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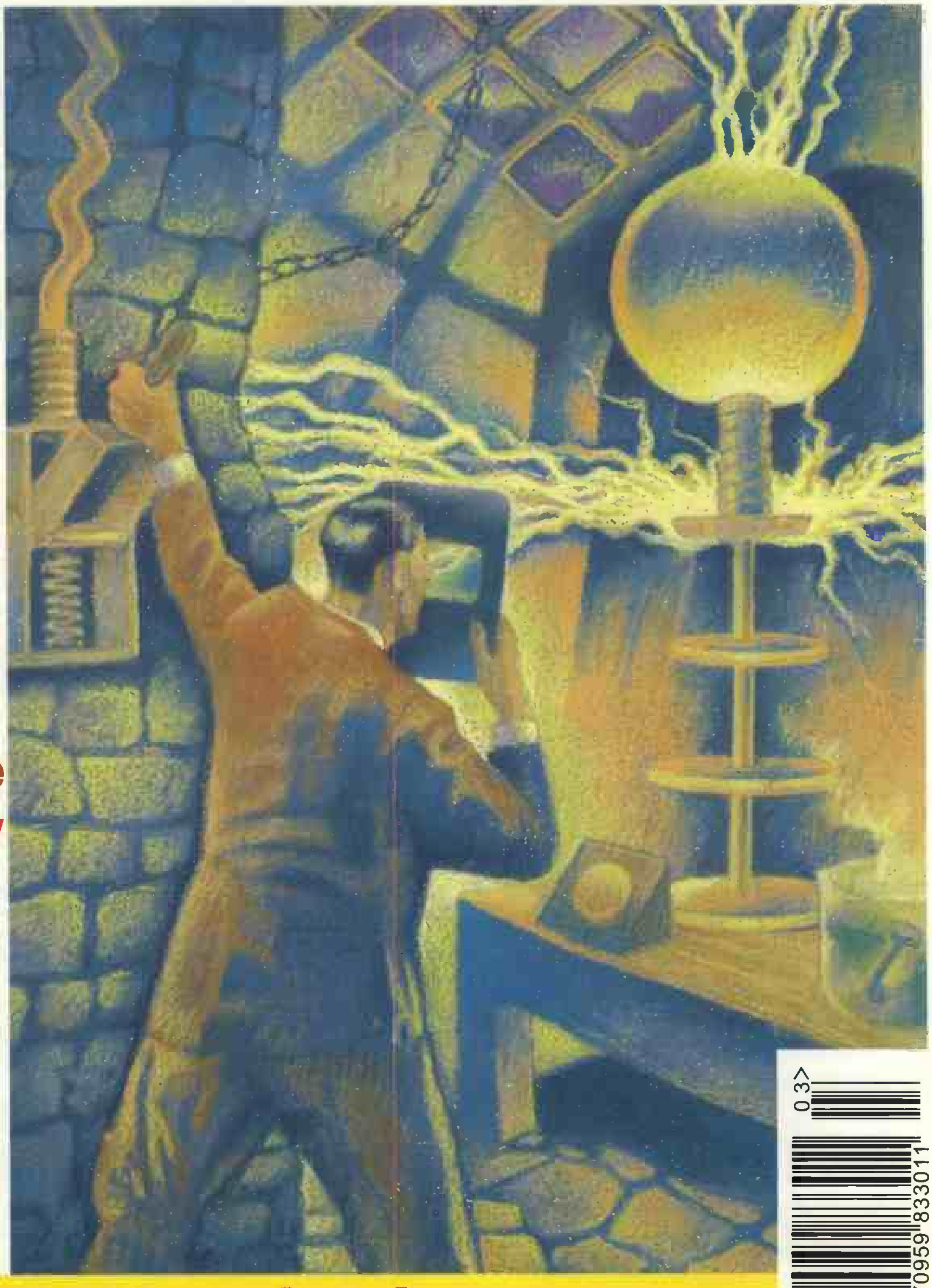


Distortion from power-amp supplies

Winning power switching circuits

Enhance RS232

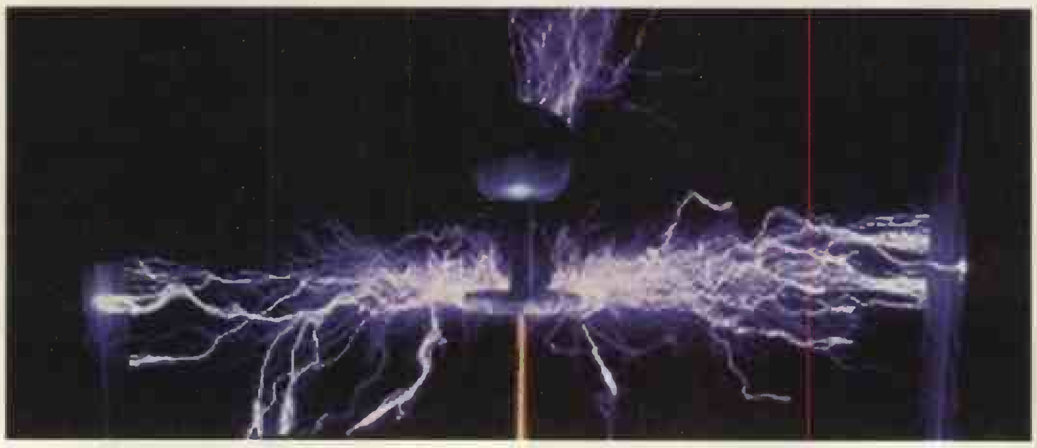
Transmission lines explained

Maths software for PCs - review

Filters adapt to improve performance

Unusual filters and oscillators

LIKE LIGHTNING? Tesla's ht generator



Very high voltages are generated by this modern version of Tesla's coil – page 190.

