε_r=4) and a GaAs substrate (ε_r=13). The actual thickness, h, of the substrate is no longer a major parameter as was the case for microstrip, provided that it is several times the ground-ground spacing d.

A CPW circuit on quartz, for example might have h = 1mm and d = 0.2mm. To a good approximation, the mode of propagation can be described as transverse electromagnetic (TEM), ie, no field components in the direction of propagation, but there is a trade-off with spacing d before this mode becomes non-TEM. Because of this close approximation, CPW shows very little dispersion (about 2%) at frequencies up to about 60GHz and thus complex circuit shapes and discontinuities can be analysed using a more straightforward quasi-static technique.

Probably one of the main design differences in using CPW compared with microstrip is that there are many instances where lines must cross each other. This may be seen in the case of the simple cross-junction of **Fig. 5**, where the integrity of the ground planes must be maintained. For this and many other microwave circuit elements, the solution is to form an air-bridge by raising one track over the other. This, of course, has a significant effect on how the fields 'see' such a discontinuity and it must be analysed. Most cad tools for microstrip are essentially two-dimensional; software development effort is going into three-dimensional synthesis methods for CPW.

We can expect to see further development and application of the coplanar technology, including attention to two disadvantages: slightly higher attenuation than microstrip and more difficulty with heat dissipation.

Waveguide

Although all transmission lines are waveguides, the term has come to apply to the dielectric-filled conducting tube; the dielectric usually being air. Nowadays, waveguide is mainly used when high powers are involved – radar for instance – to guide the signal from transmitter to antenna and in certain complex antenna feeds where minimum loss is paramount. However, the size and weight of waveguide, together with a relatively narrow frequency range for a particular size, prohibits its use in integrated circuits.

For a given size of waveguide (a x b) there is a lower frequency limit below which no microwave field can propagate; this is termed the cut-off frequency. As the frequency is raised, the first sustainable mode will form, termed the dominant mode and, with further increases in frequency, then other, higher-order modes will appear. Thus, waveguide (and all transmission lines) behave as high-pass filters but are almost invariably restricted to operation in the dominant mode.

To operate at a higher frequency, the wave-

Coaxial impedance and loss

For any low-loss transmission line, the characteristic impedance Z_0 can be expressed as:

$$Z_o = \sqrt{\frac{L}{C}}$$

where L and C are the inductance and capacitance per unit length, respectively. It is relatively simple to evaluate L and C for coaxial line to yield:

$$Z_o = \sqrt{\frac{60}{\varepsilon_r}}\ln\left(\frac{b}{a}\right)$$

This is the equation plotted in Fig. 4.

Attenuation is caused by loss in the conductors and loss in the dielectric filling the transmission line. Conductor loss α_c is given by:

$$\alpha_c = \frac{1}{27.6\pi\sigma\delta_s}\left(\frac{1}{a}+\frac{1}{b}\right)\frac{\sqrt{\varepsilon_r}}{\ln\left(\frac{b}{a}\right)}$$

in dB per unit length.

Dielectric loss is given by:

$$\alpha_D = 27.3\frac{\sqrt{\varepsilon_r}}{\lambda_o}\tan\delta$$

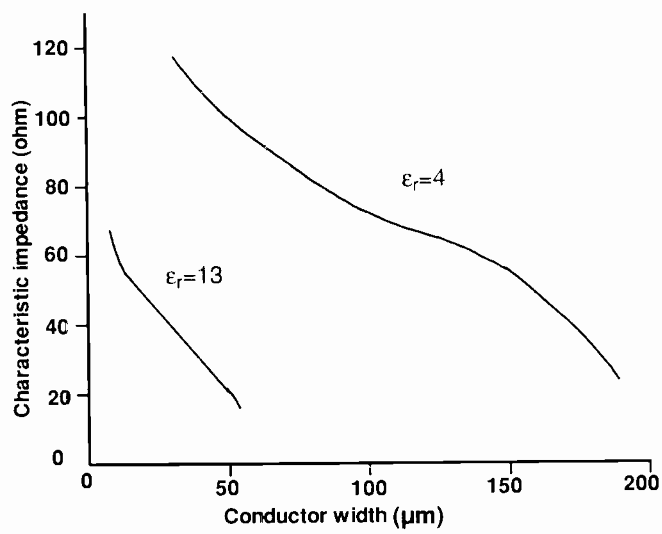
also in dB per unit length, where tanδ is the loss tangent of the dielectric (4×10^{-4} for polyethylene and PTFE).

Loss in the conductors is proportional to the square root of frequency while the dielectric loss is linearly proportional to frequency. So, at higher frequencies, the dielectric loss becomes more dominant than conductor loss.

Fig. 4. The characteristic impedance of coplanar waveguide as a function of conductor width for two substrate dielectric constants. The substrate thickness in each case was assumed to be greater than 500μm.

Table 1. Rectangular Waveguide Data

Recommended Frequency Range (GHz)	TE_{10} cutoff Frequency (GHz)	Inside dimensions (inches)
1.12 - 1.7	0.908	6.500 x 3.250
2.60 - 3.95	2.078	2.840 x 1.340
5.85 - 8.20	4.301	1.372 x 0.622
8.20 - 12.4	6.557	0.900 x 0.400
12.4 - 18.0	9.486	0.622 x 0.311
18.0 - 26.5	14.047	0.420 x 0.170
26.5 - 40.0	21.081	0.280 x 1 40
40 - 60	31.357	0.188 x 0.094
50 - 75	39.863	0.148 x 0.074
75 - 110	59.010	0.100 x 0.050
110 - 170	90.854	0.065 x 0.0325
140 - 220	115.75	0.051 x 0.0255



guide cross sectional dimensions must be progressively reduced so as to prevent the formation of higher order modes. This leads to a range of standard waveguide sizes, some of which are shown in **Table 1** for rectangular guide. With few exceptions, the aspect ratio is fixed at 2:1, this being a compromise between power handling, losses and over-moding. The internal dimensions of waveguide operating above 100GHz become quite small, so that the difficulty of creating circuit components within these dimensions increases.

The longest wavelength which can propagate down rectangular waveguide (that corresponding to the dominant mode) is equal to twice the width of the guide, $2a$. In practice, the centre frequency is made equal to about 1.5 times the cut-off frequency and actual operation is restricted to within about ± 20% of this. For circular waveguide, the cut-off wavelength of the dominant mode is $3.42r$.

The reason for restricting waveguide transmission to microwave frequencies becomes apparent from Table 1: to cover the FM broadcast band from 88MHz would require a rectangular waveguide of a size which could garage two family cars side by side.

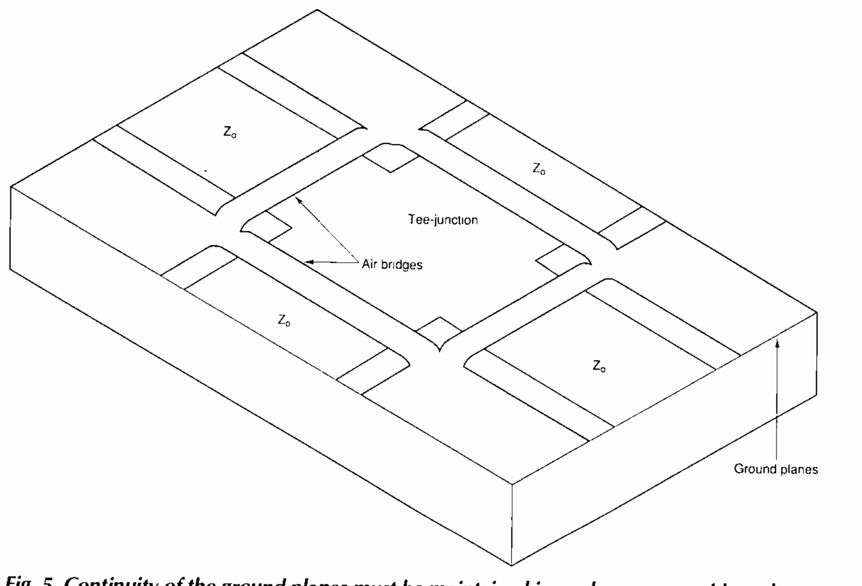


Fig. 5. Continuity of the ground planes must be maintained in coplanar waveguide and so the technique of forming air bridges over the conductor tracks is used. Consequently, a 3D CAE analysis must be developed for accurate calculation and simulation of the fields.

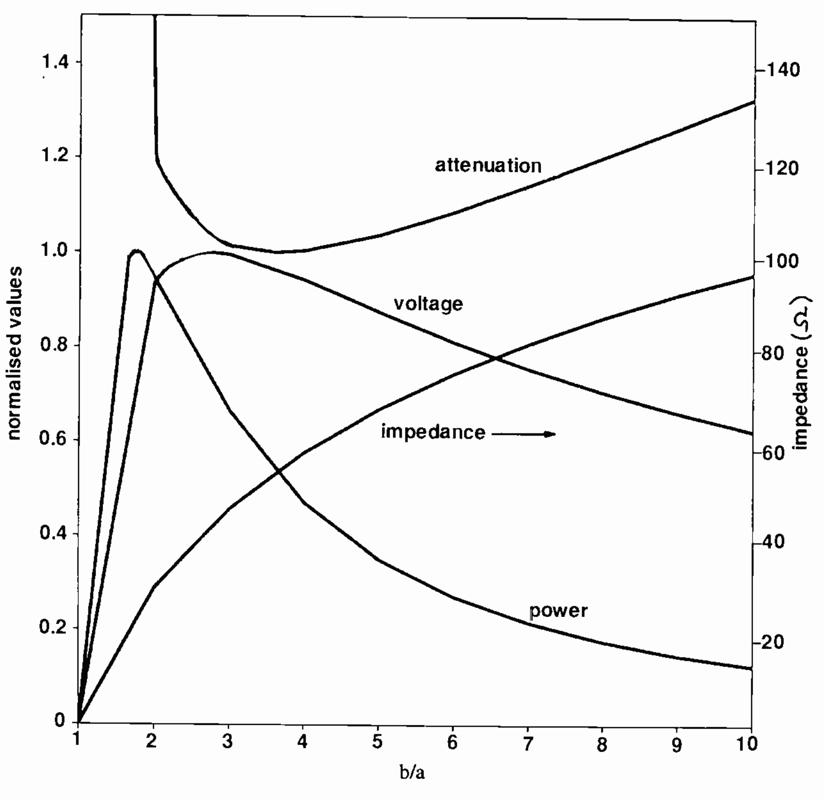


Fig. 6. There is an optimum ratio of outer to inner radii for a coaxial line depending on the transmission characteristics required. These are shown here for the case of an air-filled line where it can be seen that the usual 50Ω impedance gives a compromise between these optima.

Coaxial line

Unlike waveguide, coaxial line has no cut-off frequency: a given size will propagate signals down to zero frequency and up to infinity. The principle mode of transmission in coaxial line is termed transverse electromagnetic (TEM) meaning that all components of electric and magnetic field are at right angles to the direction of propagation. This mode requires a geometry of at least two conductors and, hence, cannot exist in waveguide. Higher order modes are possible but are almost always undesirable and start to occur when their wavelength approaches the mean circumference of the line. Thus, in practice, there is a restriction on the upper frequency of coax, usually around 40GHz.

Coaxial cable is normally filled with a dielectric material, usually polyethylene or PTFE and so the velocity of propagation is slowed from that in free space by an amount dependant upon the dielectric constant (relative permittivity) of the material, ε_r. For the TEM mode, the phase and group velocities are equal, so that the velocity in the line is simply $v=c/\sqrt{\varepsilon_r}$. Typical values for ε_r at 10GHz (there is a slight frequency dependence) are 2.25 for polyethylene and 2.08 for PTFE.

The value of 50Ω characteristic impedance has become the most widely used standard for coaxial connectors and instrumentation input/output ports. 75Ω is largely used for signal connections to dipole-type antennas since the basic λ/2 dipole has an input impedance of approximately 73Ω. Characteristic impedance and several other parameters of coaxial line are functions of the ratio of outer and inner diameters, as shown in **Fig. 6** for the case of PTFE dielectric. Coaxial impedances found in practice range typically from 20Ω to 150Ω, with standard 50Ω line having a diameter ratio b/a = 3.3 for this dielectric.

It is interesting to note that, for characteristics such as attenuation, voltage breakdown and power handling, there are optimum, but different, ratios of b/a.

Lowest conductor loss occurs for a value of 3.6, maximum voltage capability for 2.7 and maximum power handling for 1.65 corresponding to impedances of 53.3Ω, 41.3Ω and 20.8Ω respectively. ■

Next month: circuits using Impatt and Gunn devices.

Tektronix Das 9100 Logic analysers. Complete – £700.
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Marconi 6460/1 Power Meters + Heads RF various – £250 ea.
Marconi 6460 Power Meters + Heads RF various – £150 ea.
Marconi CT499 MkII RF Watt Meter Absorption 1–3–10–30–100W – 50ohm + 70ohm Convertor Adaptor – £350. Includes RF Adaptors Low Loss.
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Solatron 1170 FX Response ANZ–Led – £300.
HP59401A Bus System ANZ.
STC Optical Fibre Reflectometer OFR6 – £300.
HP3497A Data Acquisition Control Unit – £300.
Redifon Synthesized Receiver R1001–CW–AM–USB–LSB–PIC–STORE–Led Readout – 15KC/s–30MC/s – £600.
Racal LA1117 Piccolo Modem – £150.
Redifon RFS11 Pre-Selector – Post Selector – 1MC/s–32MC/s – £200.
Sayrosa 3–39MC/s Programmable Freq Syn–Type 607 + 607B – £200–£300.
Racal TA1816 1Kw Solid State Transmitter – MA1034 + 3 Pare Amps – £1,000.
H.C.D. Research Ltd Precision Oscillator 1519 5000/1000KHz – £250.
Nimbus 400Hz Convertor Mains – 240AC–50OC/s Input–Output 115V–400Hz – 500V/A Programmable 0–125V. G500 FPL – £300. G200R as above – 200V/A – £200.
Hedinair + Montford Environmental Ovens etc. Big + Small – £200–£1,000.
ICL Clean Linez Unit – £300.
HP6943A Multiprogrammer extenders – £300.
HP6525A DC P.U.O.–4000V–50M/A – £350.
Polaroid + CR-9 Cameras for Oscilloscopes – £100.
HP3710A IF–BB Transmitter, HP3702B IF/BB Receiver + 3705A Phase Detector – £250.
Moor & Reed Frequency Convertor 400C/s 3 Phase Type SFC 6K/3AXR – Solid State – 20Amps/Mains 240V AC–50C/s Input – £600.
HP7586B Plotter Large Quantity Accessories Pens etc – HP–IB–RS2320 – £1,000.
Benson 16 BZ80 Asynchrome Interface + Accessories & Pens etc – £600.
Imtec 6000 Microfilm Reader & Printer, A4–A2 sizes – Plain Paper – Various Lenses – £750.
Fluke Y5020 Current Shunt – £150.
B&K 2107 FX Analyser – £250.
B&K BFO 1022 – £200.
Tektronix Spectrum ANZ–1L5–50HZ – 1MC/s – £150 – 1L20–10MC/s–4200MC/s – £250.
FARNELL P.U. AP60–50–60V–50Amps – £1,000.
Tracor 527E Frequency Difference Meter – £350.
HP8900B Peak Power Calibrator – £250.
B&K 2425 Electronic Voltmeter – £200.
HP4437A 600 ohms Attenuator – £150.
HP6177C DC Current Source – £250.
B&K Two Channel Level Recorder – £400.
Tektronix 2213 60MC/s Oscilloscope – £300 – 2213A – £350.
Tektronix 2215 60MC/s Oscilloscope – £350.
Tektronix 2445 150MC/s Oscilloscope – £1,200.
Tektronix 2246 100MC/s Oscilloscope – £1,000.
Tektronix 2225 150MC/s Oscilloscope – £800.
Tektronix 2245 100MC/s Oscilloscope – £700.
Tektronix 491 Spectrum ANZ 10ML/s–40GHC – £1,000.
Farnell P.U. H30/100 – £500.
Schlumberger S1 4922 Radio Code ANZ – £400.
Aerial Array on metal plate 9"x9" containing 4 aerials plus Narda detector. 100–11GHZ using N type and SMA Plugs & Sockets – ex equip – £100.
Marconi TF2175 Power Amplifier – 1.5MC/s–520MC/s + Book – £100.
Schlumberger 2741 Programmable Microwave Counter – 10HZ–7.1GHZ – £750.
Schlumberger 2720 Programmable Universal Counter – 0–1250MC/s – £600.
Tektronix 576 Calibration Fixture – 067–0597–99 – £250.
Texscan Rotary Attenuators BNC/sMA – 0–10–60–100DBS – £50–£150.
HP809C Slotted Line Carriage – Various frequencies to 18GHZ – £100–£300.
HP532–536–537 Frequency Meters Various Frequencies – £150–£250.
S.E. Lab SM215 MkII Transfer Standard Voltmeter – 1000 Volts.
Ailtech Stoddart P7 Programmer – £200.
HP6181 DC current source – £150.
HP59501A HP–IB Isolated D/A Power supply programmer.
HP3438A Digital Multimeter – £150.
HP6177c DC Current Source – £150.
HP6207B DC Power Supply – £100.
HP741B AC/DC differential voltmeter standard (old colour) – £100.
HP6209B DC Power Unit – £100.
Fluke 431C High voltage DC supply.
Tektronix M2 Gated Delay Calibration fixture – 067–0712–00.
Tektronix Precision DC Divider Calibration fixture – 067–0503–00.
Tektronix Overdrive Recovery Calibration fixture – 067–0608–00.
HP5011T Logoc Trouble Shooting Kit – £150.
PPM 8000 Programmable Scanner.
Fluke 730A DC Transfer Standard.
B&K 4815 Calibrator Head – £150.
B&K 4812 Calibrator Head – £150.
HP FX Doubler 938A or 940A – £300.
HP461A Amplifier 1KC–150MC/s – Old Colour – £100.
Ailtech Precision Automatic noise figure indicator type 75 – £250.
Adret FX Synthesizer 2230A – 1MC/s – £250.
Marconi TF2512 RF Power Meter – 10 or 30 Watts – 50 ohms – £80.
Marconi 2830 Multiplex Tester.
Marconi 2828A Digital Simulator.
Marconi 2831 Channel Access Switch.
Marconi TF2337A Automatic Distortion Meter – £150.
HP489A Micro-Wave Amp–1–2GHZ – £500.
Fluke 893A Differential Meters – £100 ea.
EG&G Parc Model 4001 Indicator 4203 Signal Averager PI.

Tecktronix Plug-In AM503–PG501–PG508–PS503A–PG502.
Cole Power Line Monitor T1085 – £250.
Claude Lyons LCM1P Line Condition Monitor – £250.
Bell & Howell TMA3000 Tape Motion Analyser – £250.
HP5345A Automatic Frequency Convertor – .015–4GHZ – £350.
HP3200B VHF Oscillator – 10–500MC/s – £200.
Sencore SC61 Waveform ANZ–Microprocessor 60–100MC/s – £350.
Schlumberger 3531D Date Acquisition System – £300.
Marconi 6700A Sweep Oscillator with 1–2GHZ PI 6730A – £400.
B&K 2218 Sound Level Meter – £600.
EIP 331 18GHZ Counter-Microwave – Led – £700.
EIP 351D 18GHZ Counter-Microwave – Led – £800.
EIP 451 18GHZ Counter-Microwave – Led – £900.
EIP 545 18GHZ Counter-Microwave – Led – £1,200.
Systron Donner 6054D 18GHZ Counter – Led – £800.
Systron Donner 6057 18GHZ Counter – Microwave – Nixey – £600.
HP5340A 18GHZ Counter Microwave – Led – £1,200.
HP5340A 18GHZ Counter Microwave – Nixey – £800.
Systron Donner 6061 18GHZ Counter Microwave – Nixey – £500.
Austron 6014 FX Multiplier – £250.
Austron 2004 Receiver Loran – £250.
Austron 1201A Linear Phase Recorder – £250.
Austron 2010A Disciplined FX Standard – £250.
Microtel MSR-903 Microwave Receiver – .03–18GHZ – AM–FM – £2,000.
Microtel MSR-903 Microwave Receiver – .1–18GHZ – AM–FM – £2,000.
Microtel MSR-903A 18GHZ FX Counter for Above – £1,000.
Ailtech NM17/27 EMI/Field Intensity Meter – .01–32MC/s – £1,000.
Ailtech NM37/57 EMI/Field Intensity Meter – 30–1000MC/s – £1,000.
Ailtech NM65T EMI/Field Intensity Meter – 1–10GHZ – £1,000.
Fluke 5205A Power Amp – £1,200.
B&K 1623 Tracking Filter.
B&K 2607 Measuring Amp.
B&K 2134 Sound Intensity Analyser.
B&K 280 Microphone Power Supply.
B&K 4408 Two Channel Microphone Selector.
B&K 4910 Stroboscope.
B&K 1606 Pre-Amp Vibration.
B&K 4420 Distribution Analyser.
B&K 1014 B.F.O. Oscillator.
B&K J2707 Power Amplifier.
B&K 2305 Level Recorders.
B&K 2307 Level Recorders.
B&K 7003 Tape Recorders.
B&K 2615 Charge Amplifier.
Fluke 9010A Micro-systems trouble shooter & many Pods – £350 + Pods or Probe.
Racal/Dana 5002 Wide Band Level Meter.
Racal/Dana 5006 Digital Multimeter.
Racal/Dana 5005–S–4622 Digital Multimeter.
AVO RM215 – L/2 AC/DC Breakdown Leakage & Ionisation Tester – £400–£450.
Fluke 80K – 40 High Voltage Probes – New in Case – £100.
Watkins Johnson 340A–4 RX LF–1–800KC/s AM–FM–CW – Led Readout – £750.
Watkins Johnson DMS – 105A Demodulator–AM–FM–SSB – Led Readout – £600.
Watkins Johnson RS–111–1B–40 VHF Receiver – 30MC/s–1000MC/s – AM–FM–CW – Pan Display – £700.
Watkins Johnson 373A–2 HF Receiver – 0.5–30MC/s – AM–FM–CW – £400–£500.
Watkins Johnson Receivers from 1KC/s to 10,000MC/s also Tuning Heads – Amps–Counter Readouts – Signal Displays – Distribution Amps – HF Multicouplers – IF Demodulators – Signal Monitors etc.
Racal MA1720 TX Drive Units 1–30MC/s – £500–£750.
Racal MA1723 TX Drive Units 1–30MC/s – £1,000–£1,500.
Racal MA1724 TX Drive Units 1.6–25MC/s – £500.
Racal RA1792 HF RX-100KC/s–30MC/s – £1,000 Back Lighting.
Racal RA1772 HF RX-15KC/s–30MC/s – £600.
Racal RA17L HF RX .5MC/s–30MC/s – £100–£250.
Plessy PR2250G & H-HF RX LF to 30 MC/s–Memory–Led Readout – £650–£1,000.
B&K 2609 Measuring Amp – £250.
B&K 1613 Filter – £100.
B&K 4215 Artificial Mouth – £250.
B&K 4219 Artificial Voice – £250.
B&K 4220 Piston Phone – £120.
Dynamic Sciences R-1250 Tempet Receiver – 100HZ–1000MC/s – AM–FM – £2,000
HP3406A Sampling Voltmeter (Broadband) – New Colour – £200.
HP7404A Oscillograph Recorder – 4 Track – £350.
HP9872B Plotter – 4 pen – £300.
HP11710B .01–11MC/s – Down Convector for 8640B – £350.
HP11720A Pulse Modulator – 2–18GHz – £1,000.
HP8403A Modulator – 0.4–12.4GHz (8731–8735B) Modulators – £100–£250.
HP Pin Modulators for above – Various frequencies – 0.4–12.4GHz – £150.
HP8699B Sweep Plug-in – 0.1–4GHz – Using Yigs–Solid State – £300.
HP8690B Mainframe – £250. All PI available – .1–40GHz Sweep.
Racal–SG Brown Comprehensive Headset Tester (with artificial head) Z1A200/1 – £350.
Marconi 893B AF Power Meter – £200.
Microwave Systems MOS/3600 Microwave Frequency Stabilizer – 1GHz–40GHz – £1,000.
ACL SR-209–6 Field Intensity Meter Receiver – Pl's – 5MC/s–4GHz – P.O.R.
Ailtech 136 Precision Test RX + 13505 Head – 2–4GHZ – £350.
SE Lab Eight Four FM 4 Channel Recorder – £200.
Datron 1065 Auto Cal Digital Multimeter & Instruction Manual – £400.
Datron 1061 Auto Cal Digital Multimeter & Instruction Manual – £400.
Racal MA259 FX Standard – Output 100 KC/s – 1–5MC/s Internal Nicad Battery – £150.
Edwards E2M8 Rotary Vacuum Pumps – Brand New & Boxed – £500 ea.
Fluke 9100A Troubleshooter & Pods – New Boxed – £750–£1,000.
HP1140 & 1743 Oscilloscopes 100MC/s – £300–£450.
Tektronix PI 7A19–7A29–3A–4–6–7m11–
Tektronix 7000 Series Oscilloscopes We can supply all variations of Main Frames and Plug-ins for this range from stock up to 1GHz – £300–£3,000.

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CIRCLE NO. 115 ON REPLY CARD

Mike Button looks at the more practical side of I²C design – including a worked example of an I²C frequency synthesiser – having covered I²C theory in the January issue.

Busman's guide to I²C

Designed by Philips, the inter-integrated circuit bus allows bi-directional data communication over two wires at up to 100kbaud. Abbreviated to I²C, the bus has a defined protocol allowing transfer between any number of master and slave devices – in theory. There is automatic arbitration for multi-masters and devices can be added or removed without transgressing the bus protocol.

A list of the more popular master and slave bus ICs is presented in **Table 1**.

Writing I²C drivers

In most control applications, a single microcontroller performs all functions. As a result, the I²C bus multi-master feature is not essential. Not all proprietary slave devices need clock timing and synchronisation control so these are not always needed either.

Provided that the microcontroller has suitable spare i/o pins and there is room in its rom, a simple software driver for an I²C bus can be added to enhance existing systems. Note that the I²C bus specification states that devices can be added or removed at any time and that the bus should function correctly even if the power supply to a particular device or board is removed.

In the event of a power failure, the open collector output on the bus leads is needed to prevent the wire being pulled to ground via the inherent parasitic diode at the V_{cc} connection. This requirement is automatically fulfiled with the specified open collector wired-And logic function.

If all devices connected to the bus are supplied from a single power source – as is normally the case – a further relaxation in the bus specification is allowable. Both SDA and SCL can be connected directly to the quasi bi-directional ports of an *8048* or *8051* type microcontroller.

Using microcontrollers with active pull-up ports whose direction is software defined will need two pins per I²C bus lead. One forms an output connected to the bus via an open-collector device, the other an input connected

Table 1. A multitude of I²C controllable devices is now available. In the slave address bit descriptions, U represents a user-defined bit which is normally a hardware strap. Letter A signifies a software-defined or pre-allocated bit. Symbols in the read/write column, i.e. bit 0 of the slave address, are '1' for read only, '0' for write only and 'A' for read or write. Slave devices are listed below, microcontrollers bottom right.

Device	Description	Slave Address High Low	R/W	Manufacturer
Memory				
CAT24C02	EEPROM 256 x 8	1010 UUU	A	CATALYST
CAT24C04	EEPROM 2 x 256 x 8	1010 UUU	A	CATALYST
NM24C016	EEPROM 8 x 256 x 8	1010 UUU	A	Nat Semi
NM24C02	EEPROM 256 x 8	1010 UUU	A	Nat Semi
NM24C04	EEPROM 2 x 256 x 8	1010 UUU	A	Nat Semi
NM24C08	EEPROM 4 x 256 x 8	1010 UUU	A	Nat Semi
PCF8581	EEPROM 128 x bits	1010 UUU	A	Philips
PCF8582	EEPROM 256 x 8	1010 UUU	A	Philips
PCF8594	EEPROM 2 x 256 x 8	1010 UUA	A	Philips
PCF8598	EEPROM 4 x 256 x 8	1010 UAA	A	Philips
X24C01	EEPROM 128 x 8 bits	1010 UUU	A	Xicor
X24C02	EEPROM 256 x 8 bits	1010 UUU	A	Xicor
X24C04	EEPROM 2 x 256 x 8 bits	1010 UUA	A	Xicor
X24C16	EEPROM 8 x 256 x 8 bits	1010 AAA	A	Xicor
PCF8570	RAM 256 x 8	1010 UUU	A	Philips
PCF8571	RAM 128 x 8	1010 UUU	A	Philips
PCF8583	RAM 256 x 8 with clock calendar	1010 UUU	A	Philips
I/O				
PCF8566	LCD Driver, 96 segments	0111 11U	0	Philips
PCF8568	LCD Row Driver	0111 10U	0	Philips
PCF8569	LCD Column Driver	0111 10U	0	Philips
PCF8574	8 bit I/O	0100 UUU	A	Philips
PCF8574A	8 bit I/O	0111 UUU	A	Philips
PCF8576	LCD Driver 160 segments	0111 00U	0	Philips
PCF8577	LCD Driver, 64 segments	0111 010	0	Philips
PCF8577A	LCD Driver, 64 segments	0111 011	0	Philips
PCF8578	LCD Row/Column Matrix Driver	0111 10U	0	Philips
PCF8579	LCD Row/Column Matrix Driver	0111 10U	0	Philips
SAA1064	LED driver 4 digits	0111 0UU	A	Philips
SAA1300	5 bit high current driver	0100 0UU	0	Philips
Analogue				
PCF8591	4 channel 8 bit ADC with 8 bit DAC	1001 UUU	A	Philips
TDA8444	Octal 6 bit DAC	0100 UUU	0	Philips
TDA8442	Quad 6 bit DAC	1000 100	0	Philips
Audio				
PCD3311	Tone Generator. DTMF - Modem - Music	0100 10U	0	Philips
PCD3312	Tone Generator. DTMF - Modem - Music	0100 10U	0	Philips
PCF8200	Voice Synthesizer	0010 000	A	Philips
TEA6300	Sound Fader Control	1000 000	0	Philips
TEA6310	Sound Fader Control	1000 000	0	Philips
TDA8420/1	Hi Fi Stereo Audio Processor	1000 00U	0	Philips
TDA8425	Audio Processor with LS channel	1000 001	0	Philips
Radio				
TSA5510	1.3 GHz Frequency Synthesizer	1100 0UU	A	Philips
TSA5511	PLL Frequency synthesizer for TV	1100 0UU	A	Philips
TSA5512	PLL Frequency synthesizer for TV	1100 0UU	A	Philips
TSA6057	PLL Frequency synthesizer for Radio	1100 01U	0	Philips
UMA1010	PLL Frequency synthesizer Low Power	1100 00U	A	Philips
UMF1009	PLL Frequency synthesizer	1100 0UU	?	Philips
UMA1000T	Cellular Radio Data Processor	1101 1UU	A	Philips
TEA6100	FM/IF Tuning interface	1100 001	A	Philips
Television				
SAA1136	PCM Audio indent word interface	0011 110	0	Philips
SAA5243/4	Enhanced Teletext Circuit	0010 001	A	Philips
SAA7191	S-VHS digital multistandard decoder	1000 1U1	?	Philips
SAA7192	Digital colour space converter	1110 00U	?	Philips
SAA7199	Digital Encoder	1011 000	?	Philips
SAA9020	Field Memory Controller	0010 1UU	A	Philips
SAA9050/1	Digital Multi-Standard TV decoder	1000 101	A	Philips
SAA9068	Picture in picture controller	0010 01U	0	Philips
SAB3035/6	CPU interface for tuning & control	1100 0UU	A	Philips
SAF1135	Data Line Decoder. (Teletext)	0010 0UU	1	Philips
TDA4670	Picture in Picture Improvement Cct.	1000 100	?	Philips
TDA4680	Video Processor	1000 100	?	Philips
TDA8440	CTV Receiver Switch	1001 UUU	0	Philips
TDA8442	Colour Decoder Interface	1000 100	0	Philips
TDA8443A	YUV/RGB interface	1101 UUU	0	Philips
TDA8461	PAL/NTSC colour decoder	1000 10U	0	Philips
Bus Controllers				
PCD8584	Bus Controller	N/A		Philips

directly to the bus wire.

Flow of the functions needed for a simple I²C bus software driver is shown in **Fig. 1**. Suitable for any microcontroller, this routine provides for minimum SCL high and low times. It also takes advantage of the fact that the SCL clock high and low periods can be any ratio.

Master transmitter/receiver facilities only are provided for by the driver. It assumes that no other master will attempt to access the bus and that any slave will not attempt to perform SCL clock synchronisation.

Subroutines for the start, transmit, receive and stop conditions are shown. These need a control software routine designed to access the particular slave device being used, examples of which are given later.

The start routine assumes that the bus is in an idle state as left by the stop routine, i.e. SDA and SCL high, or that SCL is low as left by the transmit or receive routines. On initialisation, the control routine should set SCL and SDA high. Both transmit and receive routines count and control SCL timing. They assume that the software instruction set can shift, i.e. rotate, the contents of an eight bit data register for parallel-to-serial conversion.

To transfer data, the required slave device must first be accessed by sending its address. This is achieved by sending a start condition followed by a transmit with the data register loaded with the required address. For a master transmit function, the data register must be loaded with the slave address bit 0 at logic zero, signalling write. The master receiver function requires that the data register must be loaded with the slave address with bit 0 at logic one, representing a read.

In the transmit routine, the data register is preloaded from the control software. All eight bits are shifted through bit 7. The SDA lead is

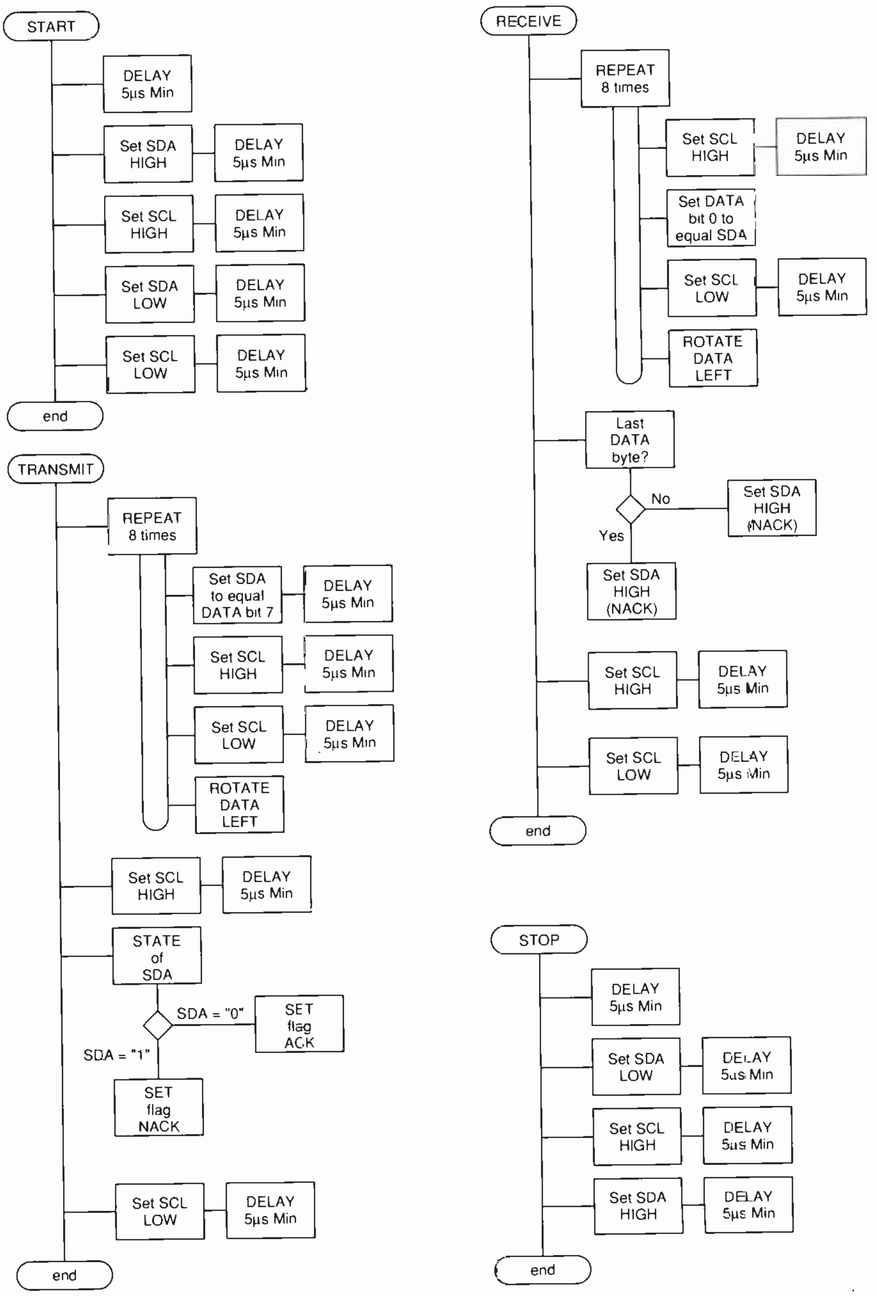
Table 1 continued. Micro-controllers with I²C hardware. The number of different microcontrollers is great. There is normally more than one version of a particular type. Each device has a base code with variant prefixes, suffixes or code change. Most of the micros are manufactured with some or all of the variants below:

- PROM type, ie mask programmable, EEPROM or external.
- PROM size.
- Working temperatures.
- CMOS, NMOS or other technologies.
- Packaging.
- Working frequency.
- Working voltage.

Base code of some of the more popular devices:

3315	8048 wide volt range for telephony functions
68000	68000 CPU with MMU/UART/DMA/timer
83C524	8051 micro-controller with 512 bytes RAM
83C552	8051 micro-controller with 256 bytes RAM 8 channel 10 bit ADC, 2 8 bit PWM, Timer 2 with capture & compare
83C652	8051 micro-controller with 256 byes RAM Same as standard 8052
87C751	24 pin package
87C752	28 pin package
84C00	8048 256 byte RAM. Bond out version.

Fig. 1. In hardware terms, I²C bus is relatively easy to implement. Since there can be many devices on its two-way, two-wire bus however, software drivers are more involved.



given the state of bit 7 prior to raising the SCL lead thus the transmitting contents of the data register to the bus in a serial form.

The routine assumes that the control routine will monitor the number of bytes to be sent. If a NACK is received from the slave or after all the bytes are transmitted to the slave, a stop or repeated start routine must be executed.

To receive a byte the slave should first be addressed by using the transmit routine. Each time the SCL lead is set high, bit 0 of the data register is loaded with the state of SDA by the receive routine. Data is then shifted left until all eight bits are loaded.

Receiving assumes that the control routine indicates whether to send an ACK or a NACK after receipt of each byte. After sending a NACK signal, the control routine must then execute either a repeated start or stop to terminate the receive procedure. The stop routine assumes that SCL is low, i.e. that the start, transmit or receive routines have been previously executed.

Note that these routines produce a relaxed form of the bus protocol. Checks for arbitration and clock synchronisation are not performed. Under the I²C patent, buying Philips I²C components conveys a license to use the devices provided that the system they are used in conforms to the true I²C specification.

Where speed is not important, the routines can be written in a high level language such as Basic and run via a microcomputer. For more serious applications, there are microcontrollers with dedicated I²C input/output lines and internal functions.

There is also a dedicated I²C bus interface for use with almost any 8 or 16 bit microprocessor or microcontroller, the *PCD8584*. It provides for multi-master, multi-slave configurations and can be accessed as a DMA (direct memory access) port.

I²C frequency synthesiser

One of several I²C-controlled frequency synthesizers, namely the *TSA5510*, was mentioned in the article *Closing the loop*, *EW+WW* June '93. This device operates at frequencies to 1.3GHz.

In the design shown in **Fig. 2**, a *74HCT4046* provides the voltage-controlled oscillator and a *TSA6057* radio tuning frequency synthesizer acts as the I²C bus device to keep cost down and simplify the circuit.

Control of the *TSA6057* is achieved by sending instructions to five write-only control registers, **Table 2**. A low cost *80552 Cameo* development system connected to a PC compatible produced the I²C bus signals.

Made by TDR, the *Cameo* board is a low-

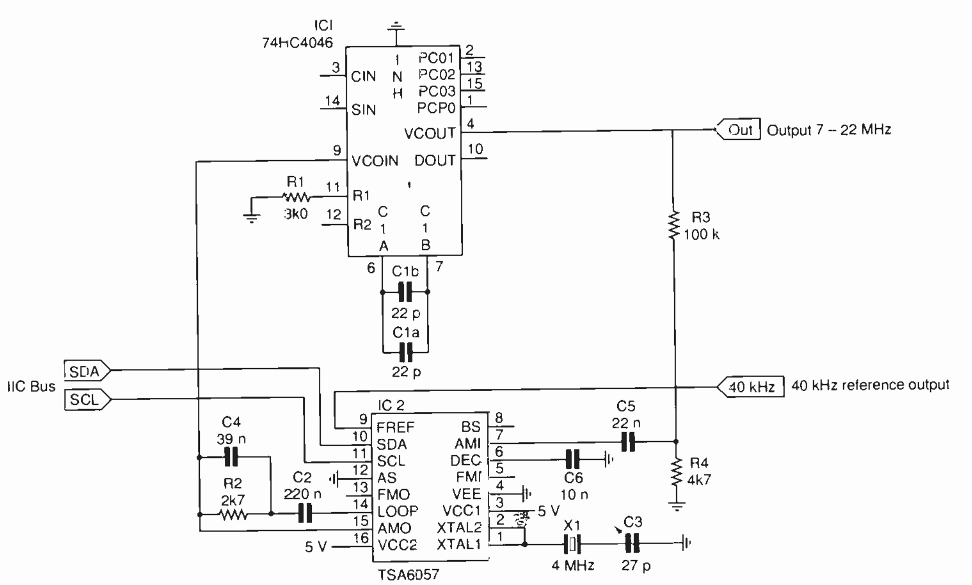


Fig. 2. Although this frequency synthesiser operates at megahertz frequencies, the signals needed to control it can be conveniently conveyed via the economical I²C bus.

cost system for developing and testing *8051* source code. Its monitor program has an I^2C driver built in. The system is supplied with an applications program which uses this driver to provide a simple RS232-to-I^2C converter.

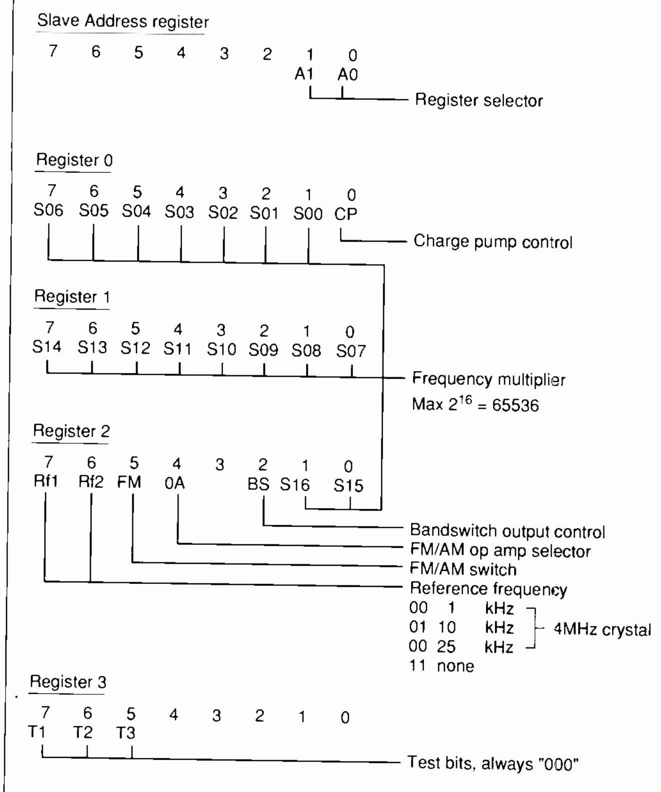
Commands sent from a communications program called Procomm are loaded via a PC RS232 port to the *Cameo* board. They then pass via *Cameo's* I^2C bus program to the *TSA6057*. Using these commands, oscillation frequency of the VCO is variable from 7 to 22MHz.

Setting of a particular frequency is easy. First, the *Cameo* I^2C-to-RS232 converter is loaded. An interrogator function then checks for slave devices by sending out addresses and looking for a response. Finally, the slave address is set.

To set a frequency of 8.192MHz for example, i.e. a 1kHz reference frequency multiplied by 8192, the internal registers need to be set accordingly.

Subaddress
0016 ;start at address 0
Data byte 0
0016 ;C_P=0, S_{0-6}=0
Data byte 1
4016 ;S_{13}=1
Data byte 2
1016 ;AM with S_{15-16}=0
Data byte 3
0016 ;test bits=0

Table 2. Control of the frequency synthesiser Fig. 2 is carried out on chip via five write-only control registers.



Using the converter command T0000401000, data transmitted to the *TSA6057* prototype resulted in a frequency reading of around 8.192MHz. Other code combinations were tried and the frequency output varied proportionately to the sent code.

I noticed considerable frequency jitter which was due to the loop coming in and out of lock. As I took no particular care with board layout and capacitor choice this jitter was not surprising. The prototype circuit did prove, however, that a synthesiser using the *TSA6057* was a practicable proposition provided care was taken in the layout and choice of components.

Date and time via I^2C

Where accurate time of day and calendar data are needed the *PCF8583* provides a low power source. The eight-pin device has a four year calendar, giving time and date to a resolution of 10ms, and a 256byte ram. An alarm function can be set to any time up to four years away. There is also a timer which can be set to count an elapsed time up to 99 in seconds, minutes, hours or days, **Table 3**.

An open-drain interrupt pin can be configured to pull low when a clock or timer alarm is activated and there is provision for allowing two devices to be connected to the bus. Clock, calendar, alarm and timing functions use the first 16 bytes of the internal ram, the remaining 240 bytes are available to the user.

While in operating mode, maximum current consumption of the device is 50μA. Current taken during access by the I^2C bus is 200μA maximum. Timing can be derived from either a 32.768kHz crystal or a mains-derived 50Hz signal.

One application, which is not immediately obvious, is the provision of a 'wake-up' facility on systems where current consumption is important. For example, during the period between taking samples, automatic data logging equipment normally performs no function. During these periods, power to all system components except the *8583* can be turned off to conserve the batteries.

By using the alarm function and interrupt output of the *8583* it is possible to provide a hardware power switch on to devices previously switched off.

Power down and power up facilities may be achieved as follows. Prior to switch off, write the relevant status data into the *8583* ram, i.e. what the processor should do next after waking up. Set the alarm or timer to activate at the required time in the future. Go to the switched off condition by programming the processor to remove the power-on latch.

On receipt of the wake up signal, the *8583* interrupt goes low, turning on the power. Power is latched on when the processor powers up. Next, status and any other relevant information is read from the *8583* ram. At the end of operating period, the system is switched off as explained above.

Driving leds

Primarily the *SAA1064* led driver is intended for controlling four seven segment leds. However it has the capability to drive 16 wires with maximum current of 21mA at up to 15V. The only restriction is the maximum chip dissipation of 1W.

The device makes an excellent relay or lamp driver. Functions of the eight internal registers are given in **Table 4**. Up to four devices may be connected to a single bus.

Arranged as constant current drivers, the 16 outputs can be controlled to sink current up to 21mA. Output current can be set in steps of 3mA from 0 to 21mA. Provided that the maximum chip dissipation is not exceeded, each output can drive a led or incandescent lamp without needing a series resistor.

Four different slave addresses can be chosen.

Provision is made to multiplex the 16 outputs to allow for 32 lamps or segments. In this mode average current per output is 50% of the set current. Both multiplex outputs are active high and can drive a minimum of 50mA.

Make I^2C go further

Open-collector And connections are used to implement the I^2C bus. These allow flexibility but at the expense of limiting interconnecting bus-line lengths. This is why it is called the inter-integrated-circuit bus.

All devices on the bus are guaranteed to work only if the rise time and fall times of the electrical signal on the SDA and SCL lines are below a certain minimum. Open-collector drivers need a resistor to pull them to V_{cc}. Total capacitance presented to this bus pull-up resistor restricts rise and fall times.

Total bus capacitance is the sum of input capacitances of all connected devices plus capacitance of the bus wire. As the distance between devices increases, capacitance of the bus wire also increases. Signal rise time depends, therefore, on the number of devices present and the capacitance of the bus wire.

Under normal circumstances bus capacitance limits the distance over which satisfactory data transfer can be performed.

Transmission line techniques can be used to assure that bus rise and fall times meet the I^2C bus specification over any length of line. Propagation velocity of the line then becomes the limiting factor, restricting maximum bit rate.

I^2C specification summary

Max. bus frequency	100kHz
Min. SCL high time	4.7μs
Min. SCL low time	4μs
Max. input rise time	1μs

Table 3. In addition to real-time clock information, this I2C-controlled IC has a 256byte non-volatile ram.

Clock Function

Dev. Addr	Bit Number	Description
00	7654 3210	**Control/Status**
	bit 0	Timer Flag
	bit 1	Alarm Flag
	bit 2	Alarm Enable
	bit 3	Data & Month Mask
	bits 5,4	Function. 00 Clock 32kHz Xtal; 01 Clock 50Hz Mains; 10 Event Counter; 11 Test
	bit 6	Hold Count
	bit 7	Stop Count
01	7654 3210	**Function of seconds**
	3210	Hundredths
	7654	Tenths
02	7654 3210	**Seconds**
	3210	Units
	7654	Tens
03	7654 3210	**Minutes**
	3210	Units
	7654	Tens
04	7654 3210	**Hours**
	3210	Units
	54	Tens
	6	AM/PM Flag
	7	12/24 hour format
05	7654 3210	**Days/Year**
	3210	Units
	54	Tens
	76	Year
06	7654 3210	**Months**
	3210	Units
	4	Tens
	765	Weekdays
07	7654 3210	**Timer**
	3210	Units
	7654	Tens

Alarm Function

Dev. Addr	Bit Number	Description
08	7654 3210	**Alarm Functions**
	bits 2,1,0	Timer Control: 000 No timer; 001 hundredths; 010 seconds; 011 minutes; 100 hours; 101 days; 110 not used; 111 test
	bit 3	Timer interrupt enable
	bits 5,4	Clock Alarm Enable: 00 no alarm; 01 daily alarm; 10 weekly alarm; 11 dated alarm
	bit 6	Timer alarm enable
	bit 7	Alarm interrupt enable
09	7654 3210	**Alarm fraction**
	3210	Hundredths
	7654	Tenths
0A	7654 3210	**Alarm seconds**
	3210	Units
	7654	Tens
0B	7654 3210	**Alarm Minutes**
	3210	Units
	7654	Tens
0C	7654 3210	**Alarm Hours**
	3210	Units
	54	Tens
	6	AM/PM Flag
	7	12/24 hour format
0D	7654 3210	**Alarm Days/Year**
	3210	Units
	54	Tens
	76	Year
0E	7654 3210	**Alarm Months**
	3210	Units
	4	Tens
	765	Weekdays
0F	7654 3210	**Alarm Timer**
	3210	Timer units
	7654	Timer tens of units

Table 4. Although the SAA1064 is intended for driving two seven-segment leds, its flexible control registers and 21mA drive capability open-up other applications.

Instruction byte

Bit No. 7	6	5	4	3	2	1	0	
0	0	0	0	0	SC	SB	SA	Select Register
					0	0	0	Control Register
					0	0	1	Digit 1
					0	1	0	Digit 2
					0	1	1	Digit 3
					1	0	0	Digit 4
					1	0	1	Reserved
					1	1	0	Reserved
					1	1	1	Reserved

Control byte

Bit No. 7	6	5	4	3	2	1	0
0	C6	C5	C4	C3	C2	C1	C0

- C0: 0 = Static Mode. Digits 1 & 2 only; 1 = Dynamic Mode. Digits 1 – 4
- C1: 0 = Black digits 1 + 3; 1 = Display digits 1 + 3
- C2: 0 = Black digits 2 + 4; 1 = Display digits 2 + 4
- C3: 1 = All outputs "on" (Test)
- C4: 1 = add 3mA to output current
- C5: 1 = add 6mA to output current
- C6: 1 = add 12mA to output current

Data Digit 1

Bit No. 7	6	5	4	3	2	1	0
P8	P7	P6	P5	P4	P3	P2	P1

Data Digit 2

Bit No. 7	6	5	4	3	2	1	0
P16	P15	P14	P13	P12	P11	P10	P9

(Data Digits 1 and 2: Controls Specified Output Pins in non multiplexed mode or when Mx1 is high in multiplexed mode)

Data Digit 3

Bit No. 7	6	5	4	3	2	1	0
P8	P7	P6	P5	P4	P3	P2	P1

Data Digit 4

Bit No. 7	6	5	4	3	2	1	0
P16	P15	P14	P13	P12	P11	P10	P9

(Data Digits 3 and 4: Controls Specified Output Pins when Mx2 is high in multiplexed mode)

END.

Max. input fall time	300ns
Time SDA must be stable prior to rising edge on SCL	250ns
Max. input low voltage	1.5V
Min. input high voltage	3.0V
Max. output low voltage	400mV
Typ. high leakage current	10μA
Max. input capacitance per device wire	10pF
Max. bus capacitance	400pF

A start condition is defined as a falling edge on the SDA line while SCL is high. Similarly, a rising edge on SDA when SCL is high defines the stop condition.

Bus capacitance limitations

In practical systems, bus capacitance is either known or can be estimated. A good approximation is to assume each device on a single circuit board presents 20pF. This comprises 10pF input capacitance and 10pF track capacitance. Capacitance per metre of interconnect needs to be obtained from the manufacturer's data. Typically, 75Ω television antenna cable is 70pF/m.

A good approximation for bus voltage rise time is obtainable by multiplying the pull-up resistor value by the bus capacitor. This value should not exceed 1μs. Although not strictly correct, the *CR* time constant gives a near enough approximation for most applications.

The minimum value of the pull-up resistor is restricted by pull-down current which should not exceed 3mA for unbuffered devices. Account of the 10μA leakage current from each device on the bus needs to be taken when calculating the pull-up resistor.

It is sometimes necessary to protect devices from bus voltage transients by inserting a resistor in series with the device pin. This further complicates the calculations. Further information on this subject is available in Philips' data sheets.

Propagation velocity restrictions

Due to the I^{2}C bus open-collector And function, all master transmitting devices monitor their own signals to detect for the presence of other masters. A received logic low when transmitting a high indicates the presence of another device.

In most systems, the delay caused by the time taken for an electrical signal to pass down a wire can be ignored. When attempting to use the I^{2}C bus over long distances, transmission delay time of the bus wires limits the data rate.

When calculating the maximum line length for a given data rate, response and hold times of all devices have to be taken into account, in addition to propagation delay.

The acknowledge signal is a typical example. A low acknowledge signal on SDA from the remote end must be received 250ns prior to the local end SCL rising edge. At 100kbaud, the local SCL clock is low for

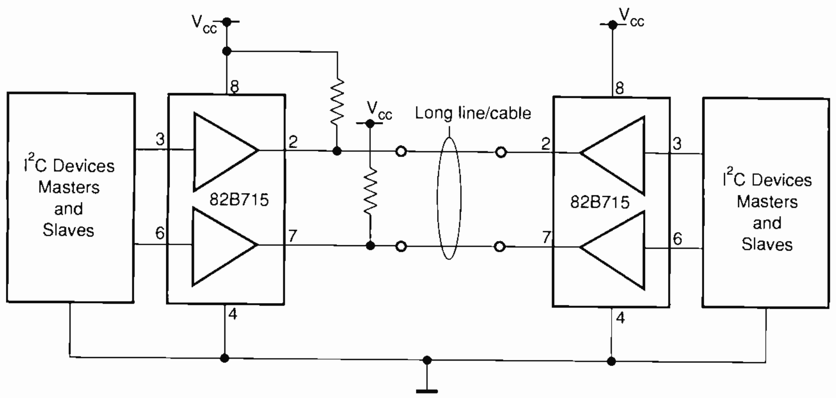


Fig. 3. Adding current buffers at each end of the I²C link extends communication distance considerably. It also allows cheaper cabling.

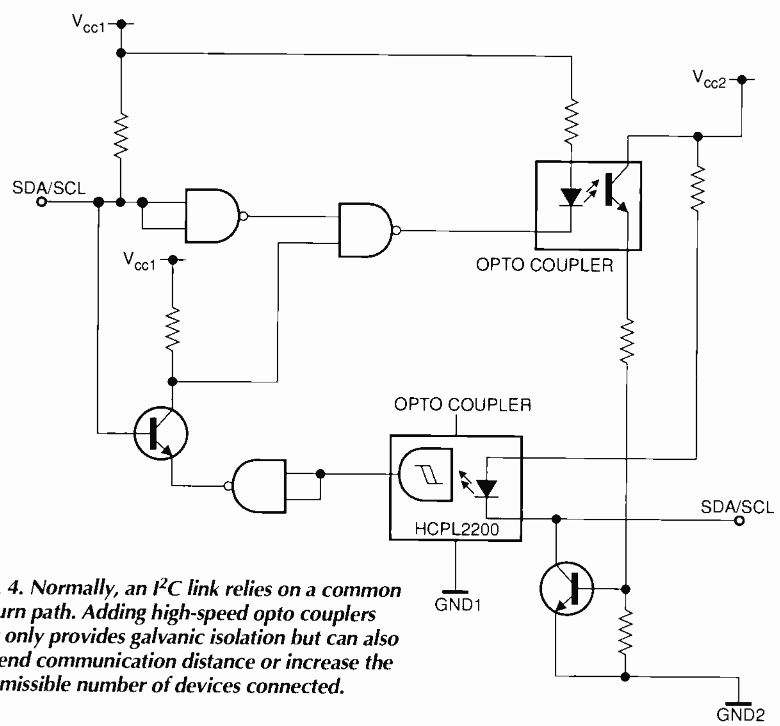


Fig. 4. Normally, an I²C link relies on a common return path. Adding high-speed opto couplers not only provides galvanic isolation but can also extend communication distance or increase the permissible number of devices connected.

4.7μs. This leaves 4.45μs, from the falling edge of SCL to detect the state of the distant SDA signal.

Where propagation velocity is light speed, maximum bus length is 4.45μs×300/2=667m. This calculation ignores device response time and the velocity factor of the cable. It is unlikely that a line length of 500m can be exceeded for a rate of 100kbaud.

Dedicated I²C bus extender

Within the *82B715* bipolar integrated circuit are two identical buffer circuits. These enable the I²C bus to be extended over longer distances without degrading performance.

Each buffer preserves the bi-directional open-collector characteristic of the I²C SDA and SCL lines. The circuit is supplied in plastic eight pin DIL or small outline packaging.

Practical communication distances are limited to a few metres by the I²C bus capacitance limit of 400pF. Using one *82B715* pair at each end of longer cables reduces effective loading capacitance on the bus. This allows communication via low cost general-purpose wiring.

Each of the buffers has a current gain of ten. Impedance seen at an input is therefore ten times the output (line) impedance.

Figure 3 shows a typical minimum system configuration.

Galvanic isolation

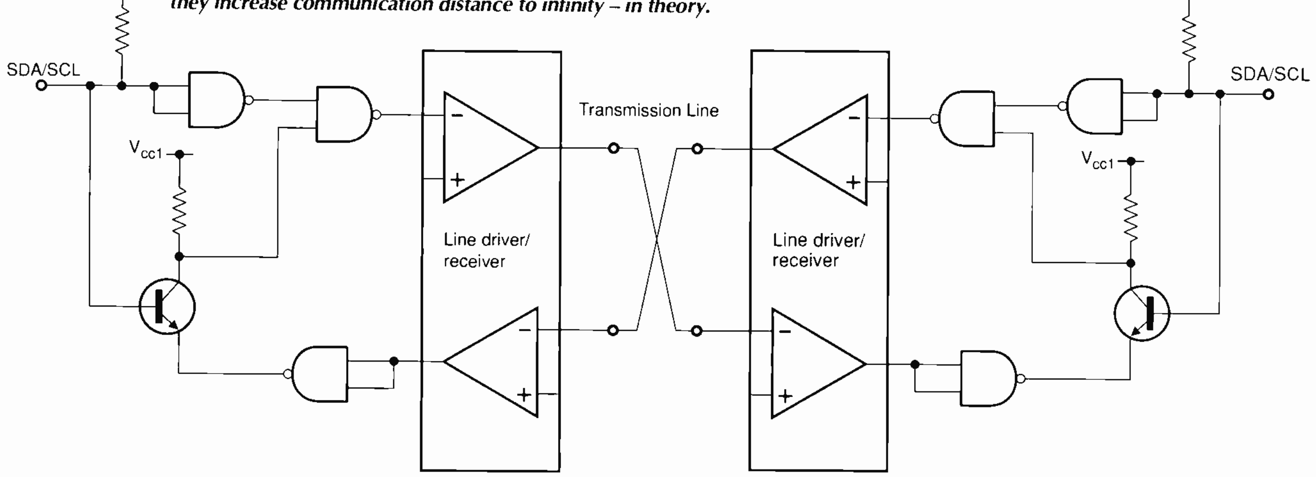
The I²C bus assumes that there is a common return between all connected devices. It is sometimes necessary to run two or more systems each with their own independent power supply.

Under these circumstances, a direct connection between the different systems is undesirable, impossible or even dangerous.

High speed opto-couplers now available make it possible to provide a galvanic barrier between I²C bus systems, **Fig. 4.** Two such circuits are needed, one each for the SCL and SDA lines. Note that bus capacitance on one side of the isolator does not reflect to the other side.

By using two or more sets of isolators, one

Fig. 5. Transmission-line techniques add a little complexity to I²C but in return they increase communication distance to infinity – in theory.



at each end of the bus, it is possible to extend the distance between systems and/or increase the number of connected devices beyond the 400pF limit.

I²C via transmission-line

An extension of the previous isolation principle is to separate send and receive signals for each of the SCL and SDA signals and send them down independent transmission lines, **Fig 5**.

With this arrangement the distance between ends can approach infinity. The only limitation is the velocity of propagation of the transmission lines which restricts the maximum baud rate. The main disadvantage of this system is the cost of two transmission lines per I²C bus lead.

Summary

The I²C bus was created to provide a low cost, yet adaptable intercommunication medium for use in the television and audio market. Although it was originally intended to be an inter-integrated-circuit bus for devices on the same PCB, the range of about 4m made it a very useful media for data exchange between units on the same shelf.

Using *82B715* buffer ICs can extend the range to several tens of meters making the bus useful for control and monitoring functions in buildings. ■

Further reading

The following Philips component application notes and data handbooks provide definitive information on the I²C bus specification and available devices.

- *I²C bus specification*, 9398-336-70011.
- *I²C bus compatible ICs*, book 4, parts 12a & 12b 1989.
- *I²C peripherals for microcontrollers*, 1992.
- *Single chip 8-bit microcontrollers, PCB83C552/562* handbook.
- *Interfacing PCD8584 I²C*, application note, Feb. 90.
- *Applications report PCF8584, I²C-Bus Controller, Mar 93*, 9390-316-20112.
- Busman's Guide to I²C, *Electronics World & Wireless World*, Jan 94.

Other publications of interest:

- *Embedded Applications Journal No 1*, Intel.
- EPROMS, EPOTs, NOVRAMs Data Book Xicor.
- Cameo Applications, TDR Ltd.
- I²C Bus Specification, Philips
- 82B715 Data Sheet, Philips

I²C driver on disk

An assembly-language listing for a simple I²C driver is available on disk. This listing is written specifically for the *PCB83C552* controller in 8051 code. It can be obtained by sending £10 all inclusive to Quadrant House, The Quadrant, Sutton, Surrey SM2 5AS. Please mark your envelope I²C software.