Cover Illustration Hashim Akib

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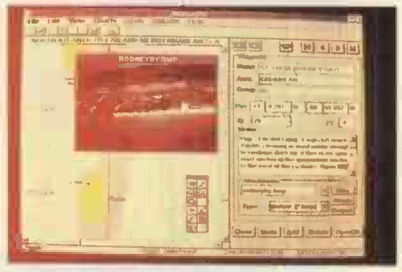
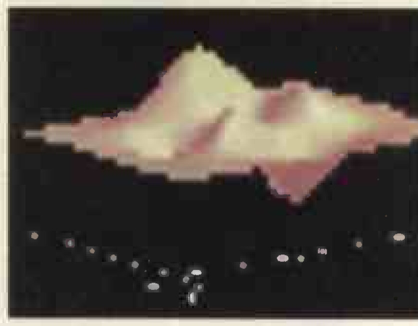
REGULARS



Multimedia hits GPS navigation – page 180.

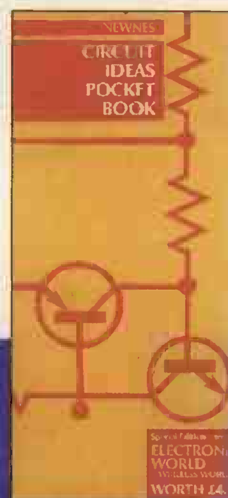


Matlab for powerful numeric analysis and graphic display. Reviewed on page 197.

Next month:

Free Circuit Ideas Pocket Book. Part 2 – another selection of circuit ideas brought together in one handy volume worth £4.95.
Plus Linsley Hood's audio attenuator, ISDN explained, 50Hz cancelling instrument front end.

APRIL ISSUE – ON SALE 30 March



Training dinosaurs

Why is so much emphasis placed on training? For engineers, isn't the first 25 years or so training enough? If we spend even more time training we will have retired before we start work.

The situation is further exacerbated by the fact that the effective retiring age appears to be decreasing all the time. Sixty-five used to be the retiring age but now more and more companies are 'letting go' their executives of 55 years old, or even earlier.

We must ask why? Why are older engineers – the people with the most experience – being asked, or asking themselves in some cases, to be let go? Why is industry losing this valuable asset of experience so early?

I believe that the answer lies within the engineers themselves. For far too long engineers have been designing, manufacturing and implementing hardware and software solutions. Every time a problem rears its head, the engineering fraternity rushes in to 'solve' it in its own inimitable way, focusing on technology.

I liken this to car manufacturers of the past. Cars then were bigger and went faster and further than ever before. That was acceptable in the sixties and seventies, but in many countries today, legislation has – at last – curbed the freedom of the automobile designer.

Cars must now be safer, more efficient and cleaner. At last the broader issues have been addressed.

Will it take a similar act of Parliament to force today's electrical engineers to consider the broader issues? You could say that it is all part of the 'green dream', but green in the sense of growing and continual rejuvenation. 'Technology has gone mad', is a phrase often heard, but I would like to rephrase that and say many of today's engineers are not thinking long term.

Remember the car designers who were forced to design cleaner cars. The long term view is now an issue and I believe is relevant to hardware and software solutions. It is no good designing 'solution generators' because people – the equipment operators – eventually lose the ability to think clearly about the problem being tackled by the system they are operating.

Why should they? After all, electronics gives them the answer doesn't it?

I am not suggesting that we abandon the technological approach. Far from it. But we must move away from the technocentric approach. Technology has to go hand in hand with people. The broader issues must be assessed and we need to become even more innovative in the use of application technology. We must develop a human-centred approach to the problem.

The thought of going back to 'school' to learn about some aspect of work is a necessary luxury. But knowledge should not be gleaned solely in this manner. If systems were designed on human-centred lines, people would be learning while they were working. This learning/earning approach must be adopted by companies and designers alike.

If we are to reverse the trend towards oblivion – maybe even at 50 – then electronics engineers must stop designing solution generators and start designing 'suggestion makers', decision support systems and tools. The distinction is difficult to make – but not impossible.

Companies need to start supporting the learning process of their staff, not by spending even more time away on training courses, but by adopting human-centred systems in their every day working life.

Organisations that adopt this approach will become learning/earning organisations. They may find it harder in the short term. But on the longer time-scale, the staff within that company will find that their knowledge is continually rejuvenated.

Not only will we see companies with a future, but we engineers will have one too.

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UPDATE

Video cd war averted?

The high-density video cd of the future looks like being the double-sided disk developed by Toshiba and Time Warner. Its victory appears assured following the decision of Matsushita, the world's largest electronics group, to support it rather than the rival Philips/Sony system.