I²C bus devices and data sheets can be obtained from Philips and their distributors. I have found Gothic Crellon at Wokingham, Tel. 0734 788876 and Quarndon at Derby Tel. 0332 332651 very helpful.

Help and further information on bus devices and the Cameo board may be obtained from me at TDR Ltd. Tel 0666 577464. TDR can obtain specific devices for experimentation.

Arnold Sugden – pioneer of single-groove stereo

Had vertical/lateral groove modulation won the battle for a stereo disc reproduction standard, Connoisseur might have been a leading record label and Arnold Sugden a leading figure in the industry. As it is, the company and its founder – pioneer of single-groove stereo recording – have drifted into obscurity. Reg Williamson looks at Arnold Sugden's achievements.

*Reg Williamson is a visiting lecturer at the University of Keele.

Arnold Sugden – founder of Connoisseur and pioneer of single-groove disc recording.

For those of us active in high-quality audio engineering, the 1950s were exciting times. The tautological term 'hifi' had yet to find its way across the Atlantic, and the few commercial products that were available were produced by small specialist companies run by enthusiasts – among them Leak and Quad.

One such small company was Connoisseur, based in Brighouse, West Yorkshire. It was there, in 1953, that one of the most important steps forward in domestic sound reproduction was made, namely the single groove stereo disc. Sadly, the man who originated that development – Arnold Sugden – is now all but forgotten.

The idea of recording stereophonic sound on disc was not new. In the USA, Emory Cook was already establishing himself with his 'binaural' discs. These had a separate tracks for each channel. As a result, they required two pickups on the disc started simultaneously to reproduce the recorded material. Although they made it impossible to preserve phase coherence, Cook's records were novel.

In 1933, the system that modern techniques are based on was patented by Alan Blumlein. He was working in the British laboratories of EMI – parent company of the familiar 'dog and horn' record label. Similar, but in my opinion less impressive work was also being carried out at Bell Laboratories in the USA. At the time EMI, with what was to become its habitual lack of commercial foresight, saw no future in it. In any case, the imminence of war forestalled any further development.

Blumlein's patents languished until around 1958, when an international standard was eventually agreed using his proposed 45°/45° system of groove modulation, shown in the diagram. Paradoxically, most of EMI's competitors demonstrated their enterprise by marketing records in that same year.

Blumlein's old company however misguidedly placed all its commercial faith in 19cm/s two-track stereo tapes – their Stereosonic system. Inevitably, because of the cost of both the software and the hardware, it failed. Belatedly, the company conceded defeat and began to produce stereo long-playing records the same as those of its competitors.

Arnold Sugden was the enthusiastic managing director, technical director and owner of the Connoisseur company. For some time, the company had been enjoying considerable commercial success in producing many high quality record playing products.

One of Sugden's most notable products was the first British belt-driven turntable. This subsequently became available in kit form and was an unusual introduction in the new market for such audio products. Even today, the BD1 turntable is much sought after by enthusiasts.

Practically everything to do with sound reproduction and recording was tackled by the tireless innovator. Connoisseur's first post-war product designed by him was a miniature lightweight cartridge and arm for quality reproduction of the old 78rev/min discs. He even invented a technique for mass producing the miniature chrome tipped steel needles that were in very short supply just after the war.

Pickup arms, cartridges, amplifiers, loudspeakers – even microphones – all appeared bearing the Connoisseur label. As early as 1950, Connoisseur was manufacturing a combined 33 and 78rev/min turntable, a high quality pickup with interchangeable heads and a twin-speed disc recording system.

All this is the more remarkable when, as I learned during a visit to see Arnold Sugden, he had left school at fourteen and had never had any formal engineering training. But, from the outset of his career, he demonstrated a natural flair for engineering design of an intuitive kind that is all too rare these days.

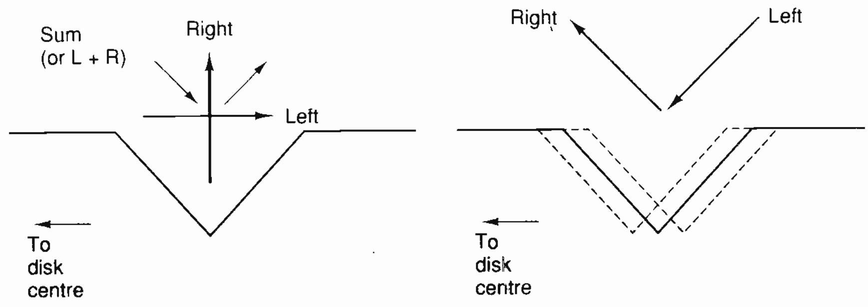
An engineer with a passion for music

Sugden's interest in high quality audio was, in common with many of us in those days, linked with a passion for music. This is no longer considered an essential pre-requisite. Shortly after the introduction of the long-playing record, he became convinced that there was a future for single-groove stereo on disc.

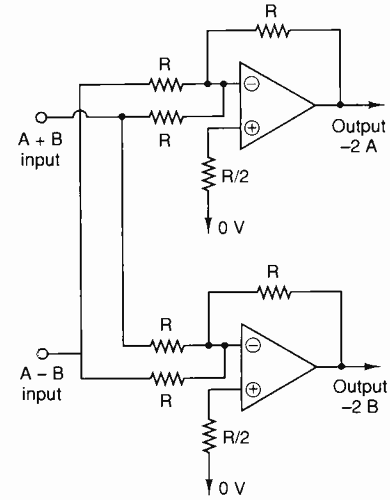
In 1953, he set to work in a disused church near the Connoisseur factory, often working alone until the early hours of the morning. He finally perfected a stereo cutter head. It was designed to be fitted to a disc-cutting table the company had already developed.

By methods both intuitive and empirical – an approach that would be frowned on today – he evolved a remarkable design that one cannot easily fault. He already knew that the mass of the cutter had to be as low as practicable. This mass, linked with compliance of the movement, has to push the first fundamental resonance of the system as high as possible.

His use of balsa wood as the former materi-



Sugden's vertical/lateral recordings, left groove, involved a 45°/45° pick-up cartridge with electronic decoding. On the right is the 45°/45° standard showing each channel modulated identically and in phase. Basic sum and difference matrix circuitry for Sugden's vertical/lateral technique is also shown.



al vividly demonstrates his unorthodox, but imaginative approach. With it he achieved a fundamental system resonance of 4.5kHz, and it was well damped. This he corrected by a tuned circuit in the driver amplifier.

An equally unusual feature of the system was its efficiency, requiring as little as 5W to fully modulate the cutter assembly. A heavy, powerful magnet was mounted separately on the cutter carrier. Only the comparatively light weight of the cutter assembly rested the stylus on the surface of the master lacquer disc.

A cartridge capable of playing his records had to be designed and made. Sugden's solution clearly demonstrates his uninhibited imagination. At that time, there was a very popular monophonic crystal cartridge. It was designed by Stanley Kelly for the Cosmocord company and called the Acos. It was very much ahead of its time, being the first crystal type with any claim to a wide frequency response and low tracking weight.

Sugden simply took two of these cartridges, mounted them together at right angles, and coupled the two styli with a fine wire link. He mounted this combination in a special moulding. As a short-term solution to an engineering problem, it was an extraordinary approach.

Inevitably however, later tests showed that its performance was markedly inferior to that of the cutter. So Sugden went on to design a high-quality crystal cartridge himself, which he marketed for the emerging 45°/45° discs.

First stereo demonstrations

By 1956, Sugden was ready to demonstrate his system. He had produced master tapes using spaced Neumann U47 capacitor microphones and a tape recorder fitted with a staggered extra head. Learning from his experience, Sugden later reduced the spacing in his miking technique. He used as little as 6in separating the pair and angled them outwards by approximately 90°. This basic configuration is still favoured today by some purists.

While he had finally decided on vertical and lateral groove modulation, his cutter could also be used for 45°/45°, given an appropriate matrix. Since doubts about the validity of Blumlein's original patents had emerged, the groove modulation had to be vertical/lateral, for the time being at least.

No conventions or standards existed at that time. On the basis that the most critical sounds would be from the string section of an orchestra, Sugden determined that the lateral cut should carry the left channel. Additionally, there appeared to be no advantage for any particular phase relationship between channels, so long as it was always consistent with the arbitrarily chosen standard. Incidentally, the lack of knowledge of what the original phase relationship was created some difficulty when attempts were made to re-matrix some of his original pressings for playback.

In the preceding three years, Sugden had recorded a diverse range of material locally. This included orchestral concerts, brass bands – a speciality of his native county, Yorkshire – and cinema organs. From this wide variety, he subsequently cut lacquer stereo discs.

Now, he looked for a platform to demonstrate his achievement to his peers. Commercial hifi shows of the kind common today were non-existent in 1956. Most new developments were demonstrated every year in London, at a two day exhibition. It was organized by what was then called British Sound Recording Society which was eventually absorbed in to the AES British Section.

In audio advancement terms, 26 May 1956 was a particularly memorable date. At that same show, Sugden first introduced his stereo discs. Peter Walker of Quad also demonstrated his prototype electrostatic speaker – another significant development in audio design. As you might expect, there were many queuing to hear these twin wonders, including me.

To put this into perspective, neither of the two major record companies in the UK were, at that time, showing any great interest in a disc system based on vertical/lateral or 45°/45° techniques. As mentioned earlier, EMI had placed its faith in tapes. Decca was working independently on an hf carrier system for vinyl discs, presumably to bypass the Blumlein patents, but with little urgency. In passing, the original Decca pickup was designed for this system and it is still popular.

To say Sugden's demonstration caused a stir would be an understatement. Almost immediately, he was besieged in Brighouse by all the record companies anxious to take advantage of his pioneering work. EMI even sent a large mobile recording studio to his factory, so that sample discs could be cut from a variety of early EMI stereo master tapes.

Many of these masters were subsequently pressed, revealing some difficulties in the pressing technique. The problems to overcome these too, were given a demonstration at the BSRA exhibition of 1957. Having acquired a pair of these pressings and using a suitable active matrix, I have been able to play them with an orthodox 45°/45° cartridge; and very impressive they are, too.

Even the record sleeves were ready

Had vertical/lateral recording been adopted as the standard, there might now have been a Connoisseur label on the market. Four discs were in fact prepared, ready for issue, and six more planned for release. Even the sleeves were designed by the enterprising Arnold.

But it was not to be. Illness, probably brought on by overwork, slowed all progress on producing stereo discs. Eventually, it had to be dropped altogether when commercial stereo discs to the universally agreed 45°/45° standard were released in 1958.

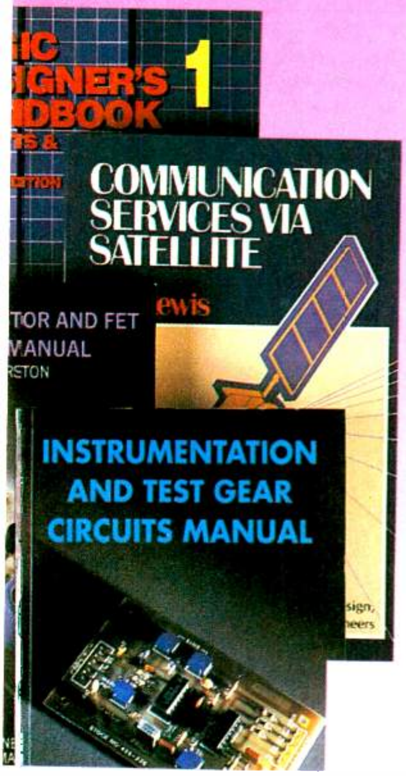
As 'hifi' arrived and highly competitive, consumer-orientated products began to appear from the Far East, the viability of Sugden's company was diminished. In the early 1970s, Arnold Sugden sold his company and retired.

The final chapter in the story of his enterprise ends on a sad note, for the company fell into the hands of asset strippers. Now, the once familiar name of Connoisseur has disappeared and the historically-important hardware is lost – an unhappy, but all too common tale in today's commercial world of hifi.

However, not so for the name of Arnold Sugden. Within audio history, he has written his own personal paragraph and undoubtedly earns a prime place amongst the pioneers. Together with many of his contemporaries, he advanced the science and art of sound recording and reproduction to a significant degree.

In the preparation of this article I gratefully acknowledge the supply of drawings and much information from the veteran Arnold Sugden himself, now into his 82nd year and to whom I offer this as a warm tribute. ■

BOOKS TO BUY



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Synchronous oscillators: Alternative to pll?

The synchronous oscillator is a fascinating circuit to play with, offering as it does an alternative approach to the PLL for signal extraction in a noisy environment. The possible variations on the theme are limitless. Ian Hickman investigates.

The phase-locked loop is a tried and tested circuit, well understood, capable of extracting signals buried in noise. However, it is not the only way of pulling out a wanted signal from noise. An ISSCC report on the efficiency of the synchronous oscillator prompted me to investigate the circuit further.

It is well known that an oscillator can be synchronised with an external signal of the same (or very nearly the same) frequency. This applies to oscillators of all sorts so that, for example, a Wien bridge audio oscillator can be locked over a small range of frequencies to a signal injected into the maintaining amplifier. This signal does not need to be a sinewave; the oscillator will lock to a low-level squarewave while delivering its normal sinewave output... in effect a high Q filter.

At rf, a weakly oscillating LC oscillator can be used in a similar way – a scheme which under the name "reaction" (or, in the USA, the more colourful term "tickling") is as old as the hills. Where the synchronous oscillator differs from this scheme is in the means by which the external signal, to which it is desired to lock the oscillator, is injected.

Fig. 1 shows the basic circuit of a synchronous oscillator as it appears in Ref. 1 and several other of the referenced papers. Here, Tr_2 is arranged as a Colpitts oscillator, with its emitter current supplied from Tr_1's collector. In the absence of any external synchronising input, the oscillator runs at the frequency determined by its tank circuit. An applied external signal modulates Tr_1 collector current to a lesser or greater extent, even chopping it up into pulses in the case of a large applied signal. This modulation will synchronise the oscillator if the injected signal is close enough to the free-running frequency, though with a standing phase difference between the injected signal and the oscillator's output of up to ±90° over the extremes of the frequency range for which lock is maintained.

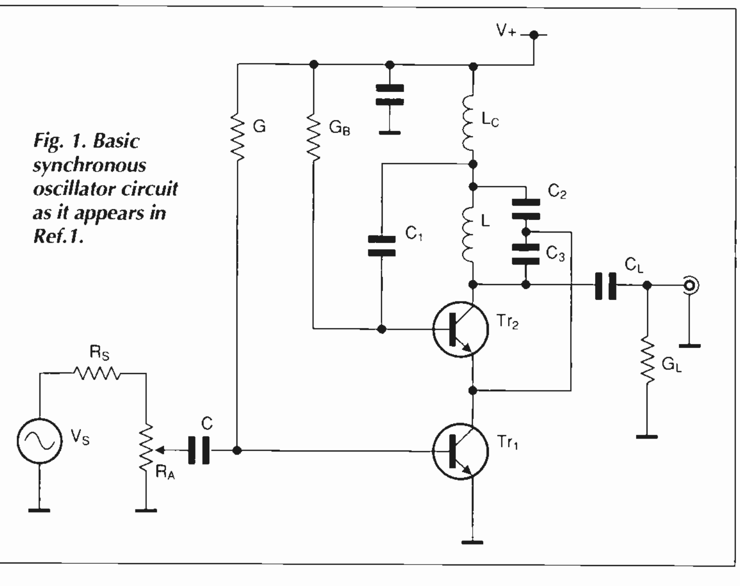


Fig. 1. Basic synchronous oscillator circuit as it appears in Ref.1.

Critical dc operating point

The stated theory of operation points out that the oscillator transistor Tr_2 is designed to run biased well into class C, with a small conduction angle, with the result that it is insensitive to noise accompanying the externally applied signal for most of each and every cycle.

Anyone used to analogue design will instantly see some shortcomings in Fig. 1, in particular the very poorly defined dc conditions, due to the use of high resistance bias sources for two devices in series. The author of Ref. 1 states that the circuit is set up so that the transistors have equal collector-emitter voltages and thus both run in a linear regime. While it is possible to select transistors or bias-resistors or both to achieve this, tracking with varying temperature and aging cannot be guaranteed even with transistors of the same type. Indeed, some of the circuits in the referenced papers actually use different transistor types at Tr_1 and Tr_2.

From Ref. 1 it appears that the circuit is run from +5V, which leaves only 2.5V V_{ce} for each device, assuming the biasing is perfectly balanced. Thus the maximum achievable tank circuit amplitude is about 4V peak to peak and the amplitude at the emitter of Tr_1 will typically be a third of this. So the conduction angle will in fact be significant.

The circuit of **Fig. 2** was therefore chosen for initial experiments. Here, the dc conditions are well defined, with the oscillator transistor's emitter current supplied from one half of a long-tailed pair. With the tank components shown, the free-running frequency was 893kHz. **Fig. 3a** shows the input waveform (lower trace) and the output (tank circuit) waveform with a signal input of 1V p/p at 875kHz (the lower end of the lock range) where the output is leading the input by 90°. Figs **3b and c** show the same at 893kHz (input and output in phase) and at 919kHz (output lagging by 90°) respectively. Fig. 3d shows the output of the synchronous oscillator with a swept input frequency.

To maintain the light circuit loading provided by the 'scope probe, the spectrum analyser was not connected to the circuit directly, but driven from the oscilloscope's Channel 2 signal output, which provides a 50Ω output impedance. The input signal was swept from just above the lower limit of lock to just above the upper limit. At this point numerous FM sidebands appear, mainly on the low frequency side, related to the rate at which the synchronous oscillator slips cycles in an unsuccessful attempt to retain lock.

Note that the amplitude of the oscillator's output is a maximum at the centre of the lock range, i.e. at the frequency at which it free-runs in the absence of an input. Ref. 6 states that this is a characteristic of an injection-locked oscillator, whereas in a true synchronous oscillator, the output amplitude is constant over the whole lock range. Why the Fig. 2 circuit exhibits this departure from the expected synchronous oscillator performance I am unable to say.

Another aspect of the synchronous oscillator's characteristics is its ability to lock to an input frequency which is an integer multiple (harmonic) of its output frequency or to a submultiple. This behaviour is shown in **Fig. 4a** and **4b** respectively, again using the Fig. 2 circuit. In each case, the input signal amplitude was −10dB ref. 1V p/p. In 4a, the 890kHz output (upper trace) is locked to an input at three times that frequency, i.e. the circuit is operating as a divider. In 4b, the 890kHz output is locked to an input at one third of that frequency, i.e. it is acting as a tripler.

The apparent amplitude modulation of the output at one half of the input frequency is one of those little mysteries that crop up all the time in electronics. I would have pursued it further, but I began to feel that in some respects, the circuit of Fig. 2 departed from the basic synchronous oscillator philosophy. Thus, for example, unlike the SO of Fig. 1, the circuit will not readily act as a doubler, quadrupler, etc., since the balanced nature of the LTP stage feeding the Colpitts oscillator results in a minimal level of even harmonics in its output.

It was time to look at the synchronous oscillator circuit as published in Ref. 1.

Is it a synchronous oscillator?

I constructed a version of the basic synchronous oscillator circuit of Fig. 1 with Tr_1 and Tr_2 both *2N918*s (selected for equal h_{FE}). G=220kΩ, G_B=100kΩ, L=14μH, C_1=4n7, C_2 and C_3 both 100p and L_c = 22μH. This gave an operating frequency of around 6MHz, with a +5V supply rail.

Although the two transistors operated with approximately equal values of V_{ce}, the oscillator transistor was bottoming heavily, viz. not operating in a linear regime. The two bias resistors were raised first to 1M and 470k, then 2M2 and 1M, but still the oscillator bottomed. Clearly the operating dynamic resistance R_d

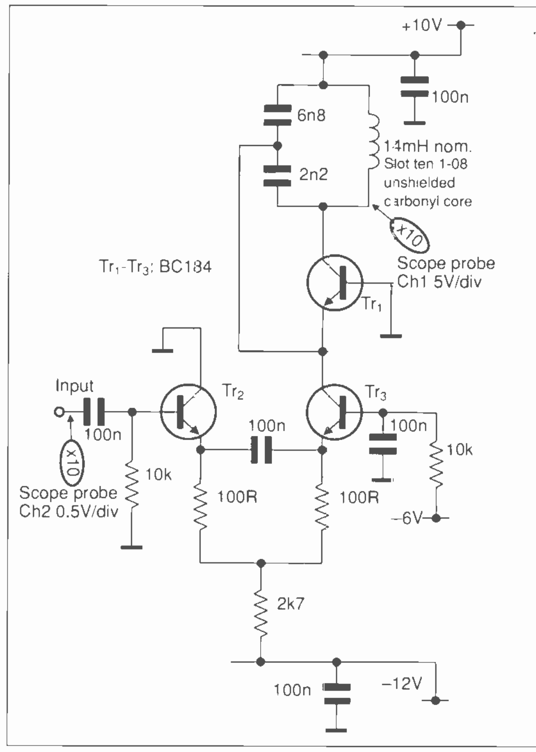


Fig. 2. Synchronous oscillator circuit used for initial experiments.

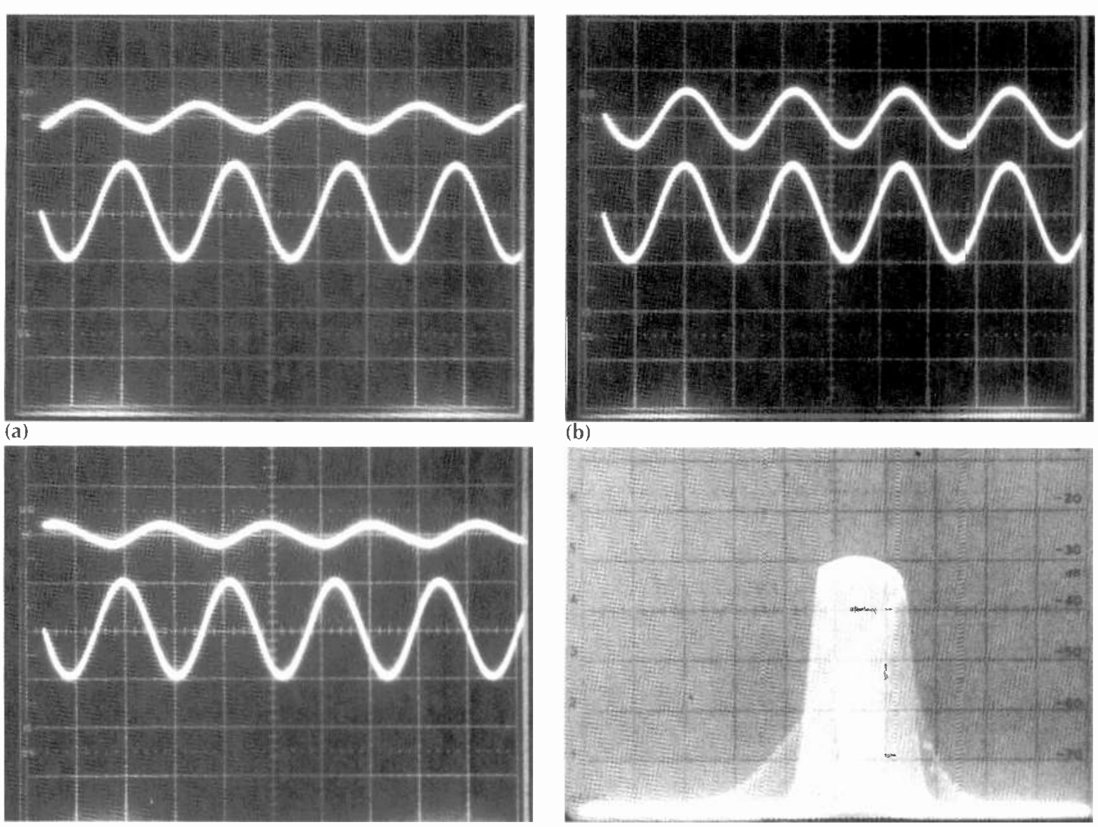
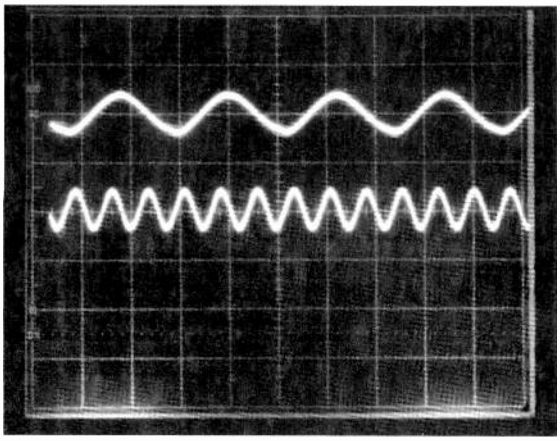


Fig. 3a. Input at 875kHz (lower trace, 0.5V/div) and output (upper trace, 5V/div) of a synchronous oscillator whose centre frequency is 893kHz, showing the output leading the input by 90° (0.5μs/div horizontal)
b. As a but at the 893kHz centre frequency
c. As a but input at 919kHz, output lagging by 90°
d. Showing the output when the input signal was swept from just above the lower limit of lock to just above the upper limit. At the latter, a whole range of out-of-lock sidebands appear. (Vertical, 10dB/div; horizontal, 20kHz/div, centre frequency 880kHz; IF bandwidth 3kHz, video filter off. Input to LTP −10dB ref 1Vp/p)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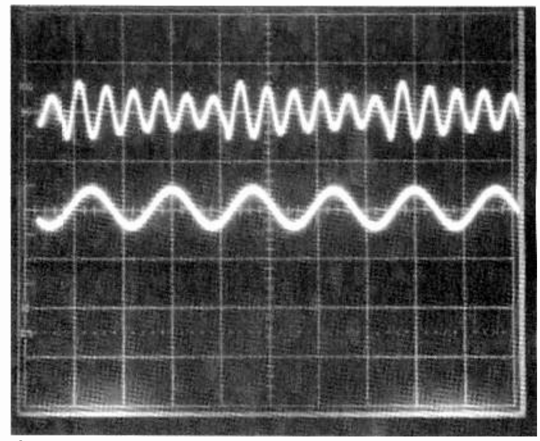

(b)

Fig. 4a. The synchronous oscillator of Fig. 2 operating as a divider, synchronised by an input (lower trace) at three times its output frequency. Scope settings as Fig. 3 a. At b. is the synchronous oscillator of Fig. 2 operating as a multiplier, synchronised by an input (lower trace) at one third of its output frequency. Scope settings as Fig. 3a except 1µs/div horizontal.

of the tank circuit was rather high, despite the heavy loading set by equal values for C_2 and C_3, but other worries about the circuit were beginning to niggle. For instance, although G_B in conjunction with the h_{FE} of the oscillator transistor sets the mean or dc value of the base potential, rf-wise, the base voltage can flap about. Tr_1 has a high output resistance and so will not define Tr_2's emitter potential, while the presence of the rf choke means that the top end of the tank is not clamped to the supply rail voltage (rf ground).

Is the rf choke really necessary, and if it is, what defines the rf voltage at the ends of the tank circuit – and hence the voltage delivered to GL, which in my case was a 'scope probe? Doubts about the choke were confirmed by Fig. 8 of Ref. 1, where a 10µH rf choke is used in a circuit operating at 560MHz. The reactance of an ideal 10µH inductor at 560MHz is over 35kΩ and it would resonate with a capacitance of just 0.008pF. Thus the impedance of the (unspecified) 10µH rf choke at the circuit's operating frequency is anybody's guess. Incidentally, this "improved" synchronous oscillator circuit has a bias source for the lower transistor consisting of 111mV from a 5kΩ source, so that in the absence of an input of considerable amplitude, the lower transistor will be cut off, supplying zero emitter current to the Colpitts oscillator transistor.

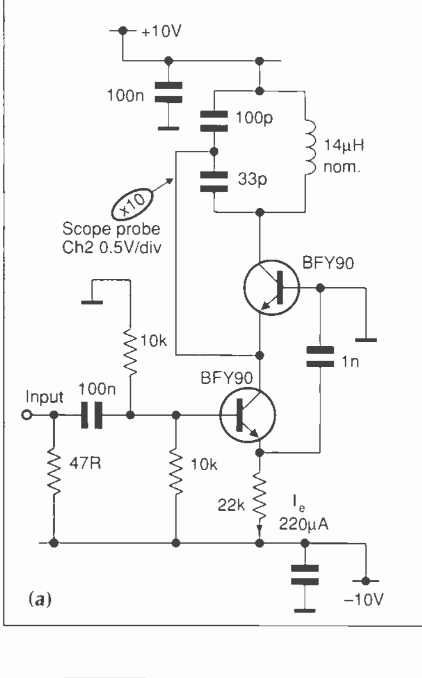
On to a new circuit for investigation, **Fig. 5a**. As in Fig. 2, the oscillator transistor's emitter current is closely defined, while the emitter of the lower transistor is at ac ground as in Fig. 1. With no external input, the circuit oscillated at about 6.5MHz, with 10V p/p at the collector of the upper transistor, a little over 2V p/p at the emitter, **Fig. 6a**.

The double exposure also shows the 'scope inputs grounded, indicating the emitter reaching a peak negative-going excursion of –0.8V relative to ground (upper trace) and the collector voltage swing centred about the +10V rail. Thus the oscillator transistor operates with a narrow conduction angle and the collector voltage well clear of bottoming at all times. I hoped it would reproduce the performance shown in Fig. **5b**, which indicates the gain of an synchronous oscillator versus input level as described in the references. Incidentally, Fig. **6b** shows the typical performance of an oscillator when the mean emitter current is not controlled to a level appropriate to the dynamic resistance of the tank circuit. Not only does the base emitter junction become forward biased, but the base collector junction does likewise at the negative extreme of its excursion. It thus appears as a forward biased diode connected directly across the tank circuit, providing heavy damping which reduces the effective operating Q.

The waveforms shown here resulted from a lower value of resistor (10kΩ) in the emitter of the lower *BFY90* in Fig. 5a, but are typical of the result when an oscillator circuit provides the active device with a fixed base current. High gain samples of the oscillator transistor will try to pass more collector current than is appropriate, and the negative going collector swing then mops up the excess base current, charging the associated base capacitors negatively in the process. The result is an overlong conduction angle, as shown.

Figure 7a shows the output of the synchronous oscillator circuit of Fig. 5 as a 0dBm input is slowly swept over a 500kHz range centred on 6.7MHz. At the start of the sweep, the oscillator is not in lock and so runs at its natural centre frequency, producing a bright trace there, while the applied frequency can be seen to be well down the skirt of the tuned circuit. However, as the sweep proceeds, the amplitude of the applied signal does not trace out the shape of the tank circuit's response curve, but suddenly jumps up to the same level as the oscillation, as the latter synchronises with it.

The lower level responses that are visible are due to the various out-of-lock sidebands seen in Fig. 3d: they would not have been visible if a spectrum analyser with a built-in tracking generator had been used. However, the top of the trace indicates the output at



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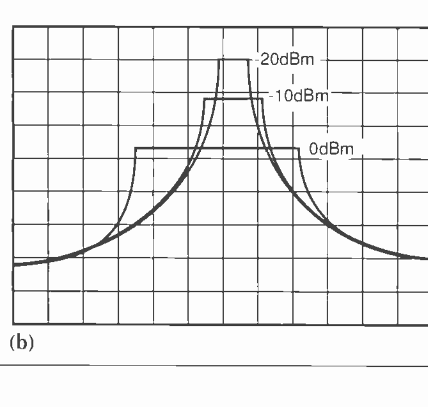

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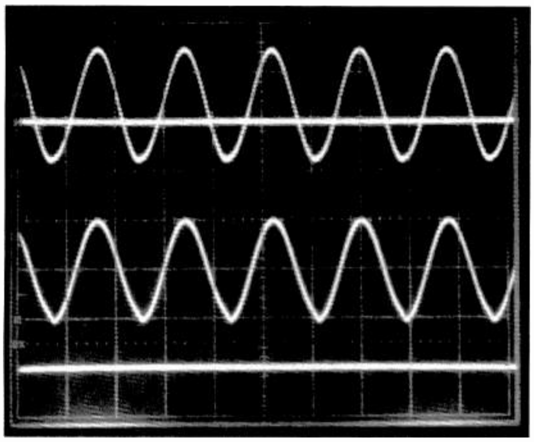
Fig. 5a. Modified synchronous oscillator circuit with robust definition of dc conditions. b. Gain and bandwidth of a synchronous oscillator as reported in the published papers.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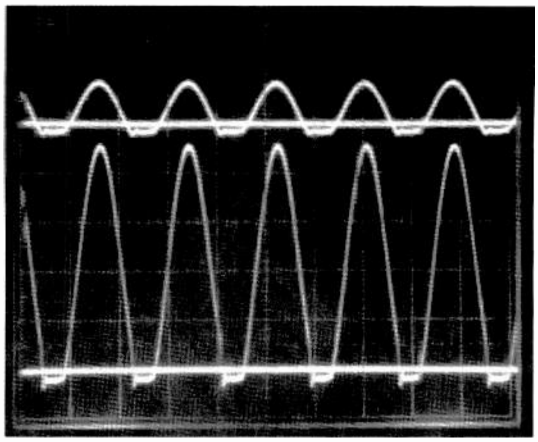
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Fig. 6a Waveforms associated with the synchronous oscillator of Fig. 5a. The emitter voltage reaches a peak of 0.8V negative with respect to the base, which is at 0V ground, as indicated by the superimposed grounded-input trace (upper trace, 1V/div) The collector voltage (lower trace) swings 10Vp/p about the +10V rail, i.e. always at least 5V clear of the superimposed ground trace (lower trace, 5V/div.) Timebase 0.1µs/div.
b. As a but with too high an emitter current. These waveforms are typical of an oscillator circuit where a fixed base current rather than a fixed emitter current is used.

the wanted frequency, whether in lock or out. Figs **7b** and **7c** show the synchronous oscillator output with the input reduced from 0dBm to –20dBm and –30dBm respectively. The result with –10dBm was intermediate between the two.

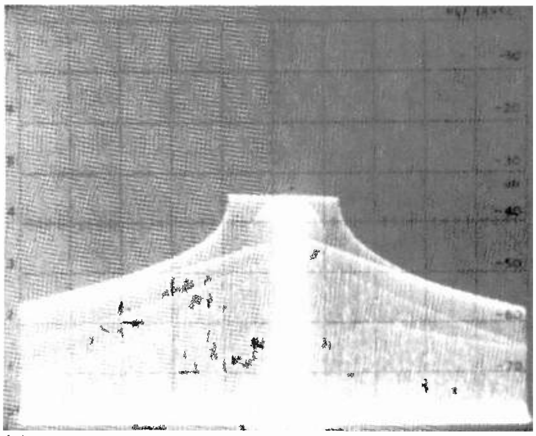
Since the amplitude of the synchronous oscillator output in lock is independent of the level of the input, it follows that the smaller the input, the larger the "gain" to the signal. Fig. **7d** shows tracings of the results in 7a-7c superimposed such that the edge of screen levels are aligned, illustrating this point and giving the same sort of result as in the published papers.

At 0dBm input, the lock range was 105kHz or 1.56% of the centre frequency, and 0.62%, 0.21% at –20dBm, –30dBm respectively. The acquisition of lock appears to be virtually instantaneous, unlike a second order PLL using a simple ex-OR phase detector, where the time to acquire lock may run to many cycles of the input.

Acquiring signals in noise

My tests to date have only covered the performance of the synchronous oscillator with a clean CW input – a very high SNR (signal to noise ratio) – whereas one of its main uses is said to be recovering a signal accompanied by noise. It seems that they are not ideal for recovering signals deeply buried in noise, Ref. 4 stating that the types experimented with and employed in various digital radio subsystems are not used below E_b/N_0= –3dB. I intend carrying out further tests in conjunction with a suitable noise generator, which will entail building a permanent version of the circuit reported in Ref. 7.

It was my hope that the synchronous oscillator would prove suitable for pulling out a signal with a large *negative* SNR that prompted the use of an LTP in Fig. 2. With suitable emitter degeneration, the LTP would modulate the input signal plus noise onto the oscillator's emitter current in a linear manner. This is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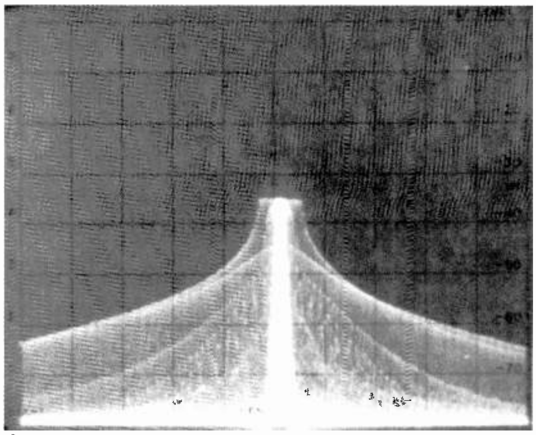
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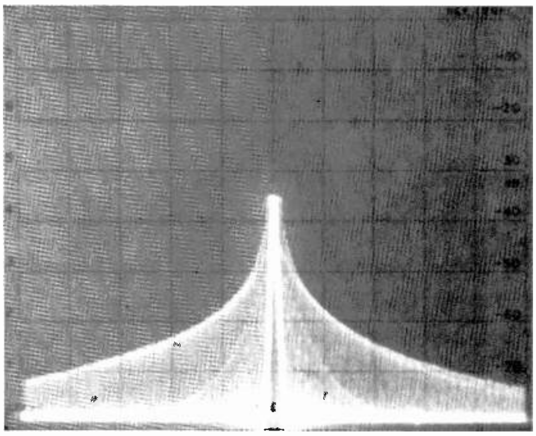
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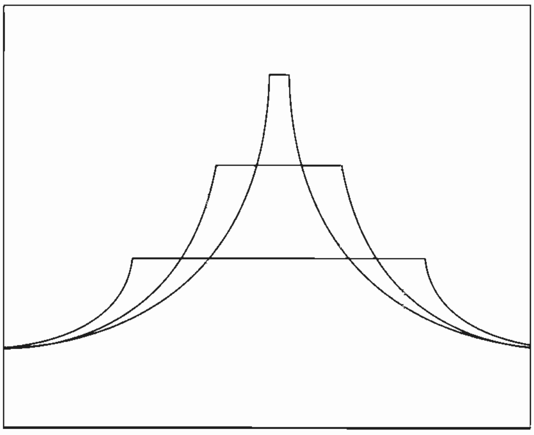
(d)

Fig. 7a. Output of the synchronous oscillator of Fig. 5a, level of input signal 0dBm. (Spectrum analyser settings: vertical, 10dB/div; horizontal, 50kHz/div, centre frequency 6.7MHz; IF bandwidth 3kHz, video filter off. Spectrum analyser driven from oscilloscope's Channel 2 output)
b. As a but –20dBm input
c. As a but –30dBm input
d. Traces a – c superimposed

important, since with a large noise voltage applied to the LTP (and no degeneration) the signal plus noise input would in effect be hard limited. Now, while hard limiting a signal well above noise improves the SNR by 3dB, hard limiting a signal buried well down in noise makes it 3dB worse. Perhaps this is the reason for the limited performance of the type of synchronous oscillator shown in Fig. 1, where there are no specific measures to linearise the transistor forming part of the signal injection network.

Ref. 6 quotes results with inputs up to 0dBm which, in a 50Ω system, corresponds to 636mV p/p. This is certainly more than the injection transistor (with no emitter degeneration or other linearising measures) can be expected to handle linearly. In fact, the transistor will start to bias itself back towards class C, so that both it and the oscillator transistor operate in a sampling mode. But with a negative SNR, this sort of limiting (by "dc restoring" the signal negative going) probably risks suppressing the wanted signal energy even more than the symmetrical hard limiting produced by an LTP.

The synchronous oscillator shares with the PLL a difficulty in giving a reliable in-lock indication with acquired signal. In a PLL used in a high SNR application, e.g. in a synthesiser, a lock detector often comes for free in the phase detector chip, along with an edge triggered "Type II" phase detector for rapid lock acquisition. However, in a PLL used for recovering a signal buried in noise, edge-triggered logic-machine type phase detectors don't work and must be replaced with an ex-OR variety, with the smaller ±90° phase range. With this type of phase detector, the pull-in time can become quite long especially at poor signal to noise ratios. (The distinction between pull-in range and lock-in range is explained in Ref. 8, which further states that the distinction between the two can become rather blurred if appreciable noise is present in the circuit.) It will be interesting to see if the synchronous oscillator still locks up instantly to a wanted signal buried in noise. ■

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Measuring pulses digitally

Frequency meters normally count the number of unknown waveform cycles occurring during a known reference time period. For higher frequencies, such cycle counting is fine. Given a 0.1s reference gating period and a 1MHz signal under test, it is possible to produce a measurement with a resolution of 1:10^5 in 0.1s. Reading a 1Hz signal to the same resolution however would need a 100,000s gating period – and a wait of more than a day for the result.

Rather than using a reference period to gate the input signal, as in a conventional frequency meter, pulse-width measurement schemes use a single cycle of the unknown input pulse to gate multiple cycles from a reference clock at a higher frequency. Width of the unknown pulse is directly proportional to the number of reference cycles counted during the gate period. Frequency is simply the inverse of this reading.

Where the unknown pulses are quite wide, or high resolution is not essential, microcontrollers with on-chip counter/timers provide a single-chip solution. Speed is limited by clock dividers within the microcontroller. Maximum clock rate for counter/timers in an *8051* controller running at 12MHz for example is only 1MHz. In addition, microcontrollers are expensive to develop and implement. Standard logic ICs provide a faster solution but significantly increase the number of components needed.

Two dedicated ICs recently launched by Numa have been designed to produce a digital output reading representing an unknown input pulse width. Called the *NT302* and *NT304*, these period-to-digital ICs provide binary counts of up to 2^{12} and 2^{16} respectively. Maximum frequency of the external reference clock is 125MHz while the maximum rate for extracting data from the first-in-first-out register is 10MHz.

Converting the width of a pulse into proportional digital data is useful not only for period measurement but also for many frequency-determining applications. The new pulse-measuring ICs covered here also open up a unique solution to fm demodulation. Martin Eccles reports.

Within the *NT302* are two 12-bit counters, gating circuits, latches for the data and control logic. As shown in the block diagram, there is also a 32 by 12-bit first-in-first-out register common to both channels and a clock divider stage with four external outputs for driving, say, a DSP chip. The two input channels accept complementary C-MOS compatible input signals. Only one other input is needed – the reference clock.

Reference clock

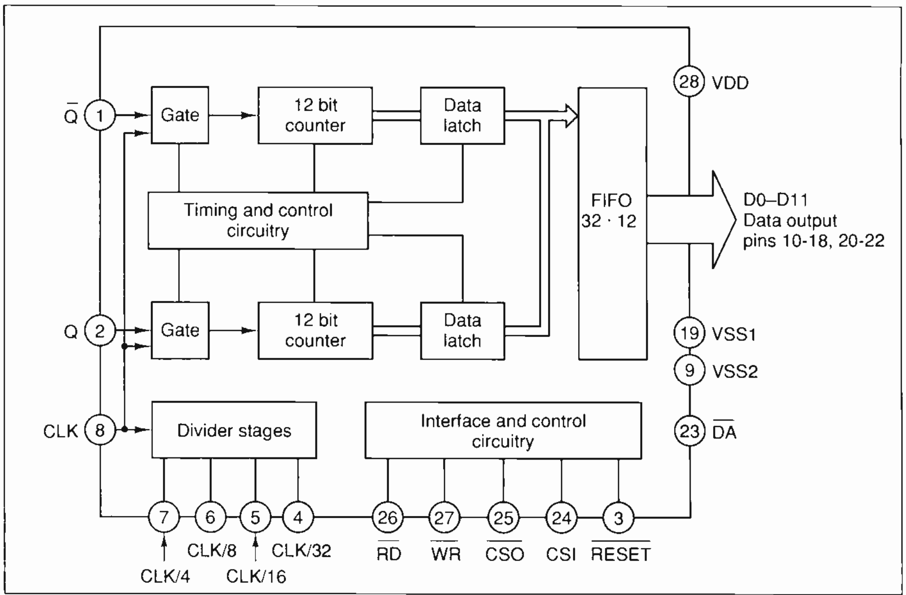
Clock frequency is simply,

$$F_{clk}=F_{in(min)}4095$$

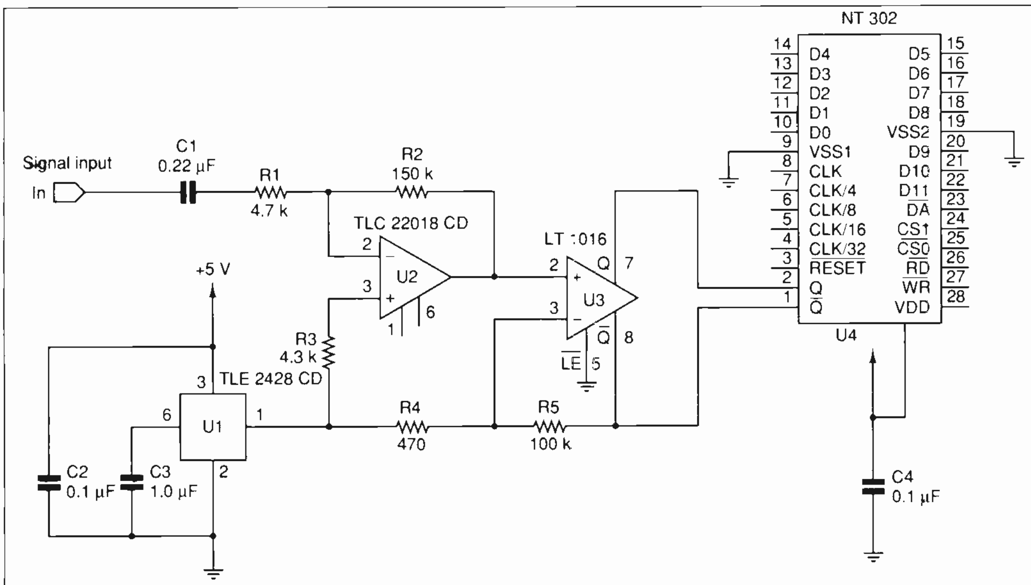
where F_{in} is the input signal.

Sampling of the first and second half cycles, namely t_{pwh} and t_{pwl}, is alternate. The 12-bit binary number representing the first half cycle is given by,

$$N_1=t_{pwh}/(1/F_{clk})$$



New period-to-digital converter IC. Frequency counters normally count zero crossings of the unknown signal during a known reference gate period. This device uses zero crossings of the unknown waveform to gate reference clock pulses to a counter. Period-to-digital conversion is generally more appropriate for pulse-width measurement at high frequencies and frequency measurement at low frequencies.



Zero-crossing input circuitry needed for correct operation of the period-to-digital converter. A 2.5V reference feeds the non-inverting input of the first op-amp while dc-blocked input feeds the other. Gain of this stage squares small input signals ready for the comparator that follows.

and for the second by,

$N_2 = t_{pwl}/(1/F_{clk})$.

For the whole period, the 12-bit binary number becomes simply,

$N_{per} = N_1 + N_2$

or

$N_{per} = (t_{pwh} + t_{pwl})/(1/F_{clk})$.

Once clock frequency is found, the minimum input frequency is established. This frequency causes a count of 4096. Should the frequency fall lower, the counters will overflow producing a false result.

In the circuit diagram is a zero-crossing detector which ensures correct operation of the *NT302*. On the left is a regulator producing a fixed 2.5V reference. This voltage may vary depending on the application. Input is capacitor coupled to inverting input of the first op-amp, whose gain ensures that the input waveform is square. Next, the comparator produces the complementary waveform necessary for the *NT302*.

On a high-to-low transition at the *302*'s Q inputs, the count is transferred from the 12-bit counter to the data latches then into the fifo. Next, the counters are reset and the system samples the next half cycle.

Latching and resetting takes approximately eight reference clock cycles and continues as long as there is an input signal. Maximum duty cycle of the input signal becomes a function of the eight clock cycle transition time and is given by,

$DC < 1 - 8(t_{pw}/T)$

where is t_{pw} is $1/F_{clk}$ and T is $1/F_{in}$. For an input frequency of 1MHz and a system clock of 100MHz, maximum duty cycle would be 92%.

Data in the fifo is shifted out on a falling edge at the READ input. If data is present at the fifo output, the data-available line, DA, goes active low. In the event of no data, DA remains high, allowing the presence of valid data to be signalled to software or hardware. Since there are two zero crossings per cycle, data can be extracted from the fifo at twice the rate of the input frequency.

The two counters are enabled alternately and there are two zero crossings per cycle. This means that 12-bit binary numbers representing each half cycle are extracted from the fifo alternately. They can be added to produce a period result, divided to define duty cycle or used as an address to access a look-up table.

Sending the data to a digital-to-analogue converter provides an analogue representation of the binary word. Relative to analogue pulse-width measurement techniques, this digital approach results in an accurate and linear analogue signal which is highly immune to DC offsets and temperature drift.

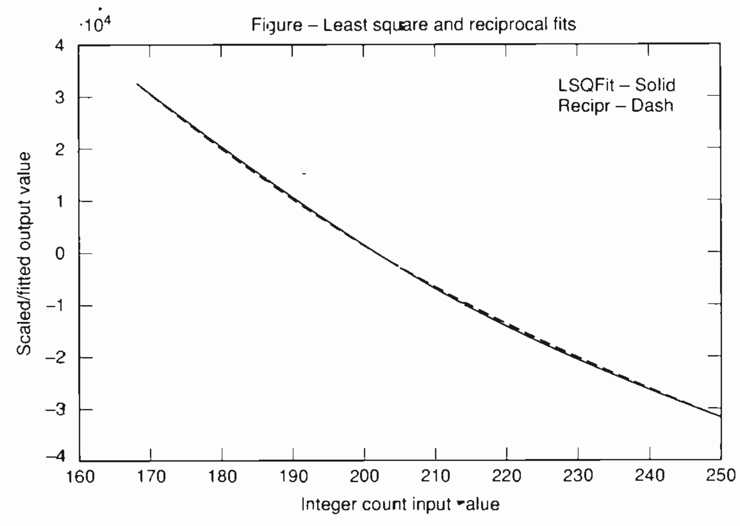
Novel fm demodulator

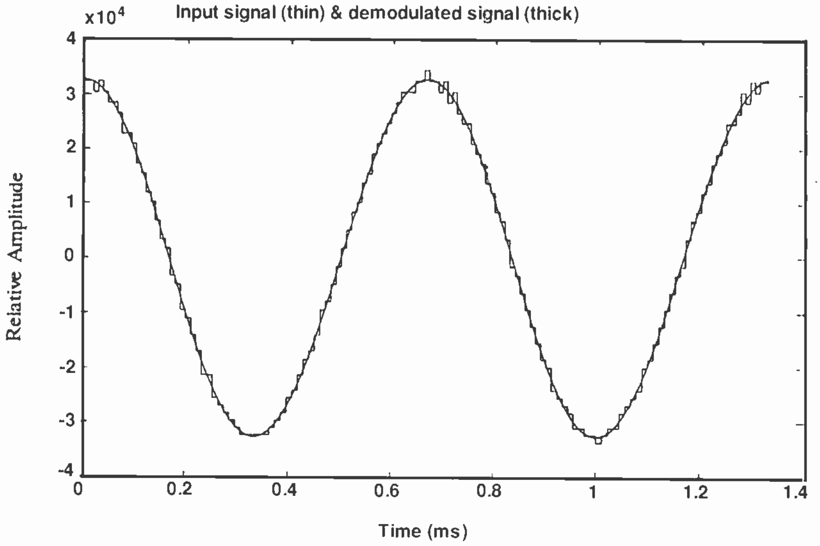
This approach to fm demodulation is said to be unique. Incoming fm analogue signals are amplified by a high-gain IF stage in which fm square waves are produced by limiting action.

FM demodulation via DSP

Reciprocal count scaling allows precise fm demodulation using non-floating-point DSP. It is difficult to see the difference between the least-squares fit and the reciprocal-fit methods. Point-for-point, the difference betwen the two methods is negligible, given a range of 65,520 data values.

Reciprocal count scaling calculates the reciprocal of the period-to-digital count values in a fixed-point microprocessor without needing mathematical division. The operation involves a look-up table performing a linearised inverse-transfer function. Should a floating-point processor be used, the reciprocal fit process is not needed.





Results obtained from a new approach to fm demodulation. Applying a high gain stage to the incoming IF information produces squarewaves that can be sampled by the period-to-digital chip with the aid of a high-speed system clock. These digitised samples then feed a digital signal processor which uses estimation techniques to produce frequency-demodulated audio data. Part of the processing involves a technique called reciprocal count scaling. In addition to being high quality, the output also has a high signal-to-noise ratio.

Using high-speed clock and counting circuits, the square waves are digitised into binary counts. These counts are sampled and sent to a digital signal processor without using any analogue-to-digital conversion.

Counts are subsequently processed in the DSP system using estimation. Noise is said to be inherently low since the technique has a noise limiting as opposed to the noise additive effect common in traditional analogue fm demodulators.

Under the proposed demodulation technique, incoming IF is squared then sampled directly to obtain its zero-crossing information. Digitally encoded, this information is fed to a DSP chip where the modulating signal is extracted using a reciprocal count-scaling software routine.

Extracted signal can be further digitally filtered before being passed to a digital-to-analogue converter. Adopting reciprocal count scaling for demodulation is said to guarantee the most accurate results attainable from a system of this type in terms of s-to-n ratio, harmonic distortion and scaling accuracy.

Digitised count information is produced so that the lowest number of reference clock counts for a given sample corresponds to the highest modulating signal amplitude. Conversely, the highest clock count represents the lowest modulation amplitude.

Numeric processing hardware in the demodulator system receives a binary representation of fm data directly from the zero-crossing circuitry in the IF strip. Estimations are made of successive digital samples.

An application article entitled *DSP peripheral IC does period-to-digital conversion* is available from Clere Electronics at Kingsclere in Newbury, Tel. 0635 298574, fax 297717. It includes full design equations for the fm decoding technique.

Algorithms for implementing the new fm decoding method are also avaliable to design engineers. These are based on the Analog Devices 2100 DSP family. ■

NEW PRODUCTS CLASSIFIED

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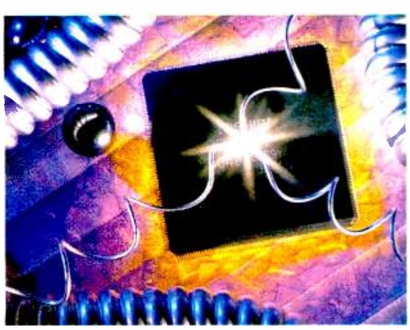
Asics

A-to-D and D-to-A converters
5MHz, 12-bit sampling A-to-D. Datel's *ADS-118* sampling analogue-to-digital converter, a 12-bit device working at 5MHz from the input leading edge only rather than a full pulse, is guaranteed to generate no missing codes over the –55°C to 125°C temperature range. Signal-to-noise ratio is 70dB and THD –69dB. Requirements are ±5V at a maximum of 1.5W. Packaging is smaller than usual in a 24-pin double-wide ceramic dip.
Datel (UK) Ltd. Tel., 0256 880444; fax, 0256 880706.

Linear integrated circuits

RF/video switch. Siliconix's *DG64X series* video switches exhibit low on resistance, the *DG643*, for example, presenting only 8Ω with an on-state capacitance of 10pF. Bandwidth to –3dB is 500MHz and current-handling ability 75mA. Packaging is SO-16 or 16-pin dip.
Siliconix/Temic Marketing. Tel., 0344 485757; fax, 0344 427371.

850MHz buffer. An 850MHz buffer amplifier by Harris, the *HFA1114*, is intended for video and instrumentation work and eases the problem of holding the full bandwidth over long cables by allowing access to the summing junction at the internal amplifier's inverting input, so that compensation networks are more easily and effectively applied. The amplifier also allows gain programming to +2, +1 and –1. Salient characteristics are 1500V/µs slewing at a gain of +1, 2400V/µs at –1 and 1900V/µs at +2, 11ns settling time to 0.1%, gain flatness to 30MHz and 0.07dB to 100MHz. Output current is 60mA.
Harris Semiconductor UK. Tel., 0276 686886; fax, 0276 682323.



50Ω pulse amplifier. Bandwidth of the Mini-Circuits *ZPUL-21* 50Ω, 24V/350mA pulse amplifier is 2.5kHz-700MHz and it is usable up to 1GHz with a flatness of ±0.6dB. Rise and fall time is 1.1ns and delay time 1.5ns.
Mini-Circuits Europe. Tel., 0252 835094; fax, 0252 837010.

GaAs switch. *GSWA-4-30DR* is a gallium arsenide switch from Mini-Circuits with a bandwidth of 0-3GHz. It switches at 25ns and isolation is 30-40dB to 3GHz, video leakage to 50Ω RF ports 30mVpk-pk typical in 500MHz bandwidth.
Mini-Circuits Europe. Tel., 0252 835094; fax, 0252 837010.

PWM controller. *UC3848* by Unitrode is the first family of current-mode PWM controllers designed for single switch forward converters, allowing primary-side control of secondary-side current in isolated switching power supplies. It offers low peak:average current ratio, low ripple current and a start-up current of 500µA. It copes with continuous or discontinuous inductor current, allowing a smaller inductor. During switch-on the device monitors switch current and synthesises inductor current down-slope to simulate inductor current during switch-off.
Unitrode (UK) Ltd. Tel., 081 318 1431; fax, 081 318 2549.

Logic building blocks

Automotive relay driver. Zetex's avalanche-rated *ZVN4206AV* n-channel mosfet is designed for automotive use and removes the need for driver-relay interface and protection. It supports a drain-source voltage of 60V and exhibits an on resistance of 1Ω, operating from a 5V gate drive. Although the device will drive 600mA coils, it needs no catch diodes, all energy being dissipated through avalanche breakdown of the intrinsic body diode, which also handles reverse-battery accidents and 50j, 65V load dumps.
Zetex plc. Tel., 061-627 5105; fax, 061-627 5467.

Low-power 0.5µm embedded array

3V cmos embedded arrays in 0.5µm technology from Toshiba now offer the option of 5V working. *TC183E* has the same core as the 3V *TC180E* series, having gate propagation delay of 0.25ns and is claimed to outperform 0.8µm, 5V asics. Each i/o pin is individually programmable from a library of 3V or 5V interface cells. The devices yield up to 340,000 usable gates.
Toshiba Electronics (UK) Ltd. Tel., 0276 694600; fax, 0276 691583.

Largest-memory microcontroller

Hitachi's H8/3048 is a variant of the company's H8/300H series of 16-bit microcontrollers, which has, says Hitachi, the largest on-chip memory in the world at 128K of rom and 4K of ram.

It offers a high-performance CPU that gives up to 1.9Mips in a Dhrystone benchmark at 16MHz and a large linear memory addressing space of 16Mbyte. Versions operating at 3.3V work at 13MHz to 1.54Mips.

The device is particularly suited to use in embedded systems where large look-up or data tables must be accessed, the large memory allowing greater flexibility while maintaining compatibility with other members of the family.

For use in GSM handsets, there is a serial port to interface to the GSM subscriber interface module and several power-down modes of operation. On-chip peripherals include an integrated timer unit providing timing operations such as PWM and multiphase outputs to drive three-phase motors. There are also a full DMA controller, a 10-bit 8-channel A-to-D converter, an 8-bit 2-channel D-to-A converter, a 2-channel serial comms interface and a watchdog timer. A timing pattern controller connecting the DMA, timer unit and i/o ports to generate timed patterns to the output pins to drive stepper motors without involving the CPU.

Hitachi Europe Ltd. Tel., 0628 585000; fax, 0628 585200.

18-bit bidirectional buffer. Combining the functions of two transparent octal latches and an 18-bit wide, 4-deep fifo in one 56-pin package, IDT's *FCT162701T* has a propagation delay of 5.5ns and solves the bandwidth mismatching between CPU and memory or CPU and i/o in *R4400/R4600*, *Pentium* and *PowerPC* applications. It also saves around 80% of board area and reduces the number of ICs by eight to one. The *162701T* allows zero wait-state working up to 75MHz and more than doubles system performance. Ground bounce is less than 600mV.
Integrated Device Technology. Tel., 0372 363734; fax, 0372 378851.

Clock distribution. Output skew of the TI *CDC330* clock distribution circuit is 0.4-0.5ns maximum and the process skew 1ns maximum. Maximum clock frequency is 67MHz for two 1-to-3 paths and 100MHz for the divide-by-two path, both for a duty cycle of 40-60%, all these figures for a maximum output current of ±32mA. Each set of outputs per path is individually switchable into tri-state.
Texas Instruments. Tel., 0234 270111; fax, 0234 223459.

Memory chips

Low-voltage memory. Made by Microchip, the *24AAxx* family of 1.8V serial eeproms in 1K, 2K, 4K, 8K and 16K versions draw only 50µA read current and 3µA on standby. They operate over the full voltage range of two AA batteries, which have an end-of-life terminal voltage of 0.9V. Features include Schmitt trigger inputs, output slope control to eliminate ground bounce and user-selectable write protection. They also have a page-write buffer holding up to 16 bytes.
Polar Electronics. Tel., 0525 377093; fax, 0525 378367.

Microprocessors and controllers

Printer chip. *68322* from Motorola is the first in a series of microprocessors intended for use in low-cost printers. It is a single-chip, dual-processor device optimised for eight page-per-minute, 600 DPI PCL/Postscript laser printers, but is also suitable for multifunction peripherals, inkjets, and other general embedded applications. The *68322* offers graphics acceleration on the one chip and is a replacement for current 68000 designs with companion asics. The device supports banding, in which a page is stored in compressed format in the printer, being output in bands as the printer runs, so reducing printer memory needs.
Motorola Inc. Tel., 0908 614614; fax, 0908 618650.

Please quote "Electronics World + Wireless World" when seeking further information

Low-power microcontrollers. Low-power, 8-bit microcontrollers in Fujitsu's *MB89120* and *MB89130* series use the same CPU as other members of the F^2MC-8L line, taking a current of only 1.8mA at 3V and 4MHz, although the devices operate from 2.2V to 6V. *MB89120* offers on-chip 4Kbyte rom, 128byte ram, a watchdog timer, synchronous serial interface, two clocks for 4.2MHz and 32MHz, two cascadable 8-bit timer/counters and 36 i/o circuits. The MB89130 has all that plus a 4-channel A-to-D converter with a 42.9µs conversion time.
Fujitsu Microelectronics Ltd. Tel., 0628 76100; fax, 0628 781484.

Multimedia processor. Claimed by TI to be the highest-performing DSP available, the *Multimedia Video Processor* (MVP) delivers over two billion operations/s based on risc-like instructions, which is the kind of performance needed for microprocessors in the video world. The device is based on TI's *TMS320* DSP family and contains four DSPs and a risc processor on the same chip. It is fully programmable and supports MPEG, JPEG, H.261 and G.728.
Texas Instruments. Tel., 0234 270111; fax, 0234 223459.

Mixed-signal Ics

RF prescaler. RF portable telephone prescaler *µPB1502GR1* by NEC operates from 0.5GHz to 2GHz, drawing only 3.2mA from 3V. Division ratios are 64, 65, 128 or 129:1.
NEC Electronics (UK) Ltd. Tel., 0908 691133; fax, 0908 670290.

Optical devices

Optical-fibre switches. AMP's Kaptron ranges of optical-fibre switches are for general and FDDI installations and provide losses of less than 0.5dB. They come in single-pole, double-throw and DPDT forms, the PC-mounting general-use types being 5V operated. The FDDI version conforms to FDDI and IEEE 802.5, being a fully reversing bypass switch that allows faulty modes in a Token Ring network to be isolated. In bypass mode, a self-test path enables the transmitter to be tested against its own receiver while maintaining less than 0.5dB loss in the external network.
Auriga (Europe) plc. Tel., 0908 274200; fax, 0908 378998.