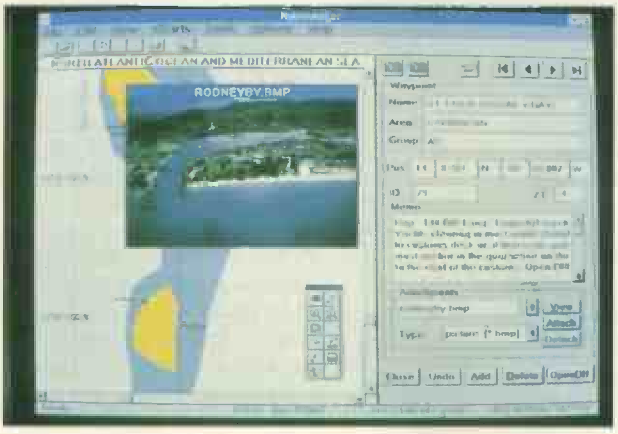
Thomson, JVC, Pioneer and Hitachi, plus several Hollywood studios, have also shown support for the new video disk standard, DVD, which can carry up to 142 minutes per side of MPEG2-quality video. It provides dual-format display in 4:3 or 16:9 aspect ratios, and multi-language sound tracks. As a cd-rom, it could carry eight times as much data on existing disks.

Philips and Sony immediately issued a statement supporting the advantages of a single format and went into top-level talks with Toshiba, aimed at contributing elements of their own system, or at least securing financial recompense for abandoning it. Their chief threat, however, is one of ruining the market's chances with a format war.

This may well happen anyway. Two-disk movies are already being released, using the existing low-density video cd standard and capable of being played on various pieces of hardware, including cd-based consoles such as CDi and 3DO, and some cd-rom drives.

Reported by **Peter Willis**

GPS receiver breakthrough enables sail-by-wire



Multimedia navigation charts show detail such as photos of landmarks.

Tall ships, and short ones, these days need more than a star to steer by, and this year's Boat Show was as ever full of additional assistance for addicts of sailing-by-wire.

The most significant breakthrough was in Differential GPS – the use of land-based radio beacons to correct the detuning deliberately introduced by the US military, and provide mariners with positions accurate to 10m. This enables, for instance, Baltic sea ferries to navigate narrow channels without reducing speed.

In British and Irish waters the service, now about two years old, is provided by the General Lighthouse Authority, under the name Scorpio, and at a hefty price – a £500 a year subscription. The breakthrough was in the form of a new DGPS receiver from Trimble Navigation, compatible with *Scorpio* and other systems, and with a five-year *Scorpio* fee built into the price at a premium of only £300. At £2895, the *NT200D* will at present appeal chiefly to commercial users and 'must-have' offshore racers.

GPS lets you know where you are, not where everyone else is. For that, you still need radar, and new lcd systems, from Autohelm and Raytheon, should encourage its use on smaller boats. Easier to read, and with convenience features such as picture-freeze, zoom and off-centre displays, they can also be overlaid with some electronic charts.

Cd-rom charts, fully coloured and using pc facilities to the full for vectored scanning to show varying levels of detail as required, looked more impressive than ever this year. PC Maritime was showing 'multimedia' charts – where photos of buoys, landmarks etc, or even videos and commentaries can be superimposed on a chart of the relevant area.

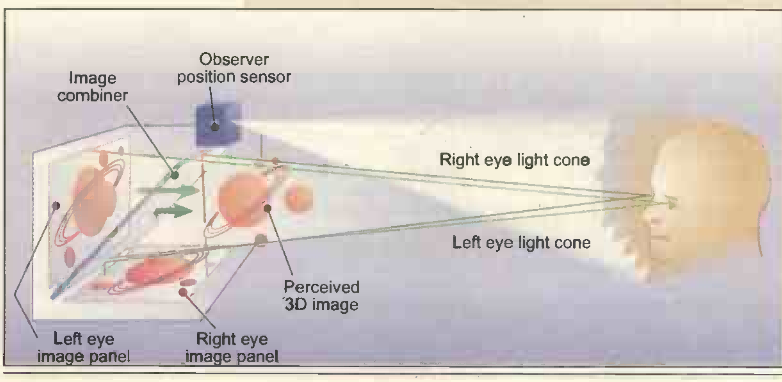
Perhaps the most practical contribution to sailing-by-wire though came from Cetrek, with its combined chart plotter and autopilot. At last, the seafarer can leave his boat to both tell itself what to do and get on with it while he settles down in his bunk with a Scotch and a copy of *The Riddle of the Sands*.

P. W.

3D tv advances

Sharp's 3D system follows the viewer's head.

British engineers at Sharp's research labs in Oxford have developed a 3D display system which could form the basis of commercial 3D-television.



The system uses sophisticated optics and image manipulation techniques to deliver 3D moving pictures, which the viewer can look around, from just two stereoscopic images.

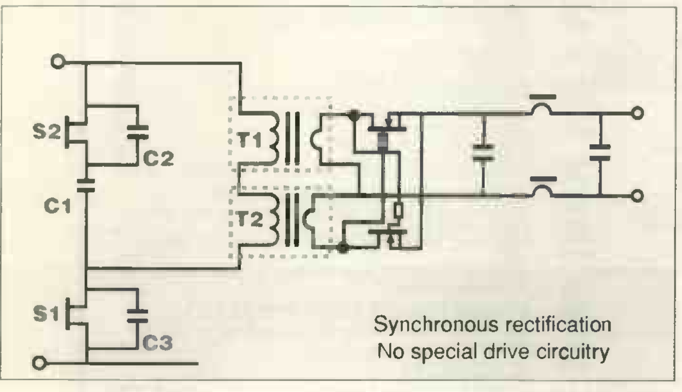
Sharp's 3D development team said the technology could be used to produce 3D tv pictures using images from just three or four tv cameras.

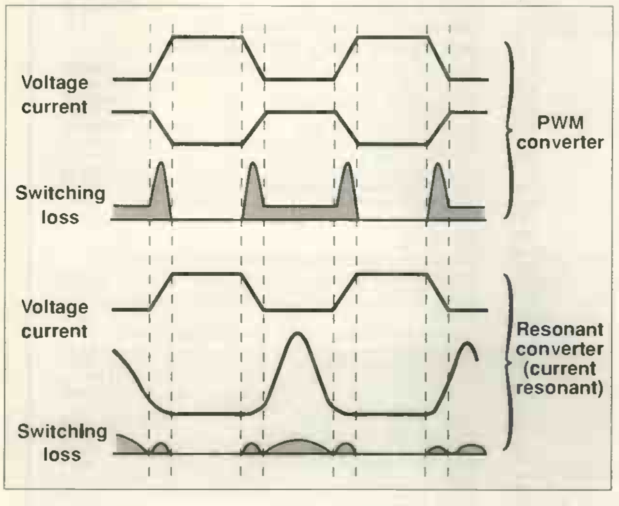
Most existing full-3D systems need to display a large number of images, one for each allowed viewing angle. Sharp gets away with just two by monitoring the position of the viewer's head and altering the two images to present the appropriate view to each eye.

At present the display can only produce 'look around' pictures for graphics sequences, but the Oxford team is currently working on video images.

Emerging dc converter technology promises higher efficiency and lower prices

Third generation dc-to-dc converters marry the quasi-resonant switching techniques of second generation types with synchronous output rectification to give 90 per cent efficiency with low output voltages for the first time.





A synchronous rectification technique is coming to the fore in the latest generation of dc-to-dc converters as manufacturers strive to achieve efficiencies of 90 per cent or more for a 5V output.

Over the past five years, the technology of dc-to-dc converters has developed from the traditional hard, on-off fast switching pulse width modulation, which gave efficiencies of 75 to 80 per cent.

These first generation resonant converters improved efficiency above 80 per cent, but suffered the disadvantage of comparatively high stresses on the switching components and variable frequency.

High stresses on the switching devices tended to require more expensive devices, thereby increasing converter costs. Variable frequency increased the cost of using the converter because of the increased complexity of system filtering.

Second generation 'quasi-resonant' converters resulted in the efficiency improvements of fully resonant converters, while reducing switching device stresses and giving constant-frequency operation.

The third generation of dc-to-dc converters marries the quasi-resonant switching techniques of the second generation converters with synchronous output rectification to give 90 per cent efficiency with low output voltages for the first time. The RM series from Coutant Lambda is one of the first commercial products to employ this combination.

In operation, primary side switches S_1 and S_2 operate in anti-phase such that when S_1 is on, T_1 acts as a transformer while T_2 acts as an output choke referenced to the secondary side by T_1. When S_2 is on, S_1 is off, and T_2 acts as a transformer while T_1 acts as a choke. Because the output choke is on the primary side, it is a smaller and cheaper component.

Also, the circuit uses two identical complex magnetics, the thereby achieving economies in manufacturing. The planar transformer has a particularly low profile for surface mount applications.

During the switching transition, current is carried without loss via capacitors C_2 and C_3. Switches S_1 and S_2 only turn on when the voltage crossing them is zero and so the technology gives zero voltage switching.

On the secondary side of the converter, the output rectifier diodes have been replaced by fets in positions R_1 and R_2.

Diode voltage drop ultimately limits any improvement in the efficiency of the converter. Field-effect transistors on the other hand behave like resistors, so that efficiency can be improved simply by putting more fets in parallel. As a result, the new technique improves efficiency by reducing both primary side losses and secondary side losses to give performance unparalelled in commercial converters to date.

Relatively little switching loss occurs in the case of synchronous rectification. In the first generation resonant converters, power is translated into heat. However, peak current in the 'on' state of the switch is generally much higher.

Overall efficiency of the converter is improved at the expense of a higher current stress in switching. But with zero voltage switching there is no on-state stress.

The overall result is a smaller system size, because no heatsink is necessary for the converter: high density converters often require a heatsink larger than the converter itself. Other benefits include a substantial reduction in cooling requirements, low cost of use and improved reliability.

Paul Gregg, *Electronics Weekly*

Linking research and industry

The Government intends expanding its scheme aimed at promoting collaboration between researchers and industry.

The Realising Our Potential Award scheme, ROPA, set up last year, is to be adopted by all six Research Councils, extending to cover all the-areas of science and engineering.

The scheme focuses some funding specifically on researchers already working closely with UK industry and commerce on basic or strategic research projects.

Defence research cuts halted

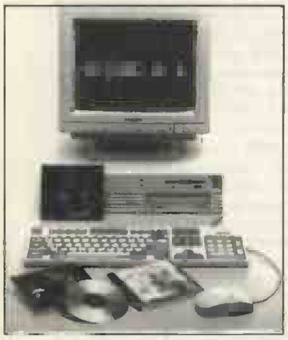
Government plans to reduce the Defence Research Agency's role in electronics component standardisation have been put on hold as a result of industry opposition.

Original plans were to save £6m from the DRA's budget by axing its electronics components watchdog, the PCS, with the loss of 70 jobs in March. Following intense lobbying from the defence industry before Christmas the MoD is reviewing its options. A source said that "more consideration is required". Budget cuts will still be necessary but the MoD has conceded that a 'smooth transistion' will be needed.

Defence companies are alarmed because the axing of the PCS and abandonment of the Defcon 17 component standardisation protocol would require them to implement, and bear the cost of self-auditing of components used in defence contracts.

The UK could also lose its official representation on international military standards committees.