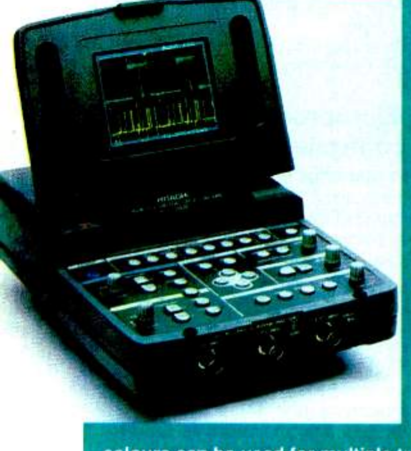
1300nm laser diode. For use as a light source in short and medium-range optical-fibre communications, Hitachi's *HL1326CN* is a 1300m InGaAsP Fabry-Perot laser diode in a coaxial package with a single attached fibre. Threshold current at 25°C is 10mA, so that there is no need for additional cooling. Output power is 3mW pulse, 3mW continuous, reverse voltage 2V, rise and fall time 500ps maximum and monitor diode dark current a maximum of 350nA.
Hitachi Europe Ltd. Tel., 0628 585000; fax, 0628 585200.

High-brightness leds. New from H-P are the *HLMP-V500*, *HLMP-VL00* and *HLMP-V100* leds for use in outdoor displays, colours being green, amber and red respectively. Luminous intensities at 20mA are 270mcd, 460mcd and 1000mcd, all the devices radiating a 60° horizontal angle and 30° vertically.
Hewlett-Packard Ltd. Tel., 0344 362277; fax, 0344 362269.

Notebook oscilloscope

Clearly inspired by the notebook computer, Hitachi's Notebook digital oscilloscope has a 4in thin-film transistor colour LCD, hinged to fold as in the instrument's namesake. It is both battery and mains powered, the battery charging automatically to last two hours of continuous use.

Performance is not compromised; bandwidth of the two channels is 50MHz and sampling speed 30Msample/s simultaneously, with an acquisition memory of 2Kword/channel storing up to 100 waveforms each of 2Kword with backup. Vertically, resolution is 8bit and sensitivity 1mV/division to 5V/division in 12 ranges to an accuracy within ±3%.



In the horizontal direction, sweep time is 5ns/div-2µs/div, ±1%; real-time sampling 5µs/div -1s/div, ±0.04%; roll mode 0.2s/div-50s/div, ±0.25%. There is 10div pre-triggering and 400div post trigger, sources being Ch1, Ch2, diff. and ext; modes auto, norm, Tv-V and Tv-H. Coupling is DC, AC, HF reject and LF reject. Since the display is a colour type, different colours can be used for multiple traces and alphanumerics, the display providing refresh, infinite persistence, dot-to-dot joint/interpolation and X-Y display. Possible processing functions are exponential averaging, weighting coefficient 2-256 and plus, minus and inversion. All pulse parameters can be measured.
Hitachi Denshi. Tel.,081 202 4311; fax, 081 202 2451.

Oscillators

Miniature oven oscillators. *OXCO* miniature ovened crystal oscillators in 14-pin dil packages by Stanler use the company's own crystals and cover the 10kHz-24MHz frequency range. Power used is 25mA from 12V at 25°C, 70mA being needed for the first 10s after switch-on. The units are in specification after 30s, ageing at the rate of ±0.7pp/year and ±5ppm in 10 years. Stability is ±0.2ppm in temperatures from 0-60°C and ±0.3ppm from –20°C to 70°C.
Stanler Components Ltd. Tel., 0376 340902; fax, 0376 322510.

300-1000MHz VCOs. Vectron has a range of voltage-controlled oscillators for those applications needing stability with a reasonable deviation. *VC-374* oscillators are available for any centre frequency between 300MHz and 1GHz, with deviation of ±10% as standard and up to ±20% as an option, linearity being ±10% or an optional ±3%. Temperature stability is either ±1% over 0-50°C or ±5% over –55°C to 125°C. Power needed is 50mA from 15V.
Lyons Instruments Ltd. Tel., 0992 768888; fax, 0992 788000.

Power semiconductors

S-M power mosfets. *Rohm 2SK* power mosfets feature 0.4-1.4Ω on resistance, fast switching and drain currents from 2A to 10A at 30-450V. For higher voltages, the *Shindengen* n-channel enhancement devices offer 900V at 1A, with 0.65Ω on resistance in the 200V, 5A device to 15Ω in the 900V, 1A type. Input capacitance for the *2SK1194* is 45pF to give switch-on/off times of 30ns/50ns.
Flint Distribution. Tel., 0530 510333; fax, 0530 510275.

Infrared power switching. *TLP250* is a photo-IC coupler from Toshiba, designed to switch the gate driving circuits of IGBTs and power mosfets. It integrates a GaAlAs led, a photodetector, a high-gain amplifier and output drive on a single chip, providing switching times of 0.15µs typical. Voltage rating is 10-35V at 11mA maximum; minimum isolation voltage is 2500V RMS for one minute at 25°C.
Toshiba Electronics (UK) Ltd. Tel., 0276 694600; fax, 0276 691583.

Fast IGBTs. International Rectifier's new *Ultrafast* 500V insulated-gate bipolar transistors operate at up to 100kHz at current ratings of 14-59A. They have higher current densities than equivalent transistors so that smaller chips are needed. These devices are also available paired with *HEXFRED* fast recovery diodes in five new Co-Packs for totem-pole applications.
International Rectifier. Tel., 0883 713215; fax, 0883 714234.

Switched-mode power supply
Better known for its transformers, Gardners Ltd has introduced the first in a planned new range of switched-mode supplies for the medium-volume European market, which will offer outputs from 15W to 100W in open-frame, lead-to-lead, plug-in and encased forms. *LCS40* is made to industry-standard pin-outs and fixing centres, five modules with single, dual and triple outputs of common voltages accepting 85-265V AC and 120-370V DC. Operating frequency is 47Hz-440Hz and the outputs may be adjusted over a ±5% range. Line regulation is ±0.3%. The range has been designed with the aim of providing any combination of input and output, within reason, at a fair cost. The company says its LCS40 is made at a price to compete with offerings from the Far East, and points out that extended lines of communication can cause problems that are avoided by manufacturing in the UK.
Gardners Ltd. Tel., 0202 482284; fax, 0202 470805.

PASSIVE

Passive components

Miniature electrolytics. A range of space-saving aluminium electrolytics, the *Rubycon REV* series, offers values in the range 0.1µF-220µF at 4-50V DC, the smallest type of four measuring 3.3mm diameter by 5.5mm high. Tolerance is ±20% and leakage 3µ. They are thermal-shock tested for conformity with reflow soldering and operate from –40°C to 85°C.
Surtech Interconnection Ltd. Tel., 0256 51221; fax, 0256 471180.

Switched potentiometers. 16mm carbon-track potentiometers from *Omeg* are fitted with integral push-push switches and also offer the option of indents to provide tactile feedback. Standard indent patterns are 1, 11, 21, 31 or 41 equally spaced indents, but Omeg can provide non-standard numbers and positions. Resistance values are 0.25W 1kΩ-1MΩ linear and 0.12W 4.7kΩ-470kΩ non-linear. Switch travel is 3mm.
Omeg Ltd. Tel., 0342 410420; fax, 0342 316253.

60MHz ceramic resonators. Murata has extended the frequency limit of ceramic resonators – a lower-cost alternative to crystals – from 32MHz to 60MHz. They are now available in the standard frequencies of 33.86MHz, 40MHz and 50MHz for use in cordless telephones, CD and digital cellular equipment and are also available to specific requirements up to 60MHz to special order. Initial tolerance on the *CSA* devices is ±0.5%, or down to ±0.2% to order, temperature drift being less than ±0.3% between –20°C and 80°C. Resonant impedance is 40Ω or less.
Murata Electronics (UK) Ltd. Tel., 0252 811666; fax, 0252 811777.

Crystals

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Murata Electronics (UK) Ltd. Tel., 0252 811666; fax, 0252 811777.

SM crystals. *HC49/4HSMX* is a surface-mounted version of the established *HC49/4H* crystal package, measuring 4.9mm by 13.4mm and 4.3mm high. Frequencies available lie in the 3.2MHz-50MHz range, with frequency tolerance and stability in the operating temperature range of ±50ppm and ±100ppm, although tighter specifications can be supplied for many frequencies. Load capacitance is 5-7pF, shunt capacitance 2-9pF and drive level 100μW maximum.
IQD Ltd. Tel., 0460 77155; fax, 0460 72578.

Small crystals. Seiko Epson's *CA-303H* mini-cylinder quartz crystal now operates over the temperature range –40°C to 85°C and has a range of fundamentals from 12MHz to 24MHz. Temperature stability is ±3ppm to ±15ppm, depending on temperature, and ages at ±2ppm/year. The package measures 8.9mm long and 3.1mm in diameter.
Advanced Crystal Technology. Tel., 0635 528520; fax, 0635 528443.

Displays

Bright multi-led indicators. Indicators using bright led arrays for sunlight viewing combine the brightness of tungsten filaments with the long life of semiconductors. The *Dialight 556* series provides the option of red, yellow or green with intensities up to 6120cd/m^2 in flat lens versions and 340cd/m^2 in 180°-viewing domed lens types. 12V, 24V and 120V versions are available, all fitting into a 1in cutout.
Dialight. Tel., 0638 665161; fax, 0638 660718.

Filters

Saw filters. New, miniature surface acoustic-wave filters by Murata are designed for use in mobile communications from GSM to PCN and DECT. Height off the board is 2mm and the devices come in a range of packages. An example is the *SAFC71.0MA50T*, which is a 71MHz version with 4.5dB insertion loss in a package measuring 6.5 by 13.3mm. Another is a 244MHz filter for PCN/DSC1800 systems, providing out-of-band attenuation of over 50dB with an insertion loss of 3.5dB.
Murata Electronics (UK) Ltd. Tel., 0252 811666; fax, 0252 811777.

400MHz logic analyser
At a maximum asynchronous sampling rate of 400MHz, TTI's *TA4000* logic analyser captures events down to 5ns in its full 8Kword memory. Timing resolution is 2.5ns. In synchronous acquisition, a 50MHz sampling rate enables handling of the fastest cisc processors. Compact, high-impedance probes minimise loading and channel-to-channel skew.
Thurlby Thandar Instruments Ltd. Tel., 0480 412451; fax, 0480 450409

Instrumentation

EMC measurement. A 16A line-impedance stabilisation network, the *LISN1600* by Thurlby Thandar, enables accurate measurement of EMC emissions at the supply input of any electrical equipment operating from a single-phase supply. It meets the CISPR 16 requirements for Band A in the 10kHz-150kHz range and Band B from 150kHz to 30MHz. A switchable 150kHz filter limits LF signals for Band B measurement, thereby reducing the required dynamic range. The measuring circuit is connected to either supply line or can be disconnected for checking the noise floor.
Thurlby Thandar Instruments Ltd. Tel., 0480 412451; fax, 0480 450409.

Low-cost oscilloscopes New instruments from Goldstar are the *OS-9020A, OS-9040D, OS-904RD and OS-3020*, the lowest price being for the 20MHz dual-trace 9020A at £229.75 plus vat. This, the 9040D 40MHz type at £340 and the 904RD at £425 are all analogue instruments, the latter having digital screen readout. At the top of the range is the 3020 20MHz digital storage oscilloscope at £638.
MPS Electronics. Tel., 0702 554171; fax, 0702 553935.

Clamp-on multimeters. New from ITT is a range of clamp-on multimeters which measure voltage as well as current and will compute power in PWM and other non-sinusoidal waveforms. *MX210* and *MX215* measure current up to 200A RMS, other functions being fundamental frequency of current and voltage, total RMS voltage and current, RMS voltage of fundamental voltage and current, total apparent power and effective power at the fundamental. Filters exclude switching frequencies for measurement up to 100Hz and a 1V FS analogue output proportional to the current can be switched from instantaneous to RMS. Jaw size is 23mm.
ITT Instruments. Tel., 0753 511799; fax, 0753 694983.

1GHz digital real-time oscilloscope. Claimed by Tektronix to be the world's first 1GHz digital real-time oscilloscope, the *TDS684A* is part of a new range of equipment for the development of advanced digital systems
It provides four channels of 1GHz, single-shot data capture, sampling at 5Gsample/s simultaneously on each channel, so that non-repetitive events are captured at full bandwidth. There are 25 automatic measurements and better than 50ps time accuracy, plus all the features associated with the earlier *TDS644A*. New triggering capabilities, slew-rate triggering and Set-up and Hold trigger address common timing problems and apparent random failures in high-speed circuitry.

Literature

AFDEC members guide. The Association of Franchised Distributors of Electronic Components has published its 1994 *Members and Products Guide* for 1994, in which nearly 60 UK distributors have their products listed, with telephone and fax numbers. Any value-added services such as assembly work and kitting are mentioned. The guide is free.
AFDEC. Tel., 0763 27109; fax, 0763 273 255.

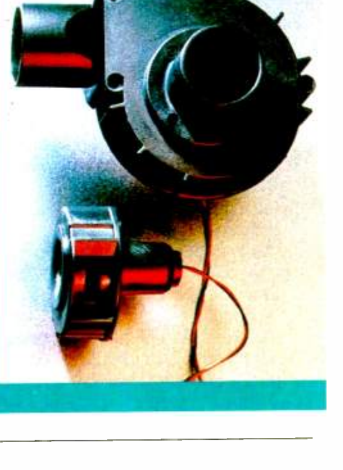
Valve and photodiode Spice models. Several application notes in the February *Intusoft Newsletter* are concerned with Spice modelling of 12AU7A double triodes, EL9000 pentodes, BPW34 semiconductor photodiodes and photodiodes in general. Also described is the interactive IsSPICE4 program for Windows 3.1, Windows NT and the Macintosh. A disk containing all circuit diagrams and Spice models described is on offer.
Technology Sources Ltd. Tel., 0638 561460; fax, 0638 561721.

Harris power. Full specifications of *Harris*'s power devices are given in a new Databook, which describes ultrafast and hyperfast recovery diodes, the *MCT* class of mos controlled thyristors and insulated-gate bipolar transistors. Application notes included describe the use of IGBTs, MCTs and fast diodes in motor control and power supplies. Among the diodes described are some operating at up to 1200V and 150A.
Harris Semiconductor UK. Tel., 0276 686886; fax, 0276 682323.

3.3V SM power supplies. Offering single, 3.3V outputs, BICC-Vero's *PK60* gives 15A and the *PK240* 45A, both at 3.3V, the PK240 having a front-end filter for use in harsh conditions. Features of both include 110V or 230V wide-tolerance AC input, long-term overvoltage and short-circuit protection, <40mV ripple, regulation <0.2% and remote on-off switching. PK60 is in a 3U by 8HP cassette and the PK240 in either a 3U by 24HP or 6U by 14HP type, both being compatible with DIN41494 subracks.
BICC-Vero Electronics Ltd. Tel., 0489 780078; fax, 0703 264159.

Inductorless DC-to-DC converter. *LTC1144* is a switched-capacitor voltage converter that accepts inputs from 2V to 18V and generates –2V to –18V. It is primarily intended to provide -15V from a positive input in automotive, applications or in battery systems with chargers and in data acquisition, and is effectively a higher-voltage upgrade for *LTC1044* and *ICL7660* devices. Only two external, non-critical capacitors are needed for charge-pump and charge-reservoir

Miniature blower
Micronel's U97 blower is a radial type providing pressures up to 750Pa and airflow of 565l/min at noise levels down to 57dB(A). Operating voltages of 5V, 12V or 24V are available in the two versions, the larger one measuring 98mm in diameter and weighing 195g. The blower's body material is polyphenylenoxide.
Radiatron Components Ltd. Tel., 081 891 1221; fax, 081 891 6839.



functions. The device runs at 10kHz from its internal oscillator, slowing down with an external C, and an ultrasonic frequency can be obtained by using the boost pin. Typical power efficiency is 93% and quiescent current 8μA.
Linear Technology (UK) Ltd. Tel., 0276 677676; fax, 0276 64851.

5W DC-to-DC converters. *Newport Components'* range of IS9001-approved converters provides output powers of 250mW-5W and are among the smallest available, measuring 19.5 by 7.5 by 10mm, some of the various package styles being surface-mounted and suitable for reflow soldering. There are 450 different models from 3V-48V in single and dual versions. Isolation is 1kV.
XP plc. Tel., 0734 845515; fax, 0734 843413.

Switches and relays

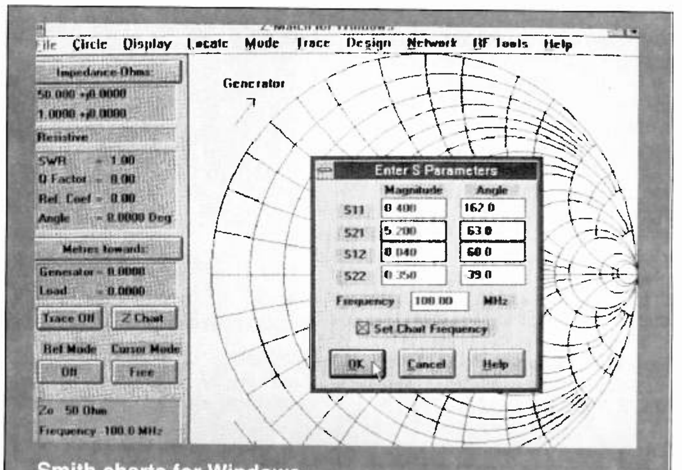
Stackable pushbutton switch. From EAO-Highland is the *Grayhill 32* series of miniature pushbutton switches, which are stackable to allow custom switch assemblies on 0.2in centres. The switches have an internally sealed plunger and are compatible with flow-soldering and cleaning operations. Status/reset or press-to-test functions are provided, with led indication. Single-pole/single-throw normally open and normally closed forms are in the same package.
EAO-Highland Electronics Ltd. Tel., 0444 236000; fax, 0444 236641.

SIP reed relays. Miniature 4-pin single-in-line reed relays for high-density mounting are announced by Coto Wabash. The *9090 Series Mini-SIP* relays measure 3.81mm by 15.24mm by 6.6mm high and are in single-pole normally open form, rated for over 10^9 operations at low loads. They have 5V or 12V coils and handle up to 10W. Options include a diode and/or magnetic shielding.
Coto Europe. Tel., (Netherlands) 01031 45 320838; fax, 01031 45 320838.

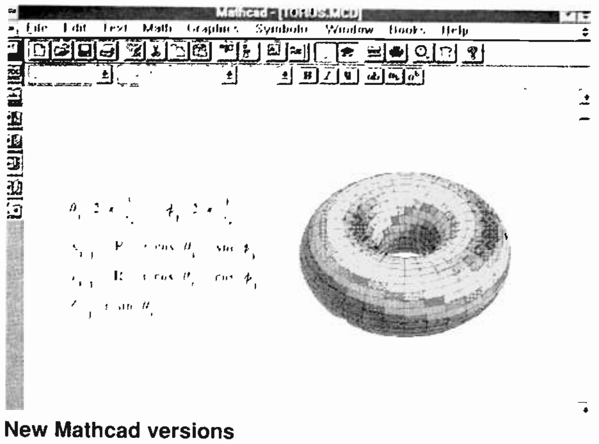
Transducers and sensors

Gear-tooth sensor. Allegro's type 3059 AC-coupled Hall-effect sensor is for the non-zero-speed sensing of ferrous targets and incorporates a voltage regulator, two quadratic sensing elements, temp. comp. circuitry, gain amplifier, filtering, a Schmitt trigger and an open-collector output driver switching up to 20mA. The differential band-pass filter virtually eliminates the effects of magnetic and system offsets and confers a degree of immunity to RF interference. Two versions cover the –20°C to 85°C and –40°C to 125°C ranges.
Allegro MicroSystems Inc. Tel., 0932 253355; fax, 0932 246622.

Pressure transducers. Subminiature versions of Kulite's *LQ-080/125* thin-line pressure sensors, the *LQ-062* series, measure 1.6 by 4.7mm by 0.4mm high, with a view to mounting them on turbine blades and other aerodynamic profiles. The sensors are available in pressure ranges between 25psi full scale and 200psi. Excitation is 5V or 10V and nominal pressure output is 100mV.
Kulite Sensors Ltd. Tel., 0256 461646; fax, 0256 47951



Smith charts for Windows
Z-MATCH for Windows by Number One Systems uses Smith chart methods to provide a set of rf circuit and system design tools in one package. It works with ordinary units, needing no normalisation, and can be switched between impedance and admittance charts. There is direct readout from the chart. Circles of constant SWR, resistance, conductance, reactance and susceptance are drawn automatically and the cursor locked to them. A receiver and system design tool calculates overall gain, noise figure and third-order input intercept point for a number of interconnected stages of filtering, amplification, mixing and attenuation. Included is a library of over 600 sets of s-parameter data for Motorola devices.
Number One Systems Ltd. Tel., 0480 461778; fax, 0480 494042.



New Mathcad versions
Mathsoft has announced new versions of its Mathcad calculation software, *Mathcad 5.0* and *Mathcad PLUS 5.0*. *Mathcad 5.0* is now easier to use and is provided with extra functions such as trace and zoom for graphics, print preview and a technical spell checker. Mathcad PLUS has the same features, but also has an entirely new set of functions such as differential equation solvers and advanced matrix algebra for more advanced users.
New features in Mathcad 5.0 include easier equation and text editing with pull-down menus and a function list, an on-line tutorial, new numeric functions including roots of polynomials and 2-D curve fitting and DDE support. For Mathcad PLUS 5.0, new algebraic functions include determination of matrix decompositions, finding eigenvalues and a complete matrix analysis system. Users may write their own routines in C or C++ or use Function Packs from Mathsoft.
Mathsoft Europe. Tel., 0344 23491; fax, 0344 873461.

COMPUTER

Computer board-level products

Peripheral interface. *PC36AT* from Amplicon Liveline is a programmable peripheral interface board using the 8255 IC to give 24 TTL-compatible digital i/o lines by way of a single 37-way type D connector. The lines are at three 8-bit ports, two of which are programmable to be all inputs or all outputs, the third being four inputs and four outputs, if required. A lower-power version for laptops is available, both types coming with demo software in Turbo Pascal and Basic.
Amplicon Liveline Ltd. Tel., 0800 525 335 (free); fax, 0273 570215.

Scorpion K4 i/o. MicroRobotics's *Scorpion K4* embedded microcontroller now has a new application module, the *K4ic*, which provides a set of analogue and digital input/output facilities to form ready-to-use hardware for much typical machine control and automation work. K4ic is in PCB form or in a metal case to CE EMC standards. Digital i/o consists of two uncommitted relay outputs handling 3A at 240V, four 2A, 24V outputs and four universal inputs, while the analogue facilities include four 8-bit inputs and an 8-bit output with selectable voltage ranges. There are also two channels of serial i/o and two g-p expansion interfaces.
MicroRobotics Ltd. Tel., 0223 323100; fax, 0223 462242.

Stand-alone audio DSP. From the Finnish company Sample Rate Systems Oy, the *M4-0202-A* is an audio digital signal-processing board operating independently of any host computer, needing only an unregulated supply and a case. It is based on the Motorola *56004* signal processing chip, has stereo analogue input and an 18 or 20-bit analogue-to-digital converter and an 18-bit output digital-to-analogue converter, which is followed by a programmable gain control. The module accommodates up to 32Kbyte of program memory and 4Mbyte of data memory, in addition to external expansion facilities. Unique to this board is the boot rom, which confers its independence from a host. Output dynamic range is over 95dB
Sample Rate Systems Oy. Tel., 010 358 31 3165 045; fax, 010 358 31 3165 046.

PCMCIA adaptors. *CARDport* is a range of PCMCIA adaptors for users of laptops, notebooks and desktop computers not provided with a PCMCIA socket. Three models are available: *CARDport eX*, which is a portable card adaptor with its own rechargeable battery, connecting to the parallel printer port of any PC and accepting memory and disk drive cards; *CARDport isa* goes in the ISA expansion bus, an additional module fitting in a disk drive bay, to provide two sockets for any PCMCIA cards; and *CARDport ide* is a low-cost desktop PC adaptor for micro hard disk drives. All are supplied with software.
Chase Advanced Technologies. Tel., 0274 841316; fax, 0274 841302.

Data communications

HDSL transceiver. Brooktree announces the *Bt8952* single-chip high-bit-rate digital subscriber line transceiver, claimed to be the first all-digital embodiment of the HDSL technique for upgrading standard twisted-pair telephone lines to take higher bit rates for fast digital services. Brooktree's *Bt8953* interfaces the 8952 to T1/E1 framers to provide a complete solution. The transceiver incorporates adaptive echo cancellation and also equalisation to compensate for phase and amplitude distortion caused by boosting the bandwidth of low-frequency copper wires.
Brooktree Ltd. Tel., 0844 261989; fax, 261906.

Notebook data acquisition. Using a 486 notebook computer as a controller, Keithley's *DacPac* is a portable data acquisition system which avoids the restrictions of the printer port or PCMCIA slot. The unit runs all software written for the PC without modification. It is the same size as the computer and can be configured for the application by selecting the acquisition board from a wide range, two such boards being possible for high channel counts or more interfaces. DacPac has its own power system working from AC mains.
Keithley Instruments Ltd. Tel., 0734 575666; fax, 0734 596469.

OEM modem. *Socket Modem 136* by TDC supports fax, voice and data, providing a BABT-approved method of providing connectivity to PSTN and cellular networks. It can be fitted with a range of component options to allow configurations for 9600b/s data, 9600b/s fax send and receive and voice, supporting V.22bis, V.22, V.23 and V.21, plus V.42bis and MNP-5 data compression and error correction. Mobile systems are supported with MNP-10 protocol for cellular and poor landlines. ADPCM compression is used for voice data.
Telecom Design Communications Ltd. Tel., 0256 332800; fax 0256 332810.

Data acquisition box. A compact, lightweight, external data acquisition box by National, the *DAQPad-1200* is compatible with the parallel printer port of a PC and includes an AC adaptor or a 9-13 hour rechargeable battery pack. It has a 12-bit A-to-D converter accepting signals from eight single-ended or four differential inputs at up to 83.3ksample/s, with programmable gain of 1-100, a 512-sample fifo A-to-D buffer, two 12-bit D-to-A converters, 24 lines of TTL-compatible digital i/o and three 16-bit counter/timer channels accessible by the user. The unit measures 5.75 by 8.375 by 1.5in.
National Instruments UK. Tel., 0635 523545; fax, 0635 523154.

Development and evaluation

78K0 starter kit. NEC's distributor, Sunrise Electronics, has a low-cost starter kit for the *78K0* series of 8-bit microcontrollers which includes a board with a shrink-dip probe software, OTP programmer, cable and documentation. Assembler and debugger software provide full-screen interface, real-time execution and single-step, and multiple breakpoints. Program loading is via an RS232C interface to a terminal or PC, the program being downloaded and started under monitor control. The µP78P014YDW used has 32Kbyte of rom and 1056byte of ram, 53i/o ports, 16-bit and 8-bit counter timers, clock and watchdog timers, and two channels of serial interface, one of which being used, if required, as an I^2C bus interface.
Sunrise Electronics Ltd. Tel. 0908 263999; fax, 0908 263003.

A different ICE. To avoid the problems caused by the growing number of microprocessor derivatives and, therefore, an increasing number of pins, Hitex has a new method of emulation in which the processor stays where it is, soldered in if required, and is interfaced to the emulator by a standard interface adaptor, the *ICE/connect*, which is supplied with a subset of the processor's signals. The adaptor is identical for all derivatives of a processor family. *teletest51 Professional* and *teletest51 Junior* are supported for the 8051 family and *teletest32* for 80C165/166/167 families.
Hitex (UK) Ltd. Tel., 0203 692066; fax, 0203 692131.

£300 16-bit development. At a much lower cost than that of in-circuit emulation, The RCS *DB16 Designer's Board* forms a complete development kit for Mitsubishi microcontrollers. It consists of the board, Mitsubishi's cross assembler, communications package and debug monitor, power supply cable, example programs in source code and documentation. The user needs only a PC with text editing.
RCS Microsystems Ltd. Tel., 081 979 2204; fax, 081 979 6910.

Computer peripherals

Solid-state disk. AMC has a 'hard disk' on a PC card, having cold-boot and full disk emulation. The *PC Flash Disk* is based on flash memory and the *Flash File System* to provide sram memory at lower cost. It provides 32Mbyte capacity on one board, is non-volatile and needs no batteries, plugging into an ISA bus to replace both the disk and disk controller with its built-in software. The board is compatible with dos and Windows and works on 8-bit and 16-bit ISA/EISA buses and will operate as drive D: when other drives are present or as drive C: in diskless systems.
Advanced Modular Computers Ltd. Tel., 0753 580660; fax. 0753 580653.

Software

Maths training. A training package intended to increase the mathematical skills of student engineers in electronics, microcomputers, biomedicine and industrial subjects is available from Feedback. It contains fundamentals such as basic maths, algebra, trigonometry, calculus, computer maths and measurement. Emphasis is on graphics and animation and each subject area has a number of modules that can be taken as a complete programme or used selectively. Speed of learning is controlled by the student. The package automatically records performance and will provide class reports.
Feedback Test and Measurement. Tel., 0892 653322; fax, 0892 663719.

Graphical debugging. Designed for Innovate's *9000* board-test system, the company's *Index* is an interactive development system providing graphical waveform displays of stimulus and response data, with mouse-driven editing tools. It reads, modifies and writes test patterns and automatically checks for potential invalid data, data and address buses being handled in both hex and binary formats. Innovate Ltd. Tel., 0604 661000; fax, 0604 660021.

Forth 5 cross-compiler. MPE's PC-hosted Forth 5 cross-compilers generate code for a wide range of target processors, offering interactive development, high compilation speed and on-line help to speed the development of embedded software. Target code supplied includes a multi-tasker with message passing and event handlers; high-level interrupts interact with tasks.
Microprocessor Engineering Ltd. Tel., 0703 339691; fax, 0703 631441.

Real-time data analysis for Windows. *Hydra-Digis* from SciTech is a Windows 3-based, multiprocessor data acquisition and control system, handling data capture, real-time analysis and display and results documentation. A measurement and analysis process is designed on screen graphically, using a mouse to connect function blocks from a library, users adding their own functions or C-code routines where necessary. The transputer object code is generated by the software, the PC being freed for interface functions. Included is one card providing 16 analogue inputs, two analogue outputs, 16 digital inputs and outputs, a trigger input and a T805 25MHz transputer. The basic system will perform a 1024-point FFT to floating-point precision at a 5kHz real-time rate.
SciTech. Tel., 0734 758857; fax, 0734 758748.

Multimedia chips. *Mwave* is IBM's solution to the problem of expanding the PC's area of activity to encompass multimedia working. It is based on the company's *IBM38MW500* digital signal processing chip and allows the software implementation of all the expansion functions for a PC or workstation, including stereo sound, speech recognition, fax, modems and telephony. All this is performed by a single chip.

Blue Micro, IBM's British OEM representative, can supply evaluation boards in standard PC AT expansion card format, a range of software modules for basic functions already being in existence. Intermetrics has a complete developer's tookit, *Mwave for Windows*, and IBM has licensed *Sound Blaster* compatibility from Creative Technology, so that Mwave will be compatible with existing audio and multimedia applications. It is already compatible with CD-rom technology and the MPC standards. Mwave is also available to incorporate at chip level, with support available from Blue Micro.
Blue Micro Electronics. Tel., 0604 603310; fax, 0604 603320. ■

Computer voice control

Talk... is a computer system by Responsive Systems that listens to a voice and displays the text on screen, no intervention from the keyboard being needed. It uses Dragon Dictate and IBM Voice Type 2.

A built-in tutorial allows the system to learn one's unique speech patterns, three vocabularies, of 5000, 7000 and 30,000 words, being available in different versions. The package consists of a 486 notebook, although Talk... can be used with any desktop computer, hardware including 12MB of ram, a microphone designed to pick up very little ambient noise, 2000 code words in addition to the vocabulary, a 100,000 word dictionary for spell checking, adaptive learning to accommodate regional accents and speech impairments, and multiple user voice models for system sharing.