CDi – the paradoxical future



A recent international conference explored several possible routes whereby CDi might increase its influence.

CDi celebrated its first million sales this year, worldwide. As an achievement, it is put into context by an installed consumer base of pcs between 10 and 15 million, and 6 million cd-rom drives.

It seems that CDi, the system developed by Philips to make interactivity and multimedia simple and accessible enough for the average consumer, is being outpaced on its home territory by more complex – albeit more versatile and capable – business machines.

However, a recent international conference explored several possible routes whereby CDi might increase its influence. These are indicated, in part, by success it is already enjoying, paradoxically, as a business tool. Its low cost and ease of operation are commending it to organisations as a one-per-branch staff training and information medium, and for 'kiosk' applications. Basically customer information points, these are extendable to enable customers to, for example, 'design' a car in their choice of exterior and interior colours, trim and accessories.

The next step is obviously to put this on the consumer's tv screen at home, via an interactive cable network. A home shopping trial is already under way in the Netherlands, with supermarket chain Albert Heijn. The disk carries product information, and magazine-type material such as menus, recipes and wine advice. For customers, an advantage of the CDi system over other forms of on-line shopping is a saving on phone bills - the order is assembled off-line and then squirted down the cable in a few seconds.

These capabilities may amaze many users. "It is important to hide the fact that the CDi player is a computer," advised one speaker. "The man in the street understands tv, but he doesn't understand computers." Potentially one of the biggest boosts for CDi could come from cd-video, but this is not a hazard-free opportunity. Many markets, Japan for one, are keen on movies-on-disk, but don't fully appreciate interactivity, and could develop a market based on straight, linear movie players. Yet such a box would have most of the capabilities of CDi built into it.

Favourable to CDi, oddly, is the promise of video-on-demand. Navigating those compressed digital channels – up to 400, according to some estimates, by year 2000 – will require a system of some sophistication. One of CDi's missions has always been to drag interactivity out from behind the computer screen in the spare bedroom, and put it in front of the tv screen, on the sofa in the living-room. Video-on-demand could be its big opportunity. Pc users who hate installing cd-rom can now cheat, by having a CDi playback board, bypassing the pc memory and disk.
P. W.

Colour differences increase fibre comms throughput

IBM is testing what it claims is the world's most advanced commercial fibre communication system based on wavelength division multiplexing, wdm, technology.

Known as *MuxMaster*, the system can squeeze up to 10 two-way channels, each of which can carry up to 622Mbit/s in each direction, simultaneously in a single optical fibre. IBM claims the system works over distances of 50km with rates up to 200Mbit/s per channel, although the range is shorter for the maximum 622Mbit/s rate.

Wave-division multiplexing, in which several different wavelengths, or colours, are used to send signals simultaneously down the same fibre, is not a new idea. Prototype wdm systems have been built in several labs, including BT's Martlesham research laboratory in the UK.

To date however, the technology has barely begun to move out of the lab and onto the market. Pirelli is one of the few companies to commercialise wdm technology, with a system that uses four different wavelengths, allowing two two-way channels to exist side by side in a fibre.

Now, IBM is conducting commercial trials of a system that uses 20 wavelengths, believed by them to be the first commercial system that uses so many wavelengths.

At each end of the fibre in the IBM system lies a diffraction grating and an array of 10 semiconductor lasers and 10 photodetectors. Each laser operates at a slightly different wavelength, in a band which sits in the 1550nm 'window' at which optical fibres are most transparent.

The lasers are positioned in front of the rating so that in each case the light produced by the laser is diffracted into the mouth of the optical fibre. Similarly, the photo-detectors are positioned where the grating will diffract light coming from the fibre at the wavelengths of the transmitting lasers .

The lasers are of the distributed feedback type, in which a grating etched into the laser cavity sets up an interference pattern. This pattern cancels out all but a small band of wavelengths, producing an output with an extremely narrow spectral width. They can switch at frequencies of up to 1GHz.

MuxMaster uses a spacing of 1nm between adjacent laser wavelengths – much wider than that theoretically needed to cope with the broadening produced by the 1GHz signal modulated onto each carrier. This works out at less than 0.01nm.

According to IBM, the extent to which the spacing between wavelengths can be reduced is limited in practice by the resolution of the grating, which determines how widely it separates adjacent wavelengths, and by instabilities which cause laser wavelength to drift.

Although very advanced for a commercial system, the technology used in *MuxMaster* is not leading edge in terms of what has been achieved in the laboratory – all the components used in *MuxMaster* were bought off-the-shelf. For example, BT, Italtel of Italy, and France Telecom have demonstrated a wdm system with a wavelength spacing of just 0.1nm. In principle, this would allow up to 50 channels to coexist in an optical fibre with erbium-doped fibre amplification.

IBM say *MuxMaster* could be upgraded to 0.1nm wavelength spacing by using a finer grating and advanced laser diodes with better stability. Data capacity would be multiplied by a factor of ten offering a staggering throughput of 125Gbit/s, but this is still far short of the potential 25Tbit/s, theoretically possible with fibre.

As well as boosting the capacity of a fibre, wdm can also be used to boost its flexibility. Each wavelength can. carry a different kind of traffic, with various bit rates and protocols. The 50km maximum range of *MuxMaster* is limited by dispersion in the fibre. The relatively short range achievable today is one of the key reasons why the technology has not been commercialised by the telecoms companies which have done most of the development work on wdm.