It works with many common dos applications such as Wordperfect, Wordstar, Word, Lotus 123 and Quicken, options including Windows voice control and Voice Tools to allow the development of specialist voice-driven applications.

Punctuation and formatting commands can be spoken or input from the keyboard, speed of text creation being around 30-40 words per minute; blocks of text can be input by a single spoken word. The whole process is faster than handwriting and faster than some typists.

This type of voice response may be added to any 386 or 486 computer provided that it has at least 12MB of ram, the cost being £999 for a 5000 word version. Completes systems start at £2899,in the 486/5000 word form. The system is modular and can therefore be upgraded.
Responsive Systems. Tel., 071 602 4107; fax, 071 603 2109.

LETTERS

From Douglas Self –

Amplified defence...

I thank Ben Duncan for his kind comments on my investigations into amplifier distortion (*Letters*, May 1994). But I also note with genuine sadness that his letter attempts to underpin a raft of unsupported assertions.

While I would happily slug it out line-by-line on every dubious statement, I doubt if The Editor would stand for it [*No. Ed*], so I will once more confine myself to power-supply rejection and "thermal distortion".

I don't know what Ben means by "economical with the facts". There were eight articles full of them. His plaintiff cry of "...where are those measurements?" is a bit rich in view of his propensity to complain without ever offering any hard facts to support his views.

Where Ben got the idea that I rejected ht regulators without trying them I do not know. I tried several versions of linear regulators in the 1970s – I still have one system in the cellar – in the innocent belief that this was a good way to reduce hum injection. But I was dead wrong. They are complicated, expensive, difficult to design for HF stability under all conditions, and generally a pain to use. It may be that Ben has some insights that render these criticisms outdated. But unless he is more forthcoming with information, it looks as if we shall never know.

Admittedly the generic amplifier I concentrated on is only one topology. But it is the basic structure of 98% of power amplifiers, and is surely a reasonable place to start. I do feel that I have shown that first-class 8Ω thd figures are obtainable with a simple unregulated supply, using reservoirs no larger than 10,000μF. I assure Ben that this also holds true for 4 and 2Ω loads, with or without various amounts of parallel capacitance up to 2μF.

Perhaps the topology that Ben prefers, but refuses to disclose, cannot emulate this performance. If so it seems like a good reason not to use it, because he has so far failed to specify any countering advantages. All we have to go on are the nuggets of data published in his article (*PSU regulation boosts audio performance*, pp.818-822, *EW+WW*, October 1992) which I reproduce below:

	10Hz	1kHz	20kHz
PSR with unreg psu	nil	–24dB	–57dB
THD with unreg psu	<0.002%	not stated	0.04%
THD with regulator	<0.006%	<0.006%	<0.006%

The data tells us that Ben's basic topology passes lf rail-perturbations straight through to the output – an unpromising start. Such an amplifier certainly appears to be unusable without very good supply regulation and I think it is only fair to wonder what is wrong with it.

The thd figures are even more enigmatic. 0.04% is fairly reasonable for hf crossover distortion at 4Ω, but it is difficult to see how any amount of supply regulation can improve this distortion mechanism. Furthermore, the 10Hz thd appears to have *deteriorated* by a factor of three and is deeply suspicious. I don't see why we should accept this without explanation.

As for my own psr vs frequency plots, I must admit that they languish in a big file labelled "pending".

PSR in a typical power amplifier depends on several overlapping mechanisms, not unlike distortion, and it will take a little time to disentangle properly. In the meantime, I would dearly like to know about these ominous "PSR-reducing real-world compromises" that Ben waves at us, but have little hope that he will tell us.

A brief explanation for Ben as to why regulated supplies are universal for powering the op-amps in mixing consoles. Typically, a power amplifier consists of one or two active stages, with all the circuit nodes available for judicious decoupling to keep the ripple out. Mixing consoles are made up of hundreds or thousands of active blocks, and the power distribution system is both extensive and intimately interlaced with the signal circuitry. Op-amps have a given psr, and if it proves inadequate there is little that can be done about it. Obviously in this case it is more sensible to remove the noise and ripple just once, rather than many times over. In particular it is vital to keep the garbage out of the audio grounding system.

Turning to thermal distortion, I find Ben's complaints unimpressive. The reason I am wasting my time looking for thermal distortion in the output devices is because he started this particular hare. But I must point out that closely examining the thd figures at 10Hz would disclose a thermal mechanism no matter where it lurked in the amplifier: a class-B output stage would be the likeliest place to find such distortion because the dissipation varies so radically over a cycle. It is also fair to assume that it would rise with falling frequency.

However, nothing can be found, even at the 0.0005% level.

I do not accept that my results are specific to the generic/Lin topology. There is no reason why they should not apply to any power amplifier – unless it has a very peculiar architecture. Possibly Ben's mystery topology is as susceptible to thermal distortion as it is to non-ideal power-supplies.

As for "thermal distortion being the HiFi industry's best hypothesis", it is news to me and certainly doesn't seem to have appeared in audio magazines recently.

I also protest about the current vogue for calling sine-wave testing "static" when a sine is anything but, having non-zero differentials up to as high an order as you care to go. Obviously Ben hopes to imply that such testing is naive and outmoded compared with some unspecified holistic new-age approach. But, mathematics is mathematics, and so a single frequency is always going to be best for finding out what actually happens inside an amplifier.

I was saddened to see that Ben regards me as an "adversary" when I had hoped we were fellow-seekers after Truth. But it is so easy to criticise if you never produce a rival design yourself. So I challenge Ben to publish an amplifier design that meets all his criteria for quality. Then we may all look at it. OK Ben... over to you.

...sad reignition

I read Jerry Mead's letter (*EW + WW*, March) with a certain sadness, which was I fear the only emotion I felt when composing my previous letter.

While I have no wish to reignite subjectivist wars in the letters section of *EW + WW*, I feel it must be wrong to let pass unchallenged erroneous statements on matters I have studied closely.

Regrettably Mead has answered none of the specific queries I raised: we still have no idea how he copes with the Hafler and Baxandall demonstrations that mystic amplifier differences just don't exist, and the

On the defensive

"For years industry has been crippled by the imperative for military development... Billions of pounds destined for the military have distorted the UK electronics industry grossly... The electronics industry must be led by marketing departments, not politicians..."
Frank Ogden, Comment, *EW + WW*, June 1990.

"Much of the relative prosperity in the UK can be traced back ultimately to our defence manufacturing activities... Arms, petrochemicals, nuclear technology ... are what separate us from a rural peasant economy".
Frank Ogden, Comment, *EW + WW*, April 1994.

Frank Ogden and Frank Ogden – are they by any chance related? Hypocrisy or MP bashing?
Richard Greatorex
London

Like it or not, the UK technology and electronics industry is still geared to defence interests and anyone who denies this would be very foolish. I don't welcome the situation, and I shall work to change it. However, I am not so much of an idealist as to ignore the money which it generates. Neither should our politicians through political expediency.
Frank Ogden

Brass monkeys

Thank you for *Help with hypocrisy* (*Comment*, April 1994).

As a professional observer and amateur practitioner for more than 50 years I have been horrified at the ineptitude of the top brass in British Industry and in particular the electronics, aerospace and automotive fields. GEC is certainly an outstanding example.
H Archer *Watford*

amplifier circuitry around the nauseating CMRR trimmer appears fated to remain shrouded in secrecy – a sort of *Trimmer Celeste*.

Mead has simply come back with more of the usual unverifiable, unattributable claims. He says that 'research has shown…' but what research? When was it done? Where can I can read it? If he has data that correlates his own audio priorities with sales figures, then can we not share some of this important information? If not, why not? And how do 'customers' in his letter turn into 'professional listeners' two lines further on?

Mead says one of his prime concerns is to correlate what he hears with measurement, and the implication of his first letter was that he wasn't making a lot of headway, though I hope he will forgive me if this is a misinterpretation. To say it one more time, there is a huge body of psychoacoustical knowledge already available for study, though I get the distinct impression this is of vanishingly small interest to subjectivists; a most strange state of affairs.

I fully sympathise with and share Alun Thomas' weariness with this debate (*EW + WW*, March), which in truth was resolved a long time ago, initially by a series of listening tests commissioned by Quad and carried out by James Moir in 1978.

Three radically different amplifier technologies (valve, conventional BJT, and current dumping) were subjected to blind testing and proper statistical analysis. The result was unequivocal; the amplifiers could not be distinguished by listening to them. Since then various other tests have produced similar results, and those that have not have proved to be fatally flawed as regards the statistical design of the experiment. It is in the light of these results that engineers feel little enthusiasm for repeating experiments that require a great deal of care and effort if they are to be of any use at all, and which can only confirm existing knowledge.

I'm sorry if Mead finds it irksome to have attention drawn to the fatal inadequacies of his experimental technique, but then *EW + WW* is hardly the right arena for those seeking uncritical adulation.

Whatever personal limitations Mead may choose to accuse me of – I assume he is trying to imply I have hearing difficulties, which is not very nice – he may rest assured that I shall continue to take an active interest in the phenomenon of uncontrolled listening tests.

Douglas Self
Forest Gate,
London

Ins and outs of amplifiers

Douglas Self's eight part amplifier series was an informative read. It brought back memories of my investigative audio experimentation twenty years ago and awoke some dormant ideas.

Amplifiers have two external audio inputs: normal signal input; and the loudspeaker output via the nfb (negative feedback) loop connection.

Suppose we connect one end of a 4Ω resistor to the live output terminal of an amplifier and drive the other with as little as 1Vrms with respect to output ground. To maintain zero output potential, the amplifier must then sequentially sink an extra 350mA in each half of its push-pull output stage.

In class A, device currents change equally but oppositely, smoothly and only fractionally, so there is little nfb loop disturbance.

In class B, virtually all current is derived from nfb-energised open loop gain. The output stage is thus driven through both device non linearities, inducing negative feedback modulation at frequencies two and four times that of the applied disturbance.

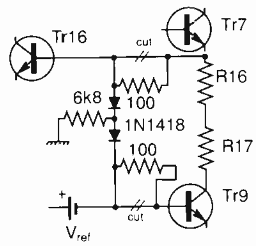
Composite loudspeakers have two or more driver resonances, with enclosures, cross-overs and leads turning up several more – well damped we hope.

Conceivably, complex programme drive will cause odd and unforeseeable back emfs, coupling almost directly with an amplifier nfb return. In class B designs, these in turn intermodulate any medium- and high-frequency audio components riding on large amplitude low notes.

It follows that higher levels of class B nfb increase the audible confusion while apparently improving dummy load measured performance figures. This could be the real reason why class A designs are subjectively pleasing, and we may need to reappraise testing procedures: what about measuring output stage distortion when it itself is driven and swept through 10Hz-50kHz by a reference power amplifier through complex impedances or real life loudspeakers? Or we could quiet test with both input and output driven – independently but synchronously.

Moving to another subject area, 20W of audio is weak by today's standards, but in class A it generates 45W of heat. Scale upwards to produce a decent 2×100W stereo amplifier and you have a small room heater.

Back in *2N3055* days, I had encouraging results with a sliding bias class A 100W rms design that quiesced with 'reasonable'



Sliding bias success

dissipation. There is now an opening to try this idea with Self's (Fig. 7) blameless design: I suggest the very simple circuit shown.

Note that with sliding A the output bias increases linearly with amplitude, differently to other classes, including AB and sustained current B.

The values shown in the circuit are estimates only. The 6k8 value should be carefully chosen according to power/impedance derivatives. C_4 should be as small as possible consistent with hf stability, and V_{ref} set to approximately four tenths of normal.

Induced current changes occur smoothly after voltage amplification stages and all is enclosed by the nfb loop, so measured distortion figures should hardly increase.

Standing currents are adequate to cope with real-life loudspeaker reactances and the output should still sound like good class A.

Tiny feet, now big, saw off my audio work and I'm unable to try these ideas. But if someone could do the soldering and testing, I'd be pleased to read their findings in this column.

Graham Maynard
Newtownabbey
N Ireland

Quiescent and controlled

In the last part of his series on distortion in audio power amplifiers (*EW+WW*, March 1994) Douglas Self suggests that it is not possible to control the quiescent current in class (A)B amplifiers by using an extra negative feedback loop. But I must disagree.

It can be controlled with a special kind of common-mode loop containing a non-linear network[1,2,3]. Using this technique, the quiescent current can be made independent of the temperature of output transistors, and a smooth transition from one side of the output stage to the other can be realised with ease.

The output transistors can be made to operate in a non-switching manner – although some parts of the non-linear feedback network will usually be cut off (nearly) completely when the output current is large, the non-linear network does not 'switch'.

Besides, with this method, common-source or common-emitter stages can be biased in class AB. Unlike Mr Self, I use the term class AB whenever the quiescent current is large compared with the junction leakage currents, but smaller than one half of the peak output current. Hence the output stage does not necessarily have to be some sort of (complementary) voltage follower.

Operational amplifiers using these class AB control loops have been described [1,2], as has a monolithic power amplifier with a class AB control loop[3].

Marcel van de Gevel
Haarlem
The Netherlands

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Power amp comments

I have been following with interest Douglas Self's series on audio power amp design.

On thermal runaway in Class A amplifiers, Self controls these by tight control of the quiescent current. The resulting intrusion of the stabilising circuitry into the amplifier circuitry seems to me to be inelegant. Given that Class A does not need to be precisely controlled, I would like to call attention to a couple of other methods.

First, constant power output PSU. If memory serves me correctly, Linsley Hood's 'Simple Class A amplifier' (*WW*, April 1969) used a PSU whose power output was essentially constant over times longer than a second or two, short-term peaks being supplied from the reservoir capacitors following the regulators. The amplifier normally consumes slightly less power than the PSU's maximum output. Should the amplifier move into thermal runaway, the system stalls like an overloaded gen-set and the output transistor dissipation does not rise. This approach also provides a measure of short circuit protection, and the circuit complexity is comparable with Self's stabiliser.

Secondly, boiling liquid cooling. To a first approximation, this approach will maintain the transistor case temperature at the boiling point of the liquid irrespective of the power dissipation. Heat is removed from the system by a condenser. It is obviously unsuitable for a commercial product below the level of the single-crystal pcb track merchants, but is not a problem for the enthusiast. Suitably meaty transistors will allow the use of water as the coolant (the air-cooled *2N3055s* in my amplifier have been running at boiling point for eight years), which brings two advantages: the availability and harmlessness of the coolant; and the possibility of tapping off the coolant to make the tea.

On dual-rail supplies versus speaker coupling capacitors, it is accepted practice to provide an amplifier with supply rails centred about 0V and a C-coupled output to avoid placing the speaker-coupling capacitor in the signal path. The logic behind this is obscure. Since the PSU rectifiers only conduct on the peak of each half cycle, for most of the time the amplifier output current is flowing through the PSU reservoir capacitors. If these are not a problem, what is so terrible about the speaker coupling capacitor? Is it simply that it is more obviously in the output signal path than the reservoir capacitors? The AC-coupled approach has two advantages: PSU circuit complexity is halved, and it is no longer necessary to ensure that the quiescent output voltage is precisely 0V.

Chris Bulman
Milton Keynes

Cables for hi-fi equipment

There is a grain of truth in some of Graham Nalty's statements (Complex cables defy physics, *EW + WW*, January) but the question is – how big a grain?

The veracity of any two statements does not mean one is contingent on the other. The door is green, the table is brown are both true, but the greenness of the door need not be anything to do with the brownness of the table.

The supercable suppliers and manufacturers are very good at selecting a diverse, convenient range of technical and scientific truths and implying that they are consistent.

If Nalty would qualify his assertions (and thereby propose a theory to explain his perception of an improved performance) then it might be possible to assess their validity.

For example, if, as he claims, silver (or silver coated) conductors make the system perform better then it must be because such conductors improve the signal in some way. The premise here, of course, is that if silver is better, then copper is worse, implying that copper in some way degrades the signal. If not, then they both degrade, but the silver less than the copper.

It would be helpful if Nalty would explain in quantitative terms how this occurs.

Drs R Yorke and BC Blake-Coleman
Southampton and Salisbury

Tack to basics

Douglas Self, (*EW + WW*, January), while preferring proper audio measurements to (improper?) listening tests, leaves out that the former are subject to restrictions inherent in Fourier's theory.

Theories are, after all, only convenient fictions to describe complex physical behaviour. To reach a manageable approximation of this complexity, theories are based on some selected simplifying presumptions.

In the audio field, for example, a theory built on well chosen assumptions would yield measurements in line with reality, so those with golden ears would agree with the engineers.

However, these well chosen assumptions will, apparently, have to come from thinking outside Fourier, considering that conventional reasoning merely resulted in 40 years of fruitless feuding across a gap between hearing and engineering. To try and bridge that gap, science could perhaps reveal how to compute Fourier's relevance to reality – maybe rated as percentage FRTR for common audio, radio and television, as well as for audio in discos and railway stations.

Trying to close the gap means research with Fourier kept on hold and only apt subjectivism to rely on in an exercise of purely practical pioneering.

Lately, perseverance in extensive, unorthodox experimentation has culminated in the deduction of precisely what, added to Fourier's theory, is needed to realise the breakthrough to novel, clearly superior responses; not to artificial signals, but to reality.

As this concerns not only electronics and acoustics, but also mechanics, especially suspensions. At least guessing FRTR has come within reach for all three disciplines.

HG Groenevelt
Rotterdam, Netherlands

Who do voodoo?

Some of the results of looking a little beyond today's simple audio performance criteria, as advocated by Jerry Mead (*EW + WW*, March), are to be found in the specialist magazines read by many hi-fi enthusiasts.

A non-technical acquaintance of mine is one such, and the methods he has been persuaded to employ in the search for improved sonic performance include: pinning the corners of his curtains back; sticking small pieces of transparent plastic around the room; variously using Litz, linear-crystal oxygen-free copper, gold-plated, bundled ribbon and ordinary mains wire for speaker cables; replacing mains plug fuses with solid copper rod; reversing one pair of speaker connections (only effective if done at the amplifier end!); painting the edges of all his CDs (to stop the light escaping!); and freezing his CDs, then wrapping them in damp towels and thawing them gradually in the fridge.

More voodoo than science, but all appeals to common sense and/or physics are dismissed as irrelevant. All that counts is that performance is perceived by the listener to be improved, as it invariably is.

Some may find these incidents amusing and relatively harmless, but serious money is involved as well. Hi-fi buffs are induced to part with large amounts of cash for special mains filters, passive preamps (sic), sand-filled speaker stands, sub-woofers and other paraphernalia. Moreover they must forever be upgrading their systems under the guidance of the magazine gurus in the endless quest for perfection.

I do not question Mead's sincerity, nor wish to imply that he is exploiting anyone. But would he take part in the sort of independent listening tests suggested by Alun Thomas? I doubt it.

Alan Dyke
Romsey, Hants

It's life... but not as we know it

Virtual reality is the current electronics hot issue. Now, by putting together several new technologies, we are probably as near as we will ever get to a Star Trek transporter.

The basic device would consist of a transmitter and receiver that can be plugged into a telephone socket. The receiver will be a virtual reality helmet with stereo headphones and two video displays. It will be connected to a box that decodes the video and audio data from the fractal compression system used to send it down the telephone system.

The transmitter will be a head-sized object with binaural microphones and a pair of television cameras, and data will be fed via a fractal compression unit to the telephone line.

Next stage would be to mount the 'head' on a small airship-type flying machine, and set its computer to keep it at head height. Then directional signals could also be sent to it from the receiving person.

Unfortunately, people would not be able to use this product to visit each other, as the person receiving the visit would still see the 'animated dustbin' or whatever the transmitter looked like where their

Unstable conclusions

What is the universe and how did it come about? When we see a stationary object we think of it as stable but it is not. The only way it can exist is to be in chaos. If it were stable it would have no energy and would not exist .

There was no 'before' the Big bang, because 'no energy' equals 'no time' and only chaos can exist. At the instant of the beginning only energy was produced.

We are told the big bang lasted only a billionth of a second, it may have taken a billion years because time was running slow. Energy is still being produced, though only a fraction of that at the beginning, and time is still fixing it into what we perceive as matter. The chaos of the universe is what has slowed new energy to virtually nil.

What the universe consists of today is exactly the same as when it did not exist. 'Nothing' time and energy have there equal and opposites rather like an equation that equates to zero. We exist but equal nothing .

The speed of light is treated as a constant but as it is energy and has no time in what time frame should it be measured?: big bang time or *EW+WW* letters time.

J Hancock
Bournemouth

friend was supposed to be.

But a simple trick could be used to get over this. The person being visited could also wear a virtual reality helmet. It would display his home image, but instead of the animated dustbin it would display the image of the person making the visit, recorded by camera at the receiver end. Editing out the VR helmets would not be impossible, so each person would see the other as if they were a real person.

John de Rivaz
Cornwall

Engineers see sense?

I was interested to read the letter by John R Owen (*EW + WW*, April 1994). At long last, engineers are beginning to understand the subjective effect of physical phenomena on the sense organs. In the past we have described neural networks purely in terms of circuits and signals. Living biological cells are different, especially in vision, where artificial systems do not appear to have any subjective experience of the images that they 'see'. Artificial vision is but a mathematical and analytical abstraction of biological vision. That is probably why, after forty years, researchers still ask why it is so difficult to get computers to do the simple things we humans take for granted?

In *Machine vision for people*, (*Update*, *EW+WW*, April) reporting on work being done at MIT, you mentioned the vital problem of the biological neural cell and electronics microchip interface. In the same issue, *Research Notes* reported how professor Warwick's robotics research was shedding light on the differences between animate and artificial robot systems.

Finally, the never-ending saga on audio reproduction quality and perception reminds us of the subjective nature of our senses. There might never be a machine that will experience pleasure as we living entities do. Perhaps the Bible may yet prove to be the ultimate authority on these matters.

Thomas McIndoe
Glasgow

Dos not dozzzzzz

Your magazine is probably not the place for extensive *Windows* vs dos debate, but I would like to take issue with Jason Ross (*Letters*, April 94) who makes the common mistake of assuming an unnecessary complexity in the operation of a system with which he is not familiar.

Ross's criticism of dos refers to "...changing to the directory containing the required program and then trying to remember its name before typing it".

Believe me, if I kept all my word processor files in a single directory there would be so many I would never find the one I wanted.

Ever since I read Kris Jamsa's useful book (*MS dos Batch Files*, Kris Jamsa, Microsoft Press, £9.95) I organise my files on the following lines: a directory reserved for batch files in my default path defined by a PATH statement in AUTOEXEC.BAT; sub-directories for each project and batch files associated with them. These have suitable mnemonic names and change activity to the sub-directory, send a set-up string to the printer (my word processor is old fashioned); use PATH and APPEND commands to provide access to the WP and associated files; call the WP program (with any appropriate command line switches); restore to root directory, and restore the default PATH.

A similar system is used for engineering projects which require several programs: a draughting program, a WP program, and possibly *Power Basic* for specialised calculations, etc. Batch files in the project directory would call up each program (preceded as necessary by PATH and APPEND) and a further batch file QUIT restores to the root directory and the default PATH.

Two tips are to change the PROMPT when in the project directory and restore the standard PROMPT when QUITing. Also, if using for programming, arrange your edit batch file (E.BAT) to assemble the program automatically when editing is completed.

It sounds complicated, but it only needs to be done once. Setting up a new project merely involves raising a new directory, copying the initiating batch file with a new name and editing the directory name.

Finally, a gui (graphical user interface) is not a panacea. Inevitably there will be problems, however standardised the applications interfaces under *Windows* or the Mac. My feeling is that three quarters of these will not be solved by introduction of a gui.

For example, I am very pleased with the ability of my recently-purchased *Wordstar Laptop* to make use of Times proportional font in my HP Deskjet printer. It can do footnotes and endnotes, and can print in two columns. But I cannot get the footnotes and endnotes to fit in columns. (Any advice would be appreciated).

David Noble
Essex

PC paranoia

Jason Ross (*Letters*, April 1994), has misunderstood the views I attempted to express in a previous letter (December 1993). My aim was to question the motivation of the electronics and computer press in excluding machines not using a Microsoft operating system. They also ignore the fact that *Windows* is so poor in comparison with, for example, Acorn Risc OS.

Many of arguments used to justify the 'PC only' policy are spurious. Reviewing an Acorn *A3020* machine one expert complained: "...it only has 2Mbyte of ram and a floppy disc, I use a PC with 4Mbyte and a 225Mbyte hard drive".

Bully for him! But perhaps he should have been asking why it takes so much extra hardware to get from *Windows* anything like the same level of functionality given by Risc OS on the Acorn.

Another recently urged schools to replace their BBCs with PCs. One reason was that: "...high flyers will need to know how to install *Windows* applications".

Ignoring the fact that he assumed schools were still limited to 32K-byte machines designed in 1981, surely the question is why it should be so difficult a task. Working under Risc OS, which many schools will be running on Acorn machines, the task is trivial. Why not under *Windows*?

Most personal computers probably run no more than two or three programs regularly. So if you are never going to need them, what does it matter that there are several times more programs available for the PC than for the Atari, Amiga and Archimedes put together?

A good wimp based gui (graphical user interface) is a pleasure to use, as Joel Sciamma found with his Mac (*Letters*, April 1994). The reluctance of some PC users to swap from dos to *Windows* suggests that for them its advantages fail to outweigh the problems.

No doubt Windows will eventually mutate into a fast, compact and convenient operating system to be hailed as yet another triumph for Microsoft. But this would happen a great deal faster if the trade media dropped their infatuation with dos and *Windows* machines. In particular it might be useful to differentiate between what is vaguely promised in press releases and what is actually being delivered today.

Simon Wyre, (*Letters*, March 1994), suggests Acorn open up its architecture to third parties. Arm chip sets, used in Acorn's machines, can be bought by any manufacturer, but British computer makers don't appear to want them. So the *Arm600* chip is exported back to us in the Apple *Newton*.

Is no one planning to incorporate Arm chips into a new product? It might even merit a mention in *EW+WW*.

Les May
Lancs

Bose and bouquets

Several readers have pointed out that in my article (*Using momentum to dethrone Einstein*, *EW+WW*, September 1993) I omitted to mention the important work of Indian physicist Satyendra Bose, which provided a firm mathematical basis for quantum theory.

Although Planck described the radiation of energy in terms of quanta, he did not regard light as actually consisting of packets of energy. For him, the concept was only a mathematical device – one which could explain the interaction between light and matter.

The Indian physicist Meghnad Saha thought that photons were more like particles than waves, and he was able to explain radiation pressure in terms of light quantum.

Bose, working in obscurity in East Bengal, took a similar view. In 1923, he attempted a mathematical derivation of Planck's law of black body radiation. He succeeded only when he treated photons as though they were particles. His paper on the subject makes not even a mention of either waves or electromagnetism.

To derive Planck's formula, Bose had to employ a new kind of statistics. Uncertain of its validity, he sent his paper to Einstein, who recognised its implications, translated it into German, and had it published in the journal *Zeitschrift fur Physc*. Einstein sent a copy of Bose's paper to Erwin Schrodinger, together with Louis de Broglie's famous paper which predicted that electrons can behave either as waves or as particles.

In a letter to Einstein, Schrodinger frankly admitted that it was Bose's paper that made him realise the importance of de Broglie's prediction.

The years that followed saw the development of quantum theory. But Bose took no further part in it. He returned to a subject nearer to his heart – the teaching and popularising of science. Just before his death in 1974, he described himself, rather sadly, as a comet which came once and never returned again".

John Ferguson
Surrey

■

CIRCUIT IDEAS

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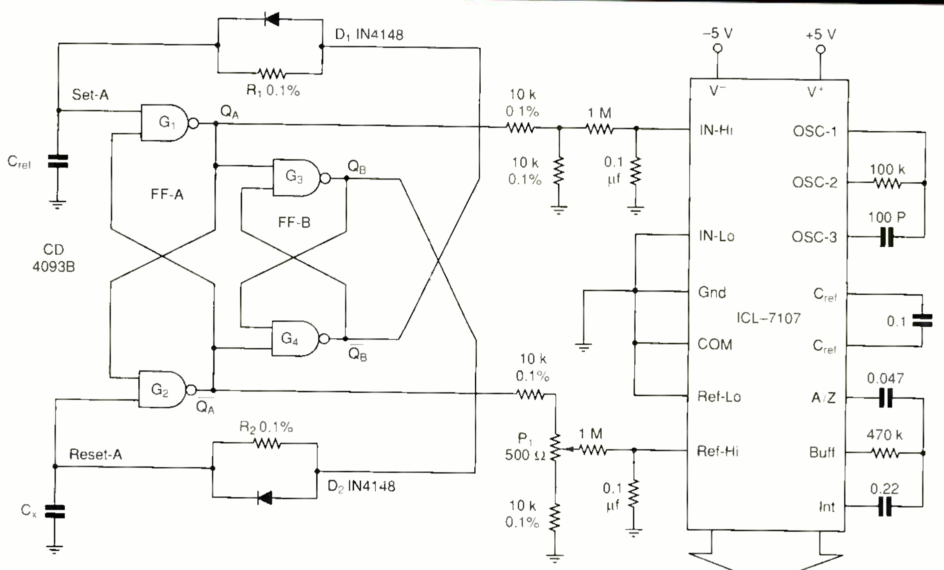
With only two ICs and without the need for a regulated supply or a crystal clock, this arrangement measures the capacitance ratio between two components to within 0.3%, presenting the reading digitally.

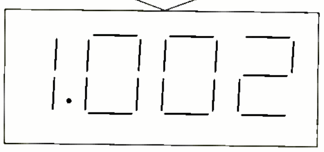
A latch-up-free astable flip-flop, made up of the four gates G_{1-4}, produces complementary square waves, the on times of which are independently determined by C_x and C_{ref}. At switch-on, the two capacitors are at zero and QA and QA\ are both 1, the second flip-flop accepting this input condition and toggling one way or the other indeterminately. Assuming that QB is 0 and QB\ is 1, QB\ charges C_{ref} to the supply voltage by way of D_1. G_1 and G_2 inputs are now 1 nd 0 respectively and the outputs become QA and QB\ = 0, QA\ and QB = 1. *Cx* now charges and *Cref* discharges via R_1, this voltage toggling the flip-flop when it reaches the threshold of Set A. The new state discharges C_x and the cycle repeats.

The *ICL7107* 3.5-digit A-to-D converter now indicates the ratio of integrated average voltages at the high and low inputs, its "thousands" decimal point being turned on. A potentiometer on the reference input is set so that the display shows a reciprocal reading when the capacitors are interchanged.

M S Nagaraj
ISRO Satellite Centre
Bangalore
India



Cref - range	R1 = R2
50 – 100 P	1 M
100 – 1000 P	100 k
0.001 – 0.01 μ	10 k
0.01 – 0.1 μ	10 k
0.1 – 1 μ	1 k



Capacitance ratio meter with a maximum reading of 2:1 and a range of 50pF-1μF. Accuracy is 0.3% and no voltage regulation or clock input is needed.

A simple way to high-pass and band-pass filters

Using only a low-pass filter consisting of an *R* and a *C*, this circuit will function as an active high-pass filter and, with the addition of a similar circuit with other values, a band-pass filter.

Inputs to the *AD620* instrumentation amplifier are direct *and* via the *RC* low-pass section. At low frequencies, both inputs see the same signal, since the *RC* filter passes the signal and there is a common-mode input to the amplifier. Higher frequencies are attenuated by the filter and the amplifier receives the difference, which is amplified.