Telecoms firms focus on long-haul traffic, with a huge number of wavelengths over long distances, like between cities or different countries

Early applications identified for *MuxMaster* include disaster recovery, where a computer at a site hit by some calamity downloads all of its stored data to another location. Other potential uses are tipped as being medical imaging, distributed peripherals and the transfer of uncompressed video, for example between tv studios. ■

Karl Schneider, *Electronics Weekly*

Best rf article '95

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

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

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

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

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

Win a £4000 programmable signal generator from Hewlett-Packard

The prize for the coming year's award is a £4000 Hewlett-Packard HP8647A 1GHz programmable signal generator. It features HPIB interface, solid state programmable attenuator and built in AM-FM modulation capability.

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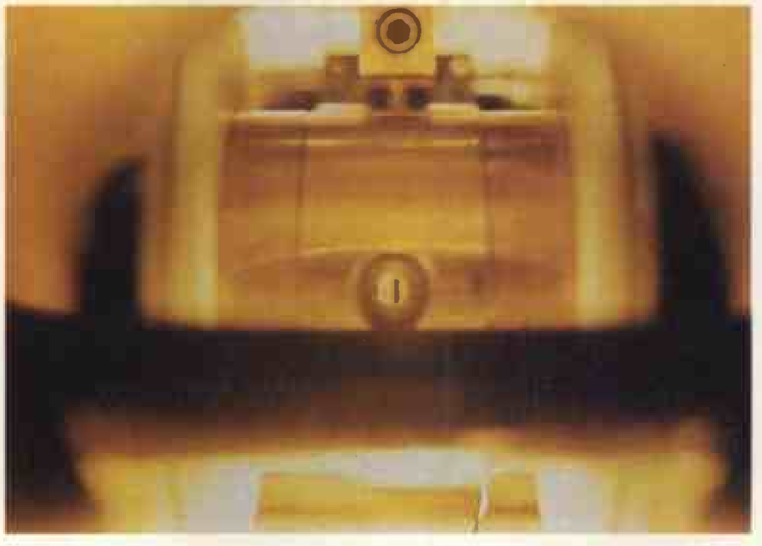
RESEARCH NOTES

Jonathan Campbell

Laser demonstrates singular success

Successful operation of a laser using a single, isolated atom is being hailed as a fundamental advance in laser physics and our understanding of how atomic systems radiate in an enclosed laser resonator. The long-sought development, by researchers at the Massachusetts Institute of Technology's George R Harrison Spectroscopy Laboratory, marks the first time that laser oscillation has been achieved with only one atom in the laser resonator.

Present lasers use amplifying media composed of billions of atoms or molecules to achieve the enormous photon multiplication required for coherent light generation. Such large numbers are required to overcome low emission efficiencies per atom and photon losses of existing laser resonators.



Close-up view of the cavity assembly of the MIT microlaser. The two bright stripes in the centre are the supercavity mirrors while the dark stripe, about 1mm wide, is the mirror spacing.

But because the atoms in a conventional laser interact with each other as they emit photons, fundamental features of the atom-photon interaction process are lost.

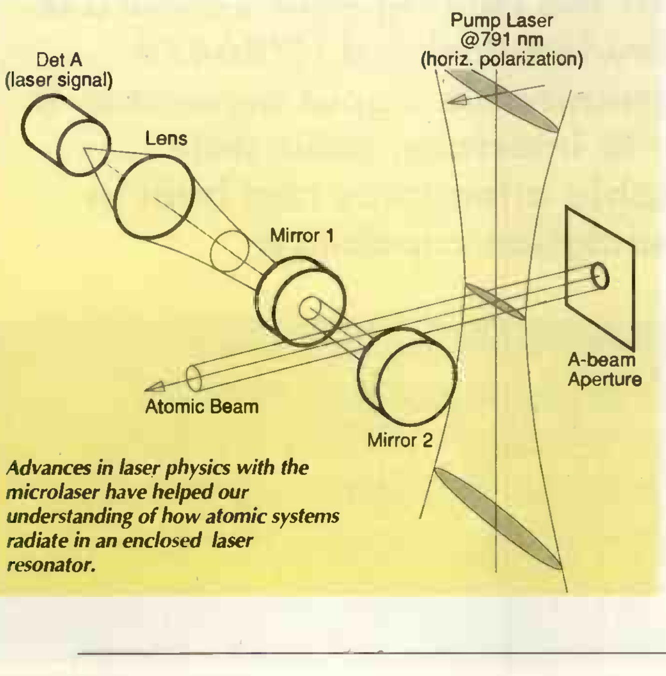
The keys to MIT's single-atom laser are a laser resonator with extremely efficient photon storage capability and selection of an atomic transition with appreciable strength and negligible spontaneous emission – atomic barium transition at a wavelength of 791nm.

The new laser device has been developed by a large team team at MIT led by Professor Michael Feld, director of the laboratory, and PhD researcher Kyungwon An, from Korea. It is composed of two supercavity mirrors – having a capability for storing photons 10,000 times greater than conventionally – separated by 1mm, with a beam of atomic barium flowing through the gap between the mirrors. The barium atoms are raised to an excited state before they enter the resonator, and the flux of atoms is kept small enough to ensure that one atom or fewer is inside the resonator at any moment.

Laser operation is initiated when a barium atom traversing the cavity emits a photon into the empty resonator. As successive atoms flow into the resonator, photons are emitted with greater and greater likelihood, leading to an equilibrium state with many photons inside the resonator. But a steady state is never reached, and even in this equilibrium state the system continues to undergo atom-photon oscillations. Laser light is coupled out of the resonator through the mirrors provided with a tiny amount of transmission.