Effectively, the output of the circuit is the amplified output of the derivative of the *RC* section:

$$\frac{V_2}{V_1} = A\left(1-\frac{1}{1+s\tau}\right)$$

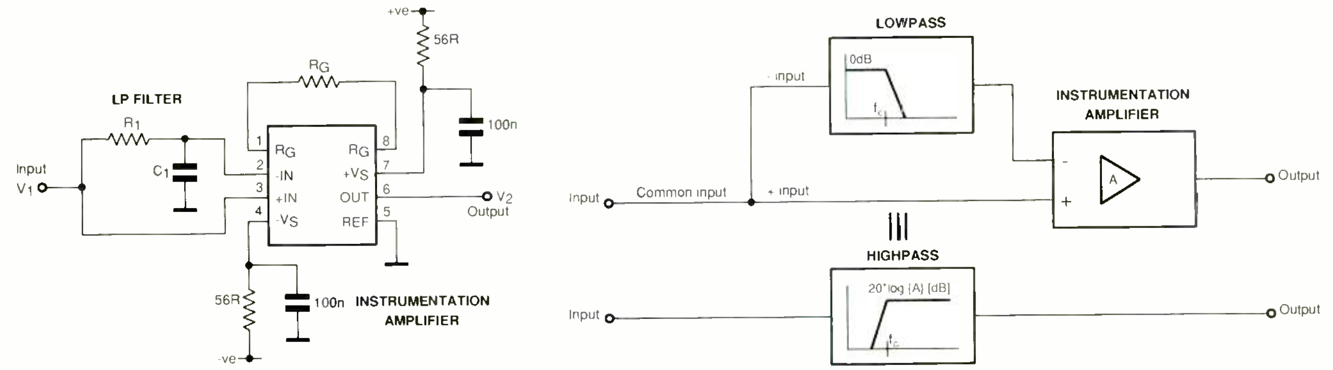
$$= A\frac{s\tau}{1+s\tau}$$

where $\tau=R_1C_1$ and s is the complex frequency variable.

A band-pass filter can be made by inserting another *RC* section, preferably in the common input before the original filter to give the usual low-pass/high-pass cascade, or in the non-inverting lead. These arrangements give different results and, if the cut-off frequencies are identical, the second connection gives zero output, while the first still behaves as a band-pass filter.

The circuit has found application in the amplification of a small, higher frequency signal superimposed on a slowly changing one.

Jaroslav Chum
Geophysical Institute
Prague, Czech Republic



One RC low-pass section followed by a differential amplifier becomes a high-pass filter and can be made into a band-pass type.

Binary-to-BCD converter

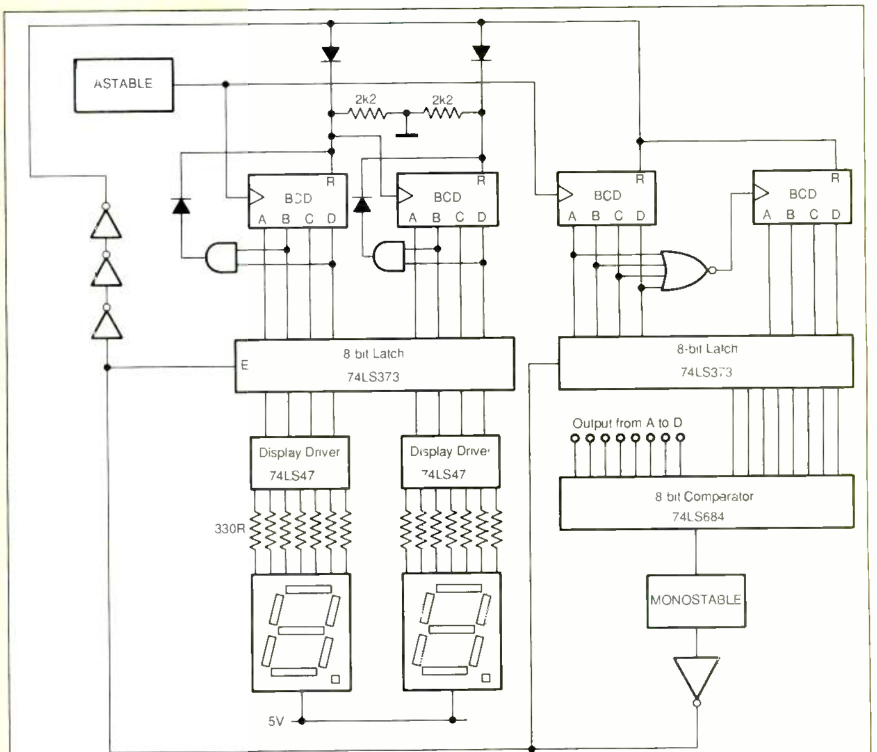
Converting an eight-bit binary input to binary-coded decimal can, on occasion, be useful in showing the output of an analogue-to-digital converter in decimal form.

In essence, the circuit clocks binary and BCD counters together until the output from the binary counter equals the input from the binary A-to-D converter, only the BCD output being indicated.

A 555-based clock at 7.3kHz drives both sets of counters to a maximum of 99 counts or until equality between the inputs from the A-to-D converter and the binary counter is detected by the *684* comparator. At this point, the comparator triggers a monostable flip-flop, which latches the BCD counter output so that the count is displayed. After a short delay, the same signal resets the counters, which remain reset and the count displayed for about 1s until the monostable pulse finishes, when the process recycles. Very little flicker is evident, since the process is fast.

To avoid the circuit cycling when the count is correct, a second latch between the binary counter and the comparator, so that a constant equality signal from the comparator and monostable ensure that the display only updates when the input from the A-to-D converter changes.

Richard Maggs
Rhiwbina
Cardiff



Clocking BCD and binary counters together and displaying the BCD output effectively converts binary to decimal, allowing the display of, for example, the output of an A-to-D converter. As it stands, the circuit is suitable for a 99-count input.

Special-purpose 74-series binary-to-BCD parts are a simpler solution for two digits but this circuit may prove more convenient as the number of digits increases – Ed.

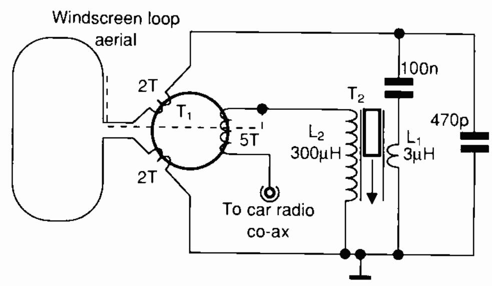
Car radio loop aerial

This windscreen loop aerial is suitable for long and medium waves and VHF FM. An existing whip can be left in situ or an extra wire can be run parallel to one leg of the loop.

The 100nF capacitor, series inductance of L_2 and the loop tune the aerial to long wave, while medium waves are tuned by L_1 and the capacitance of the coaxial cable from the aerial to the radio. For VHF, the ring core T_1 has an outside diameter of 6.5mm and T_2 is the oscillator coil from a MW receiver, with the slug fully in.

There is a performance improvement over a whip, particularly in tunnels.

D Di Mario
Milan
Italy



Windscreen loop aerial foils vandals and improves performance.

High-voltage, current limited power supply

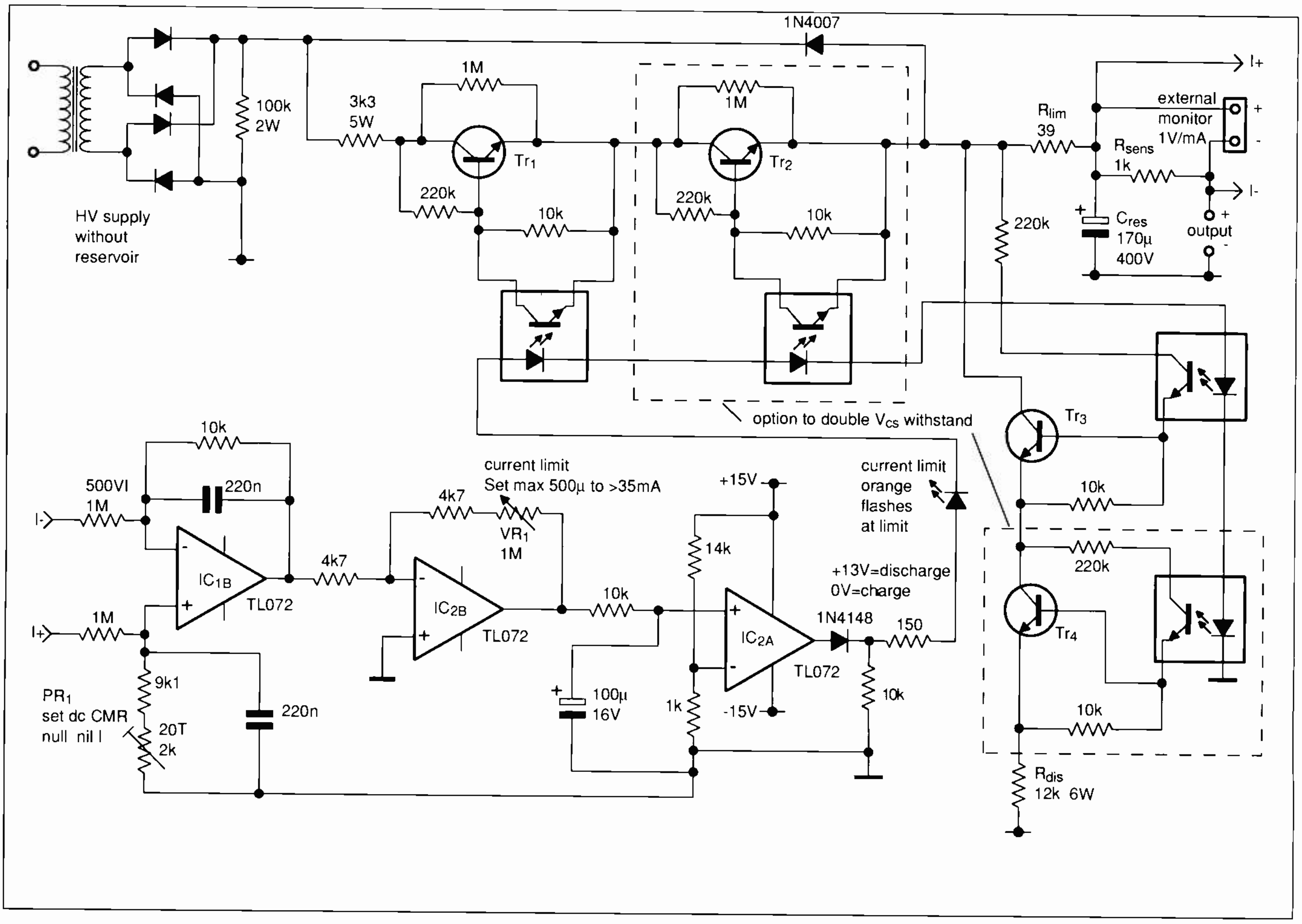
This inexpensive supply was designed to test the breakdown voltage of components and high-voltage semiconductors, using switching techniques to supply current-limited outputs from 100V to 500V at low dissipation. The output voltage is unregulated, ramping up until the set current limit is reached.

Input is up to 500V of raw DC, any reservoir capacitor being removed and possibly reused as C_{res} at the output. A switch at the input would allow the test voltage to be applied after connecting the device under test. Collector/emitter voltages of $Tr_{1,3}$, or $Tr_{1+2,\ 3+4}$ if the the extra devices to double the voltage rating are used, must withstand the input voltage but, since all the transistors are either on or off, dissipation is low and safe operating-area derating is unnecessary; TO-92, E-line or TO-126 packaged devices may be used. Collector-current ratings should be well above the current limit. The reservoir capacitor C_{res} should have low equivalent series resistance, R_{lim} being set to limit inrush current.

Amplifier IC_{1b} reads the voltage drop caused by output current in R_{sens}, which can be varied in value to allow current limits other than 20mA. Although common-mode voltage is irrelevant in a floating supply, the input to the amplifier is attenuated to allow grounded HT supplies, IC_{1b}'s ±10V CM limit coping with ±500V at the input if the 1MΩ resistors are properly rated.

Gain of the second amplifier sets the current limit and its 3MHz gain bandwidth product limits servo loop bandwidth and rate to around 15kHz at a gain of 200. Used as a comparator, the third amplifier controls the opto-couplers.

Ben Duncan
Lincoln



High-voltage current limiter/stabiliser

The basic circuit shown in Fig.1 is a current limiter using a junction fet, which operates according to

$I_{lim} = -V_{gs}/R > I_{ss}$,

in which $0 < -V_{gs} < -V_{gs(off)}$.

For higher power, the circuit of Fig. 2 is better, the mosfet being fitted with a heat sink. The mosfet holds off the supply voltage and R_4 is a current sensor, the voltage across it driving the bipolar transistor base to apply feedback round the circuit. A 10V zener provides a supply to R_2Tr_1 and, since it is connected to the top of R_4, takes into account the current in R_1.

Asthe circuit stands, the current limits at 100mA, with an output impedance of around 100kΩ and with a maximum applied voltage of about 175V although, since 1000V mosfets are available, the circuit can be scaled up.

Replacing the bipolar transistor with a programmable zener gives better stability and a higher output impedance of about 1MΩ, both the quoted impedances being estimated from the performance table shown. ■

CJD Catto
Cambridge

	voltage	10	15	20	35	50	100	110
Fig 2	current	94.0	96.5	97.5	98.5	99.0	100.0	100.1
Fig 3	current	67.0	93.0	99.5	100.0	100.05	100.1	100.1

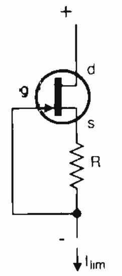


Fig. 1. Essentials of a fet current limiter, where R_4 senses the current.

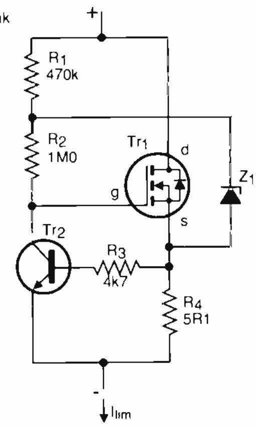


Fig. 2. A mosfet handles higher power, the bipolar device providing feedback to the mosfet base.

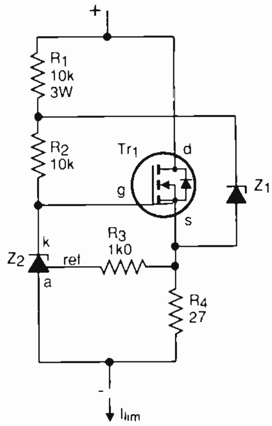


Fig. 3. For improved current stability and higher output impedance, the programmable zener replaces the transistor. From the table shown, an estimate of output impedance is 1MΩ. Note the 3W resistor at R_4.

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Filtering out design problems

***Choice of a filter for rf power amplifiers must not only take account of intended applications, but also of the components available. Norm Dye and Helge Granberg explain the theory – and practice – of good filter design. From the book* RF Transistors: principles and practical applications.**

Norm Dye is Motorola's product planning manager in the Semiconductor Products Sector, and Helge Granberg is Member of Technical Staff, Radio Frequency Power Group (Semiconductor Products) at Motorola. Their rf transistors book includes practical examples from the frequency spectrum from 2MHz to microwaves, with special emphasis on the uhf frequencies .

Filtering for output harmonic reduction is needed in virtually all solid state rf power amplifiers – especially those for radio communications. Specifications depend on the application, frequency of operation and power level.

Wide-band amplifiers have much higher harmonic content than narrow-band ones, the worst cases existing at the lowest frequencies. An amplifier with a bandwidth of 10-175MHz, for example, would not only pass the second harmonic up to 85-90MHz and the third harmonic up to about 60 MHz, but will amplify these frequencies along with the fundamental.

Harmonics generated at higher than these frequencies fall outside the amplifier's bandwidth. Their amplitudes will gradually diminish at increasing frequencies, at a rate depending on the amplifier's gain characteristics outside its intended pass band.

At 10MHz, the same amplifier would pass and amplify signals all the way up to the seventh even and fourth odd order harmonic. Viewing the fundamental waveform on an oscilloscope would show it to resemble a square wave more than a sine wave – which is normal considering the wide distribution of harmonics. In a single-ended amplifier, the second harmonic is the most troublesome. But in push-pull amplifiers, the even-order harmonics are suppressed. In one with a 4-5 octave bandwidth, the amplitude of the second harmonic can be 30-50dB below the fundamental, depending on the circuit balance, yet the third order harmonic may be attenuated only 10-12dB.

In narrow-band systems, band-pass filters can be used for rf power amplifier output harmonic filtering. But such filtering is almost exclusively done with low-pass filters.

Extending normalised tables

Normalised tables that help in designing low-pass filters – available in textbooks and other publications – can be used for designing high-pass, band-pass or low-pass varieties. Only a few simple calculations are necessary to derive the normalised values from the tables, and filters so designed can be computer analysed to verify their accuracy.

The technique is applicable up to frequencies where lumped constant elements cease to be practical. But at high uhf and microwave frequencies, microstrip techniques can be used to realise the components needed in the low-pass filter design. Chip capacitors can be used up to 800-1000MHz, but lumped inductors become ineffective above 500MHz.

When calculating lengths of line in microstrip to obtain distributed *Ls* and *Cs*, remember that microstrip is not a true

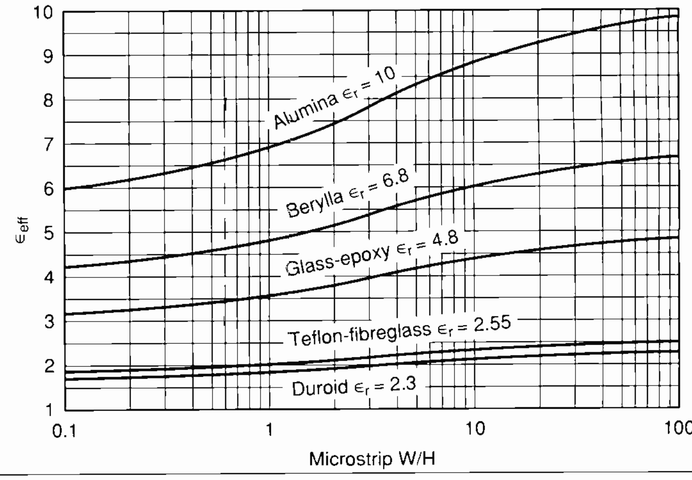


Fig. 1. Effective dielectric constant versus the width-to-height ratio for some of the most popular microstrip substrate materials.

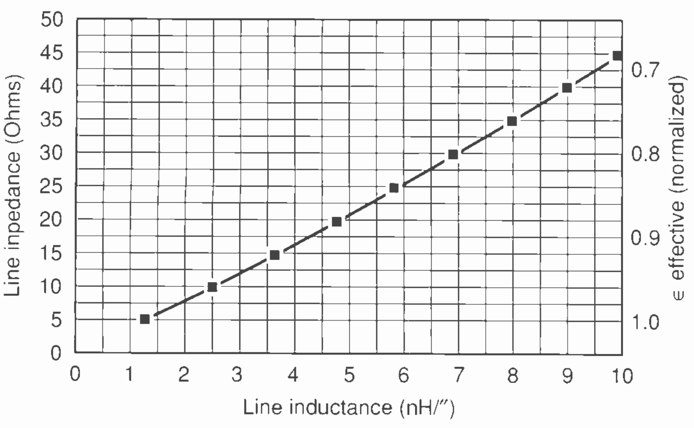


Fig. 2. Conversion of microstrip line impedance to inductance per unit length. The scale at the right shows the ε_{eff} relative to ε_r vs line impedance.

transverse electric magnetic (tem) mode of propagation. Velocity of propagation is reduced by the medium in a similar manner to that encountered in true tem mode transmission lines such as coaxial cable and stripline, but, to lesser extent.

Velocity reduction, similar to wavelength reduction discussed previously (March, *EW + WW*), can be expressed as $v_m = v_0/\sqrt{.}\varepsilon_r$.

In the case of microstrip, the 'effective' relative dielectric constant ε_{eff} is related to ε_r by a formula involving the W/h ratio of the microstrip line.

The formula:

$$\varepsilon_{eff}=(\varepsilon_r+1)/2+((\varepsilon_r-1)/2)(1+(12h/W))^{-0.5}$$

shows that ε_{eff} approaches ε_r as the term h/w becomes small – the condition occurring as the characteristic impedance of the line approaches zero.

Microstrip design

Common practice when designing microstrip circuits is to use curves that show the effective relative dielectric constant as a function of W/h of the microstrip (**Fig. 1**).

By using the 'effective relative dielectric constant,' velocity v_{eff} can still be referred to in the medium, given by the expression:

$$v_{eff} = v_0/\sqrt{.}\varepsilon_{eff}$$

where v_0 is the velocity of light in a vacuum (3×10^8m/s) and ε_{eff} is the effective relative dielectric constant. The length l of a microstrip line for a specific inductance is:

$$l = Lv_{eff}/Z_0$$

where L is inductance in nH, v_{eff} is effective velocity in the medium given in the expression above, and Z_0 is the characteristic impedance of the line in ohms.

If we select the desired value of L to be 2.5nH and use microstrip line made with alumina (ε_r = 10) having a Z_0 of 10Ω, then ε_{eff} = 8.9 (Fig. 1) and $v_{eff} = 30/\sqrt{.}8.9 = 10.1$ cm/ns. Substituting in the above formula: $l = (2.5 \times 10.1)/10 = 2.53$cm.

A plot of inductance for different line impedances is shown in **Fig. 2**. For convenience, ε_{eff} normalised to ε_r is also plotted on the vertical axis. The graph clearly shows how the value of ε_{eff} approaches ε_r as line impedance approaches zero.

Note that in the above example, Z_0 = 10Ω and Z_{in} and Z_{out} are assumed to be 50Ω.

We have seen previously (Fig. 6, March, *EW + WW*) that W/h = 10 for 10Ω line if the dielectric material has a relative dielectric constant of 10. In practice, if the line impedance is not more than 10% of Z_{in} and Z_{out} – ie less than 5Ω, the value of ε_{eff} can be assumed equal to ε_r.

The industry standard characteristic impedance for rf power test systems and equipment is 50Ω, so all subsequent material

Chebyshev or Butterworth?

Two basic types of low pass filters are commonly used to suppress harmonics of the desired frequency in the rf output of solid state amplifiers: Chebyshev and Butterworth.

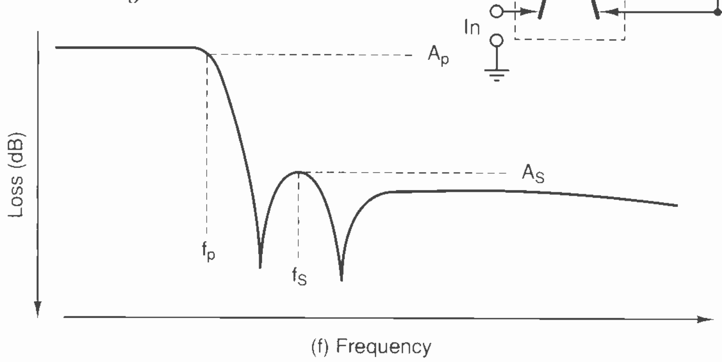
Several variations of each include the elliptic function filter (also called the Cauer-parameter filter), the constant-K filter, and the m-derived filter.

Each has its own characteristics, performance, advantages and disadvantages. In addition filters can be divided into inductive input and capacitive input categories.

Most rf power applications make use of either a straight Chebyshev or one modified to provide the elliptic function characteristics – having a sharper cut-off, but lower far band attenuation, as shown below.

The modification also provides deep and sharp notches in the out-of-band attenuation. Some of these notches can be fine-tuned to the specific harmonic frequencies, improving out-of-band attenuation but decreasing return loss. The option is also only practical in fixed frequency or very narrow frequency range applications.

Fig. 3. Typical response curve of a six element elliptic function low-pass filter. The A and f designations refer to the definitions given in text.



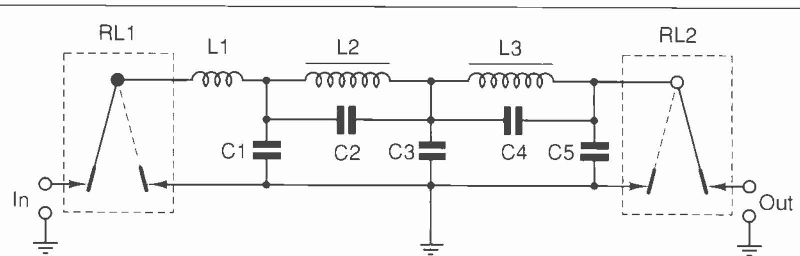


Fig. 4. Schematic of a six element L-input elliptic function low-pass filter. The relays RL1 and RL2 were added to make possible switching of multiple filters.

and data refer to this impedance unless otherwise noted.

Choosing a filter

Using only a plain Chebyshev or Butterworth filter to suppress harmonics of the desired frequency in the rf output of solid state amplifiers makes it is difficult to obtain sufficient attenuation. This is especially so for the second harmonic, except with a well balanced push-pull circuit which provides even order harmonic suppression by itself: the push-pull configuration has a cancellation effect, as mentioned earlier.

Low pass filters have – in general – characteristics which attenuate unwanted harmonics by reflecting them back to the source. The effect is to alter the impedance for the harmonics, which was originally 50Ω without the filter.

Series-*L* input type low pass filters present a high impedance for the harmonics: with a shunt-*C* input type, the harmonics will see a low impedance load. In each case, linearity and efficiency of the signal source are affected depending on the original harmonic amplitude.

A series-*L* input low-pass filter, when used with a solid state amplifier, can be expected to have superior efficiency characteristics. On the other hand, the shunt-*C* input filter results in better linearity and harmonic suppression, though it can result in abnormally high rf currents in the line connecting the signal source to the filter.

In the shunt-*C* input filter, harmonics are directed to ground, so the amplifier will see a lower than optimum output load line – especially at the harmonic frequencies.

The filter's effect is reduced when inductance is added between the amplifier output and filter input. Also, a transmission line will convert it to a series-*L* input type.

In a series-*L* input filter the harmonics are either dissipated in the series inductance of the filter or reflected back to the amplifier output, presenting a higher than normal load line to the amplifier.

Severity of the above characteristics for each filter depends on the initial harmonic content of the amplifier and on frequency of operation in the pass band of the filter. The exact conditions under load mismatch in each case depend on the phase angle of the mismatched load, in addition to phase delay of the filter.

One solution to avoid problems is to use a diplexer in the amplifier output.

The device is actually nothing but a dual set of filters, one low pass and one high pass. Their cut-off frequencies are designed so that the low pass filter passes only the fundamental frequency to the primary load. The high pass filter presents a high impedance to the fundamental, but passes the harmonics and 'dumps' them into a secondary load. The solution looks ideal, since 50Ω loads would be presented to both the fundamental frequency and the harmonics. The secondary load is only required to handle the power level of the harmonics, typically up to about 10% of the power level at the primary load.

Table 1. Component values for six filters covering the 1.6-30MHz spectrum. The normalised values, as well as the actual values of the low-pass filters, are shown in the lower part of the table. The upper part shows the specified numbers from which the normalised and actual values are derived.

	30MHz	21MHz	14MHz	7MHz	3.5MHz	1.6MHz
f_p (MHz)	32.0	23.0	14.5	7.5	4.0	2.10
f_s (MHz)	48.0	34.0	20.0	10.0	5.8	3.20
Ω_s (f_s/f_p)	1.50	1.48	1.38	1.34	1.45	1.52
A_p (passband ripple)	0.044	0.044	0.044	0.044	0.044	0.044
A_s (dB)	–45	–45	–45	–45	–45	–45
C_{ref} (pF)	100.0	140.0	220.0	420.0	760.0	1430.0
L_{ref} (nH)	250.0	349.0	525.0	1050.0	1970.0	3650.0
Normalised values for C_N or L_N	**Actual values = C_NC_{ref} (pF) or L_NL_{ref} (mH)**					
C_1 = 1.190	119.0	161.0	262.0	500.0	905.0	1700.0
C_2 = 0.426	42.7	60.0	94.0	179.0	324.0	610.0
C_3 = 1.312	131.2	184.0	288.0	551.0	997.0	1875.0
C_4 = 0.232	23.2	32.5	51.0	97.5	177.0	332.0
C_5 = 0.696	69.6	97.5	153.0	292.0	529.0	995.0
L_1 = 0.899	225.0	314.0	472.0	944.0	1770.0	3280.0
L_2 = 1.118	279.5	390.0	587.0	1174.0	2200.0	4080.0
L_3 = 1.181	295.0	412.0	620.0	1240.0	2325.0	4310.0

But in spite of its advantages, the diplexer is practical only in single-frequency or extremely narrow spectrum applications, except for certain military equipment where space and economy are of less importance.

Design procedure

Design of low-pass filters using tables of normalised element values is relatively straightforward. All that is required is to set the specifications (**Fig. 3**):

1) Cut-off frequency f_p (in MHz);
2) Stop-band frequency f_s (in MHz);
3) Reference low-pass (W_s) = f_s/f_p;
4) Required stop-band attenuation A_s (in dB) at frequency f_s;
5) Allowable return loss *RL* (in dB) or voltage reflection coefficient $|\Gamma|$ (in %), which determines the band-pass ripple A_p (in dB);
6) Input and output impedances (*Z*).

Using the tables – from several sources – a filter type can be selected and its degree (how many elements are required to satisfy the conditions specified) determined.

Sometimes the compromise necessary to reduce the number of elements and complexity of the filter may result in only a minor degradation in performance characteristics. Normally eight filters are required to cover the 1.6-30MHz hf frequency spectrum without gaps, even with sharper cut-off elliptic designs. But sometimes frequency bands such as the ham radio spectrum up to 30MHz (1.6, 3.5, 7.0, 14, 21, and 30MHz) can be covered with only six filters of the same type because continuous coverage is not required.

Putting theory into practice

Let us examine a low-pass filter design with f_p = 32MHz and the lowest usable frequency of 24MHz, for example. Then f_s would be 2 x 24, or 48MHz, resulting in $\Omega_s = f_s/f_p = 48/32 = 1.50$. A_s is set at –45dB and the maximum allowable reflection coefficient $|\Gamma|$ at 10%.

For all the above reasons, a series-*L* input type filter seems to be the best choice. Two different configurations could be used: one designed with series-*LC* shunt elements; the other with shunt-*LC* series elements. Shunt *LC* is far more popular and is easier to implement in practice, since it uses fewer inductors. To make the filter a series-*L* input type, it must be an even degree, meaning that it must have an even number of elements.

Tables of normalised values for elliptic low-pass filters indicate that at least six elements are required for A_s = –45dB, $|\Gamma|$ = 10% and Ω_s = 1.50. If the reflection coefficient $|\Gamma|$ is converted into a decimal: 5/100 = 0.050, the mismatch loss or pass band ripple can be calculated as $-10\log_{10}(1-(0.05)^2) = 0.011$dB. This – as well as the actual values – is theoretical, assuming an infinite *Q* for the elements. There will not be a large change if the loaded *Q* is 100. A *Q* of 20 however would result in a considerable loss in A_p and a 1-2dB decrease in A_s. The largest influence of the reduced *Q* can be noticed in the depth of the notches, which are not of primary importance.

Next we find the reference *L* and *C* values:

$$L_{ref} = R/2\pi f_p = 50/201 = 249\text{nH}$$

and

$$C_{ref} = (1/R)/2\pi f_p = 0.020/201 = 99.5\text{pF}.$$

These values multiplied by the numbers from the normalised tables will result in the actual *L* and *C* values required for the filter elements. So values of the actual elements of the six element *L*-input elliptic function low-

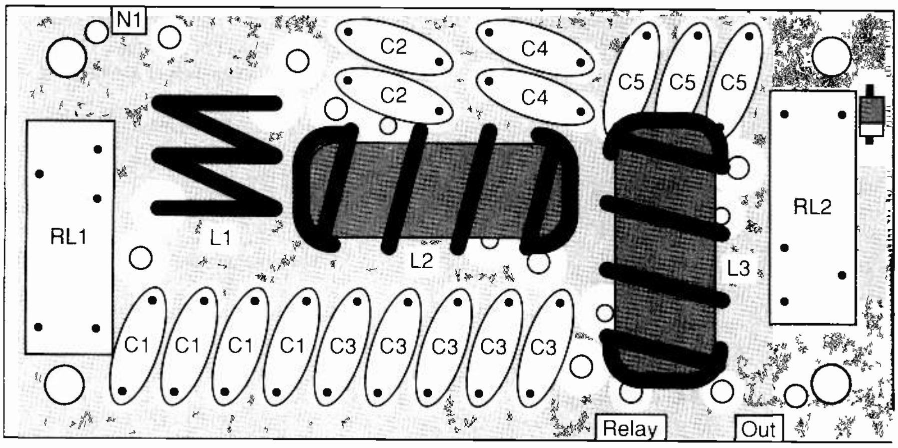


Fig. 5. Layout of a low-pass filter. Note the 90° positioning of the inductors and the paralleled capacitors in C1-C5. The grey area represents a ground plane on top of the board. RL1 and RL2 are provided for switching of multiple filters.