With an average of one atom inside the resonator, the researchers measure an emission rate of more than 10 million laser photons per second, from which they estimate that about 11 photons are stored in the resonator.

Several technical obstacles had to be overcome to make the experiment work. A laser supercavity exhibits resonant behaviour over a very narrow spectral range, making it difficult to keep the laser cavity resonant with the atomic transition. To ensure atom-cavity resonance, an extremely stable resonator had to be developed to keep the mirror spacing fixed to less than a billionth of a centimetre. It was also necessary to ensure that the pump laser, which excites the barium atoms before they enter the resonator, was kept in resonance with the atomic transition. Finally, a high efficiency detection scheme employing an avalanche photodiode was used to achieve a counting efficiency of 36% – far superior to the efficiency of conventional photomultipliers at this wavelength.



Advances in laser physics with the microlaser have helped our understanding of how atomic systems radiate in an enclosed laser resonator.

Research could push fibre comms to 200Gbit/s

Optical fibre under development at Corning Inc in the US could rewrite the limits for communication system capacities. Results of initial tests carried out on the fibre at the University of Rochester suggest that Gbit/s transmission rates of might be possible, with the distance between amplifiers significantly increased.

In fibre optics there are two adverse and opposing effects that limit communication capacities: a group velocity dispersion which is always present and a non-linear self-phase modulation. In conventional systems, designers try to balance the two effects to keep the optical pulses intact. But unavoidable blurring of the pulses over long distances and at high data rates puts a ceiling on transmission capacities.

Companies such as AT&T and Nippon Telegraph have tried to reduce the blurring by using ultra short pulses from a soliton laser (also under development). As the pulse travels down the fibre, its intensity level modulates the fibre's index of refraction, allowing the pulse to propagate with no change in wave shape.

But even these ultra short soliton pulsess face dispersion problems. Their energy drops as they travel down fibres. This change in intensity affects their self-phase modulation in a non-liner manner. If their energy is periodically boosted, the solitons can became unstable.

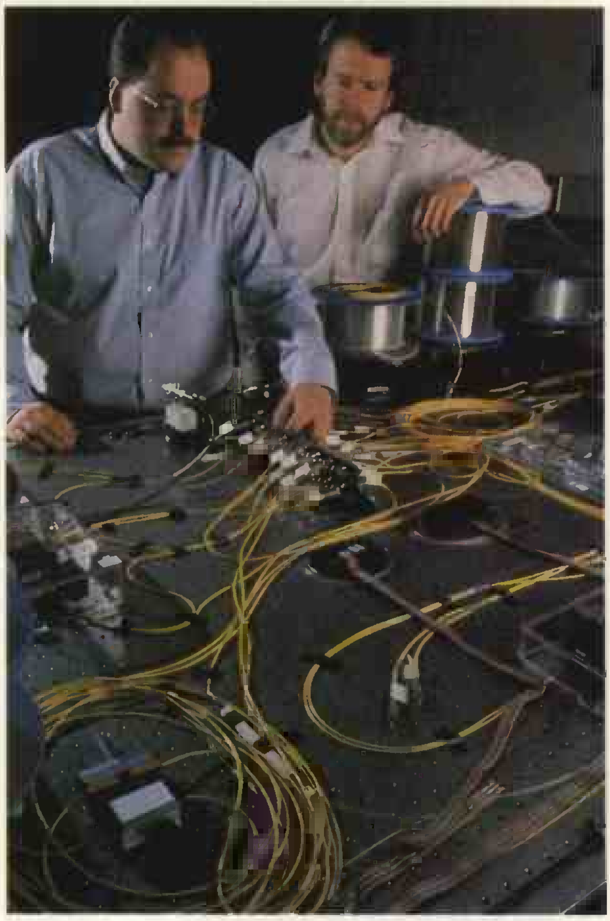
Corning's solution is to build a special dispersion-decreasing fibre to compensate for the energy fluctuations, matching the soliton's decreasing energy with a fibre whose dispersion decreases proportionally.

To test the fibre, Andrew Steintz at University of Rochester built a passively mode-locked soliton laser based on amplifiers doped with the rare element erbium. Using the laser he sent 1ps pulses through a 40km stretch of fibre spooled in the laboratory. The solitons emerged intact. But sending solitons from the same laser through conventional fibres resulted in signals being quickly degraded.

The Rochester result is the first step in assessing how the fibres could be used to improve soliton transmission. Next stage is to put in a full system.

Alan Evans, senior research scientist at Corning, says he expects that to show even more dramatic effects.

Information capacity of soliton communication system using the Corning dispersion-decreasing fibre looks to be almost unlimited. Researchers are already talking in terms of 200Gbits/s. But the preference may be to reduce the bit rate and instead opt for fewer amplifiers, more widely spaced, on a system. The result would be a useful boost to the economics of soliton communications.



Corning's new fibre undergoing tests at University of Rochester.

Lithography with magnetic allure

Storing individual bits of information in single-domain magnetic particles is an attractive goal for scientists looking to boost capacities of magnetic storage densities. Now a team at the Department of Electrical Engineering, Stanford University, looks to have brought realisation a step nearer by developing a procedure for high-resolution patterning of magnetic recording films.

Using direct-write electron beam lithography and a multistep sputter etching process, the researchers report that small islands of polycrystalline magnetic thin films have been successfully produced (*J Vac Sci Technology*, B, 12, (6), 1994, pp.3196-3201).