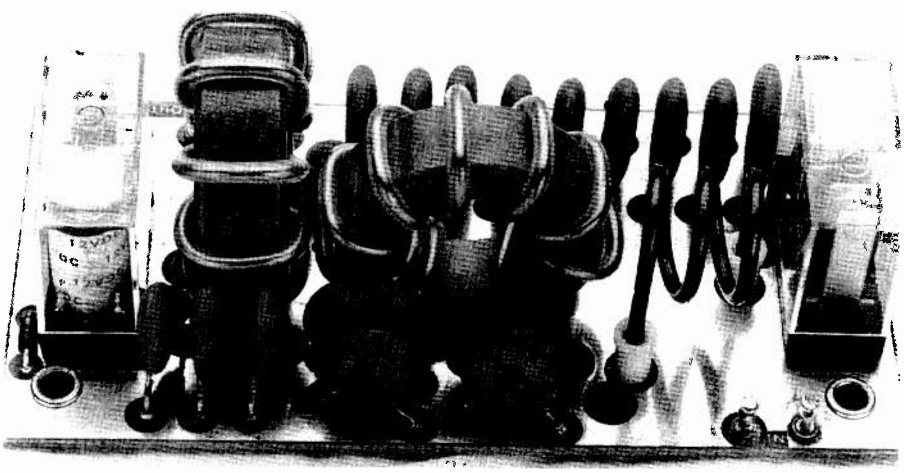


Fig. 6. Switched filter shown in Fig. 4.

pass filter (**Fig. 4**) will be: L_1 = 225nH, C_1 = 119pF, L_2 = 280nH, C_2 = 43pF, L_3 = 295nH, C_3 = 131pF, C_4 = 23pF, and C_5 = 70pF. The component values for six filters covering the 1.6 to 30MHz spectrum are given in **Table 1**.

In a practical design these numbers can be rounded to the nearest standard values. The inductors are somewhat limited in their tolerances, since fractions of turns, of course, cannot be realised and the capacitor tolerances are 5%, at best, as standard stock items. **Figure 7** is a computer plot of the expected response of the filter.

Component characteristics

Paralleled multiple capacitors filters are not always recommended. But in this case no abnormalities were noticed when the responses were checked with a sweeper and a spectrum analyser – although this may not be the case at higher frequencies. Paralleling of multiple capacitors allows their current carrying capabilities to be increased, and allows inexpensive disc types to be used. Composing values closer to the non-standards required in many instances is easier too. High voltage (3000V) types were selected, since under certain load mismatch conditions they may be subjected to high rf voltages. Typically the rf voltage ratings of such capacitors are approximately 30-35% of their dc ratings.

Inductors should be located so that adjacent ones are at 90° angles to each other, minimising the possibility of mutual coupling. L_1 is an air-wound inductor and L_2 and L_3 are wound on phenolic, non-magnetic toroids for their more convenient shape factor.

At lower frequencies (below 15MHz) all inductors can be toroidal to reduce their physical sizes. The toroid material permeabilities (μ_r) should be 10 or less, limiting selection to powdered iron since the minimum μ_r of ferrites is around 30.

Relays can be relied on for filter switching at hf. (**Figs. 5**, **6** and 7) but they should be of a low contact inductance type such as *Omron G2R* or equivalent manufactured by Magnecraft and several other companies.

At higher frequencies, where excessive parasitic inductances become increasingly critical, pin diode switches are probably the only choice for a designer.

Calculation of the rf voltages and currents in each element tell us the wire size required for the inductors and what the ratings for the capacitors should be.

For certain elements such as L_1, C_1 and C_5 (Fig. 4), manual calculation would be fairly straightforward. For the remainder, it would be complex, but programs now exist for the pc and Macintosh which optimise existing designs and plot the S_{11} and S_{21}, etc.

Other programs handle a more comprehensive design and calculate the nodal currents and voltages as well.

Skin effect

In the design example, L_2 and C_4 are exposed to the highest currents and C_3 is subjected to the highest rf voltage.

At a power level of 2000W, for example, the numbers are 11.7, 11.4A and 556V respectively. Since the skin-effect makes the current carrying capability of copper wire under rf conditions frequency-dependent, use as heavy gauge wire as practically possible to increase the conductor's surface area.

In addition to heating effects, conductor size affects the inductor Q values.

Skin depth (d) of a copper conductor is approximately 0.009mm at 100MHz. Based on this reference, for practical purposes, the skin depth versus frequency is:

$$d = 0.009 \times \sqrt{(100/f)}$$

where f is the actual frequency in MHz and 100 is the reference frequency in MHz. The resulting final numbers for 1.6 and 30MHz are 0.071mm and 0.016mm respectively. Usually five to six skin depths are considered adequate for good engineering practice, although rf current carrying capabilities diminish in the deeper layers.

Prompted by these numbers, experiments were conducted with inductor wire sizes of AWG #16 and #14, and no significant differences in performance or operation of the filter were noticed.

But for rf voltages in high power filter elements, the wire wound on ferrite or powdered iron toroids should be covered with some type of high temperature sleeving such as ptfe to prevent possible arcing to the core. High voltage, inexpensive disc capacitors suitable for high power filter applications up to 30-40MHz are available from several manufacturers.

The filter described here has been tested and operated, at length, a power level of 2kW. Common pc board construction techniques were used, with a ground plane on the component side, continuous except for clearances provided for feed-throughs (Fig. 6). The pc board material is *G10* epoxy-fibreglass with a dielectric thickness of 1.6mm and 2oz copper on each side.

There were questions in the early design stages, over whether *G10* dielectric material would have excessive losses at 30MHz, but this has not proven to be true.

At vhf (up to 200MHz) power levels of 2kW would be difficult to obtain without use

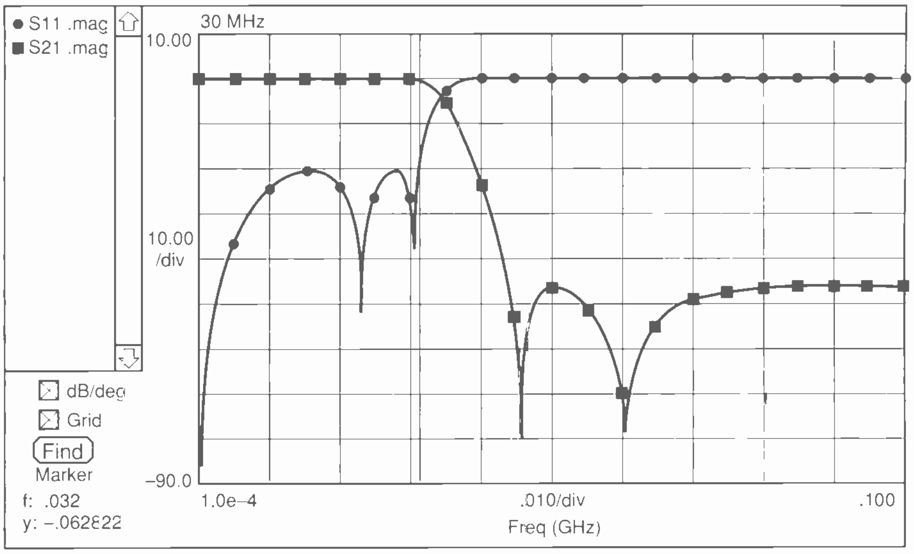


Fig. 7. Computer analysis of the response of the 30MHz filter shown in Figs. 4, 5 and 6 with the component values given in the text. The marker is set at 32MHz and the response is shown up to 100MHz. S21 is the actual filter response versus frequency and S11 is the return loss in dB.

of high quality components. Toroidal inductors as used in a hf design may be impractical and multi-layer ceramic capacitors such as ATC type *100E* or Tansitor type *MPH* are recommended.

At uhf, the high power filter design requires a completely different mechanical concept, such as etching or depositing the inductors and capacitors on a low loss substrate. Leadless chip capacitors would also be required to provide any degree of repeatability.

A single unit can always be 'tweaked in' to meet the specifications. But there may be a problem if the item has to be mass produced – especially with the inductors, which must definitely be air-wound or of stripline construction. As the frequency increases, the power handling capabilities of the components decrease exponentially due to higher losses and shallower skin depth of the conductors.

Filter switching with relays, feasible at hf with certain types, is not possible at higher frequencies because of the high series inductances of the relay contacts.

In rf power applications at vhf and uhf, pin diode switches are about the only way to switch signals. But at higher power levels the number of diodes required per filter makes this technique costly. ■

Designing antennas for wireless networks

Wireless local-area networks generally use lower microwave and upper uhf, typically between 800MHz and 3GHz and often at 2.5GHz. Antennas for such systems are the subject of application note *AN143-1.3* from GEC Plessey, which provides both an explanation of the various types of antenna and two worked examples.

Antennas for wireless lans are installed on the equipment enclosure which is either rectangular or cylindrical. They are usually around one or two wavelengths in size. Since the enclosure is in the near field of the antenna it significantly affects performance. It is best to think of both the antenna and the enclosure as a composite radiating structure.

Gain is usually defined relative to an idealised isotropic antenna; an isotropic element would produce a gain of 0dBi in all directions. Higher gain in one direction can only be achieved at the expense of reduced gain in another direction. The isotropic antenna is however unrealisable.

Usually, an electrically small antenna produces a dipole-type polar diagram with nulls in two directions and omnidirectional in the plane normal to the null axis. In wireless-lan applications, it is desirable to have these nulls vertically upward and downwards, and the omnidirectional plane horizontal. Vertical monopoles or dipoles, or normal-mode helixes all have the appropriate characteristics.

Radiated energy from an antenna is polarised, usually either linearly or circularly. In linear polarisation the electric field vector lies in a defined direction normal to the direction of propagation. Circular polarisation involves an electric vector rotating around the propagation direction so as to lie always in the plane normal to that direction.

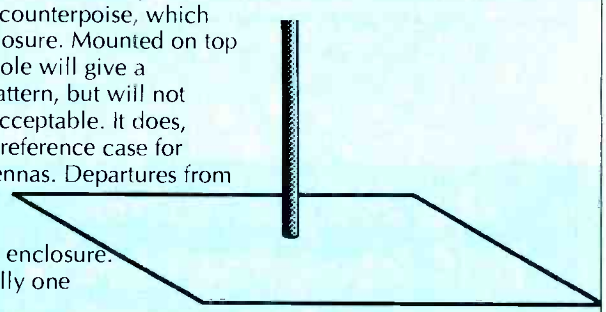
Most efficient power transfer is obtained when transmit and receive antenna polarisations are matched. In the extreme case of orthogonal polarisation, very little coupling, i.e. around –8dB, takes place between antennas. As a result, a vertical dipole and a horizontal dipole will not couple, while two vertical dipoles will.

Linear polarisation generally leads to simpler antennas; circular polarisation is produced by helixes or by phased feeding of a pair of linear antennas. This isolation between orthogonal antennas can be used to advantage in multiple access systems.

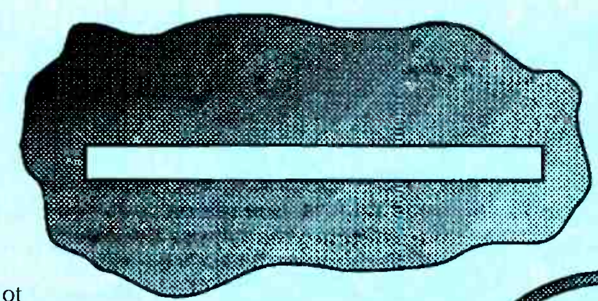
In order to maximise power transfer efficiency, the antenna must present a reasonable match to the transmitter/receiver. A poor match can also cause excessive reflected power in transmit mode which is returned to the transmitter with unfortunate consequences for the output devices. Low voltage-standing-wave ratio is relatively easy to achieve in wireless lans due to their narrow bandwidth.

Bandwidth should be qualified by defining

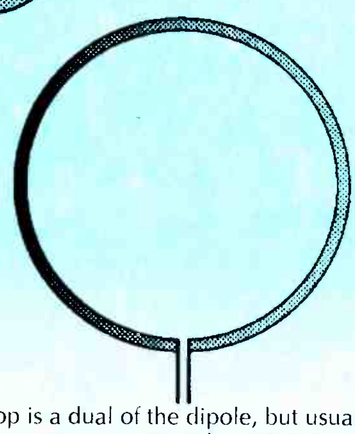
Dipole
Dipoles are simple antennas with well-known characteristics. If there are particular space constraints, they can be configured in free form with modification to the radiation pattern. The dipole is linearly polarised and typically a half wavelength long. The best location is on top of equipment such as a cellular phone or walkie-talkie. It is possible to print the dipoles onto a dielectric substrate for internal mounting but performance is reduced.



Monopole
Closely related to the dipole is the monopole. It requires a ground plane or counterpoise, which may be the equipment enclosure. Mounted on top of the equipment, a monopole will give a relatively clean radiation pattern, but will not generally be aesthetically acceptable. It does, however, provide a simple reference case for comparison with other antennas. Departures from the ideal monopole pattern are readily identified and attributed to the equipment enclosure. A monopole will be typically one quarter wavelength tall.



Slot
Slot antennas are closely related to the dipole, and are very suited to integration into an equipment enclosure. However they need a ground plane, and often a backing cavity. This can lead to the antenna being relatively narrow band. The antenna can be driven from an unbalanced feed across the slot, the feed point being chosen to give an acceptable input impedance.



Loop
The loop is a dual of the dipole, but usually requires more space. Its polarisation is complementary to that of a dipole.

the parameter being considered. An antenna may produce acceptable gain over a particular bandwidth, but its VSWR may only be acceptable over a much smaller frequency range. It is common however to talk of bandwidth as the frequency range over which all parameters of interest meet their specified values.

Bandwidth is often quoted as a percentage, and is obtained from the ratio of the highest frequency to the lowest. Bandwidths of 5% to 10% are common; an antenna with a bandwidth over 20% would be considered broadband while one with less than 2% would be narrow band. Special antenna designs are available with theoretically unlimited bandwidth, but they are physically large.

Enclosure interaction

Interaction between an antenna and objects near to it is particularly marked when the object is a few wavelengths in size. If the body is very small, say less than one wavelength, the total radiation pattern is usually dipole in form.

When the body is very large, say more than ten wavelengths, the antenna can usually be designed to be independent of the body by providing a ground plane. Between these limits, the body is not sufficiently large for a suitable ground plane to be incorporated, nor is it so small that the dipole pattern is obtained.

Interaction of the antenna with its surroundings needs to be considered. Scattering from nearby objects such as the equipment operator and desk top will affect the radiation pattern. Anechoic chambers should be used for antenna characterisation since they allow evaluation in a scattering-free environment.

EMC

Since an antenna is designed to provide electromagnetic coupling between free space and the interior of an equipment enclosure, there are EMC implications. As far as possible, the rf 'window' represented by the antenna should exclude out-of-band frequencies.

In addition, the antenna structure must not penetrate the EMC screening surrounding the equipment. These constraints can present a challenge to the antenna designer, who can only achieve acceptable antenna performance if the antenna is adequately separated from conductors and ground planes.

Multipath

Wireless lan antennas operate in a high multipath environment. This is combated to a degree by diversity working using two antennas spaced a quarter wavelength apart. Reflections from vertical and horizontal surfaces tend to preserve linear polarisation, while changing the handedness of polarisation. Linear polarisation is therefore preferable.

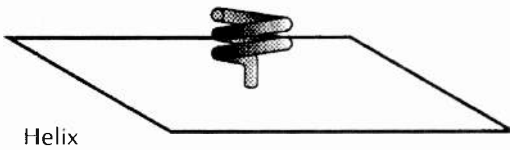
Simple sleeve antenna

A sleeve antenna can be made using tubing for one arm of a half wave dipole. Coaxial cable is passed through the tubing, and the centre conductor of the coaxial cable forms the other quarter wave arm of the dipole.

Screening of the coaxial cable is trimmed back and connected to the top of the tubing. For best performance, lengths of the tubing and the projecting section of the inner conductor should be adjusted using a network analyser for minimum VSWR at 2.45GHz.

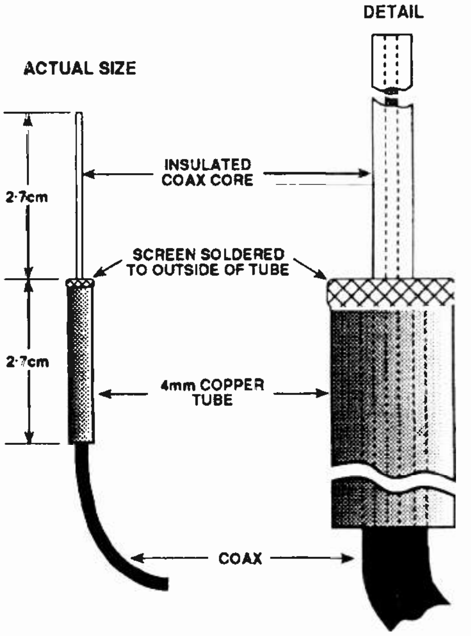
Tubing diameter is 4mm and the coaxial cable is 2mm outside diameter. An overall covering of heat-shrink tubing could be used to protect the antenna, with adjustments to the length of the dipole elements to allow for the consequent change in electrical length.

GEC Plessey Semiconductors, Cheney Manor, Swindon, Wiltshire SN2 2QW. Tel. 0793 518000, fax 518411.



Helix

The helix antenna can be very compact. In general, helixes have several radiating modes. These include axial mode, in which the helix radiates a circularly polarised wave along its axis, and normal mode, which is of particular interest in wireless lan applications. In normal mode, the helix radiates a linearly polarised wave with a maximum in the plane normal to the helix axis. The radiation pattern is identical to that of a monopole, but the helix is more compact than the corresponding monopole.



A simple dipole can be made using coaxial cable core for one pole and tubing slid over the cable insulation for the other.

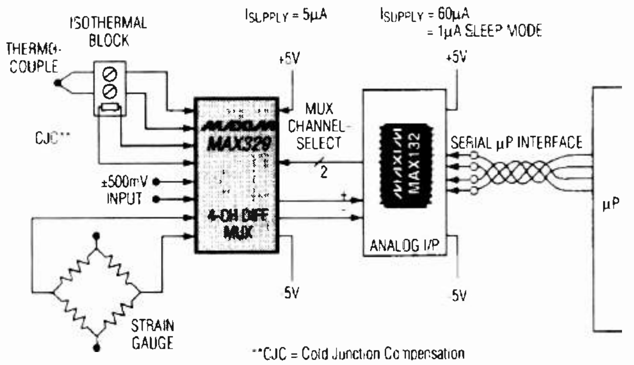
±18bit a-to-d conversion at 60µA

Resolving ±500mV inputs to 2µV, the *MAX312* consumes only 60µA while operating or 1µA in sleep mode. No intermediate signal conditioning is needed; thermocouples, strain gauges, etc, plug directly into the analogue switch and are routed straight to the converter.

Serial interfacing reduces board space. It also simplifies remote data acquisition and makes isolation easier. To expand the number of channels, the ADC's four digital outputs can directly control a multiplexer or programmable gate array. This eliminates glue logic and additional serial control lines.

Inherent 50/60Hz rejection makes the converter suitable for industrial environments. Only brief details of how to apply the device are contained in Maxim's fifth *Analog Design Guide* but an evaluation kit is available.

Maxim Integrated Products, 20b Horseshoe Park, Pangbourne, Reading, RG8 7JW. Tel. 0734 845255, fax

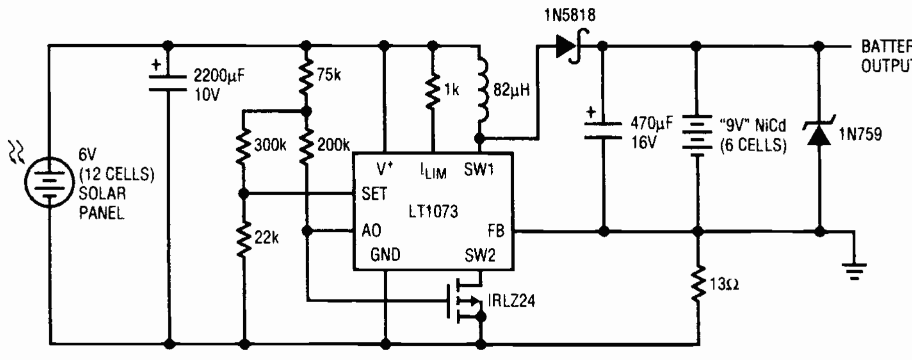


This high-resolution, low-noise analogue-to-digital converter has a 10pA input figure and interfaces directly to low-level output devices such as thermocouples and strain gauges.

Solar-powered battery charger

Remote instrumenation such as used for weather analysis often relies on solar power with battery back-up. Equipment being driven can consume considerable power in bursts while measurements are being made but very little averaged over relatively long standby periods.

Nickel-cadmium cells are ideal for this type of application provided that charging circuitry prevents overcharging and contributes negligible current leakage during hours of darkness. This circuit, from *Linear Technology Chronicle* Vol. 2, No 2, provides constant charge current over a wide range of light levels from a small solar panel. When solar-panel output falls to zero, the circuit automatically disconnects from the battery.



Unattended data gathering instruments often rely on solar power with battery back-up. Charging circuits require two features – low leakage in hours of darkness and overcharging protection.

Power for the *LT1073*, configured as a boost converter, is provided by a 6V solar cell array. Current is monitored via the 13Ω resistor, which keeps charging constant at 16mA.

The IC has a low-battery detector which, together with the mosfet, unloads the solar panel until there is enough light to develop 5V. At this point, the chip turns on and remains on until panel output falls to 4V.

Removing the battery can causes excessive output voltage so a 12V clamping zener is added. Note that the '9V' NiCd battery comprises six cells and actually provides about 7.2V.

***Linear Technology**, Coliseum Business Centre, Riverside way, Camberley, Surrey GU15 3YL, Tel. 0276 677676, fax 64851.*

Fax machine design

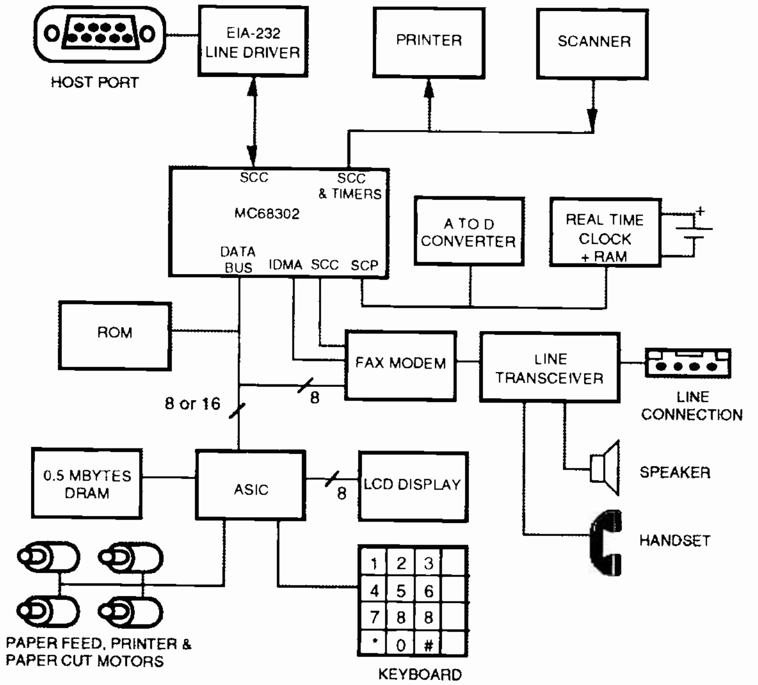
Circuit-function outlines for a high-performance fax machine are presented in Motorola's *DC411* application note entitled *An MC68302-based fax machine*.

Groups 2 and 3 are accommodated, while group 4 is an expansion capability. Upgrading to group 4 involves using the *MC68302* built-in ISDN channels and ISDN interface chips to replace the group-3 modem chip. Maximum communication speed is 14,400bit/s. The design supports error correction and integrated voice/fax facilties.

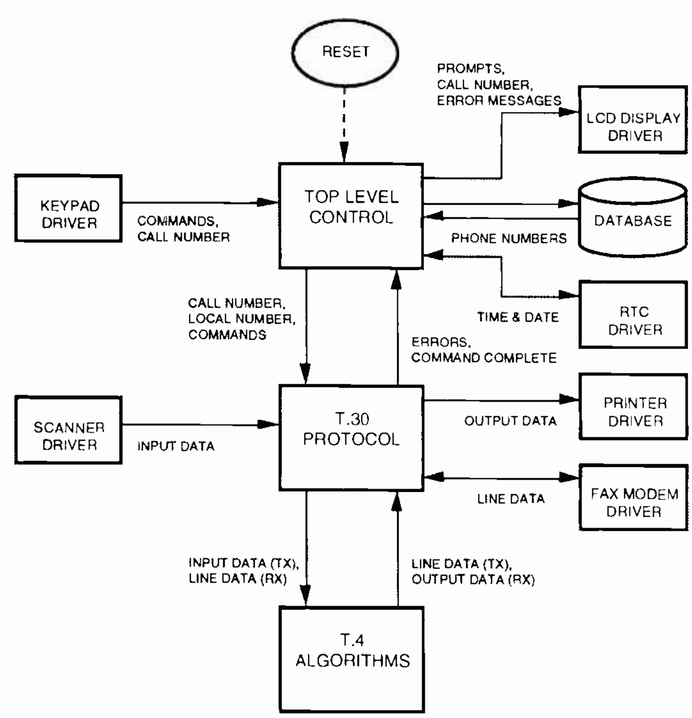
Although the note covers both hardware and software, it is intended as a design concept only. The functional block diagram and software chart shown here are the only illustrations presented. Since the note mentions specific devices for the various functions however, all that remains for the designer to do is collect the various data sheets, interface the chips and write the software - well almost.

***Motorola, European Literature Centre**, 88 Tanners Drive, Blakelands, Milton Keynes MK14 5BP. Tel. 0908 614614.*

Elements of a high-performance fax machine for groups 2 and 3. Communication at up to 14,400bit/s is possible.



Using integrating components such as the MC68302 at the heart of a fax machine reduces software complexity.



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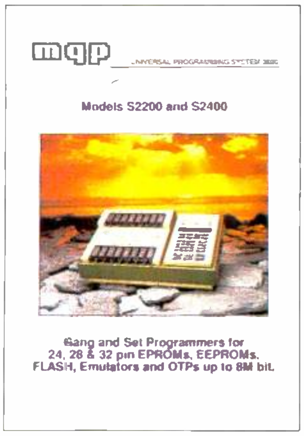

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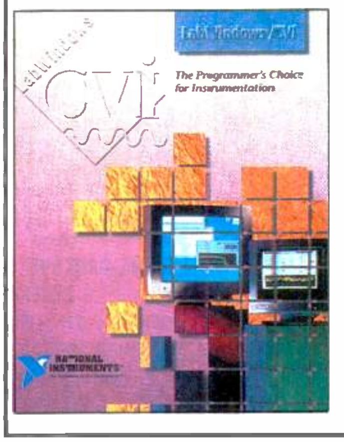

The LabWindows/CVI Brochure, from National Instruments, explains how users can build instrumentation applications on Windows PCs and Sun SPARCstations using the ANSI C programming language and LabWindows/CVI data acquisition, analysis, and presentation libraries.

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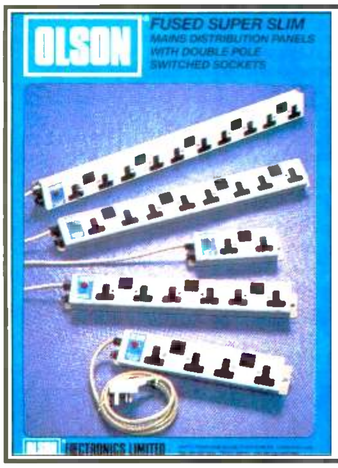

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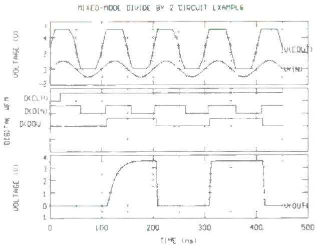

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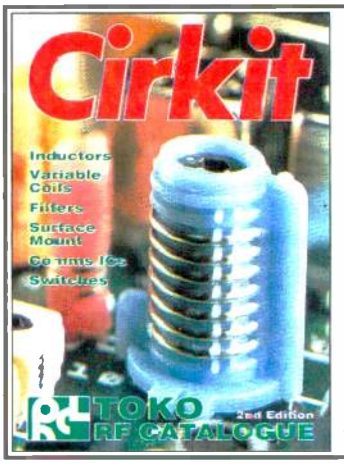

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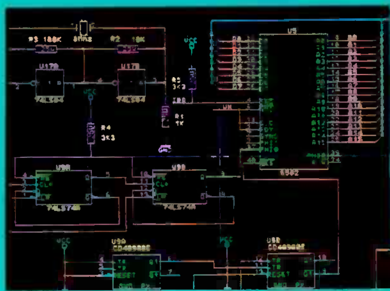

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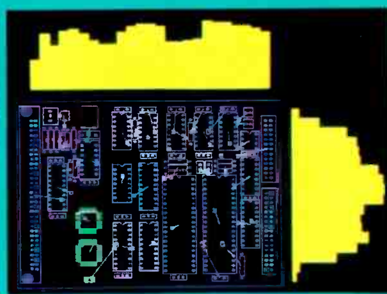

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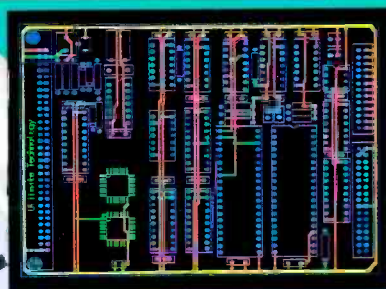

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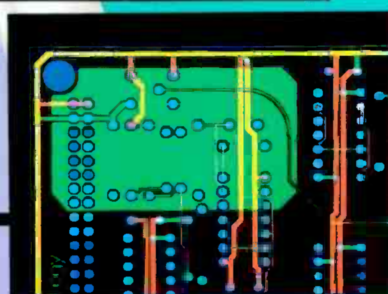

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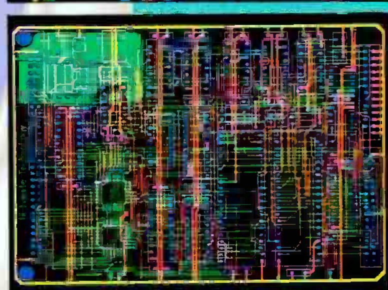

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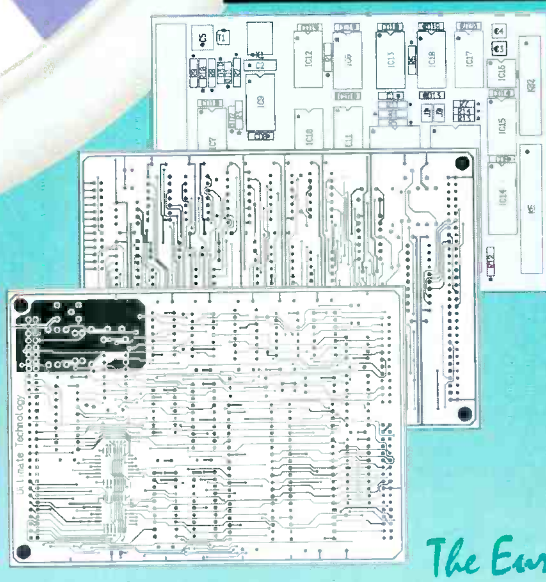

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