Magnetic recording medium normally used in hard disks is a thin unpatterned polycrystalline film of cobalt alloy with a grain size of 10-20nm. But the recorded bits, consisting of many magnetic grains, have rough edges, producing fluctuations in the average positions of the bit boundaries and so jitter during bit reading.

The possibility of a recording system based on a square array of circular tracks of lithographically-produced islands of magnetic material is generating intense interest, because signal to noise ratios would be substantially improved. Storage density of such a medium should be limited only by the lower size limit for single domain particles ($\approx$10nm).

Researchers R M H New, R F W Pease and R L White have high hopes for their technique because this noise is now one of the most important limiting factors in increasing bit density in hard disks. The Stanford patterning processes involves eight steps – more than any previously described – but the team says that virtually any type of magnetic thin film can be patterned as a result.

Patterning begins by depositing 10nm of carbon onto a previously deposited 20nm magnetic film eg cobalt. Silicon oxide is then sputtered on, followed by a 50nm polymer (pmma) layer which is used for the direct-write e-beam lithography. After a 10nm chromium mask is deposited in the areas exposed by the electron beam, previous layers are removed step by step using the chromium as a mask.

Sputter etching through the magnetic film then leaves the islands of magnetic material covered only by the silica which is finally removed. In a 20nm cobalt magnetic film, feature sizes down to 0.1μm have been produced. Tests show that the patterned magnetic islands are physically isolated from each other.

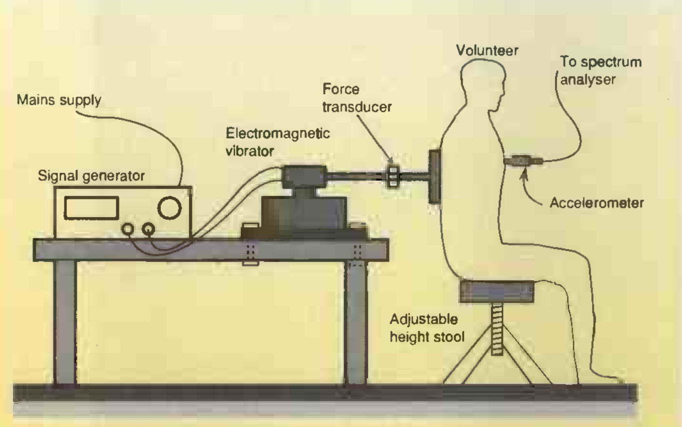
Commercial interest will be helped by the fact that the process can be used for films already used in conventional recording media.

Meta-physio-therapy?

Knowing our chest resonant frequencies could improve mechanical physiotherapy.

Back in 1925, TS Eliot observed "We are the hollow men". What a pity he didn't have the advantage of medical technology now being developed by MJ Goodwin of the School of Engineering, Staffordshire University. Using Goodwin's measurements of resonant frequencies in the human chest he could doubtless have said exactly how hollow.

Of course Goodwin's work is connected to a much more serious subject than mere poetical metaphysics. Physiotherapy of the chest to help dislodge mucus caused by some lung diseases is a burden for both patient and physiotherapist. Using mechanical devices to apply the physiotherapy has been tried. But little attempt has been made to tune the forcing frequency to the structural dynamic characteristics of the human chest. That is the goal at which Goodwin has been aiming (*Proc Instn Mech Engrs*, Vol 28, pp.83-89).



In his experiments, subjects were required to sit upright on a stool with their backs against a light, stiff pad. This pad was attached a electromagnetic vibrator driven by a signal generator whose frequency was adjusted manually.

A force transducer placed between the vibrator output rod and the pad recorded the magnitude of the exciting force delivered to the subject. An accelerometer, held firmly against the front of the chest, was used to record the resulting response of the body.

Outputs from both the force transducer and accelerometer were fed to a two-channel spectrum analyser to identify the amplitude and frequency content of both signals. A clear well-defined peak in the spectrum was observed coinciding with the forcing frequency.

Then, using a white noise forcing signal, Goodwin was able to identify a clear peak in the spectrum, enabling the volunteers' chest resonant frequencies to be identified. The results varied between 18.5 and 35.3Hz. This indicates that any mechanical physiotherapy would have to be highly patient specific.

Goodwin hopes eventually to be able to estimate resonant frequency based on a patient's size. One of the next steps would be to assess the efficiency of physiotherapy provided by a mechanical device properly tuned to allow the researchers to take full advantage of resonance.

The treatment certainly looks promising, as forcing close to the resonant frequency should cause a much larger chest response than usual, meaning shorter treatment times for a patient.

CAD treatment makes mosfet amps fly

Mosfet rf power amplifiers, much improved over those currently available, are promised as result of a practical computer-design method formulated by two researchers at the Department of Electronic Engineering, University of Natal. The cad technique developed by G A Hoile and H C Reader (*IEE Proc Circuits Devices Syst*, Vol 141, No 6, pp.433-438) makes use of continuous interaction between computer simulation and physical circuitry to produce an optimised result.

Hoile and Reader looked at the typical application of commercial am aircraft radios operating at 118-136MHz, with high peak envelope powers of around 100W. Up to now, design of such amplifiers has been largely based on empirical methods, with simple linear theory used to obtain approximate initial values.

Although *Spice* modelling of bipolar transistors is not unusual, obtaining models of commonly used devices is difficult. But Hoile and Reader have been able to draw on their own mosfet model, previously reported, and combined it with the *Libra* harmonic-balance program.

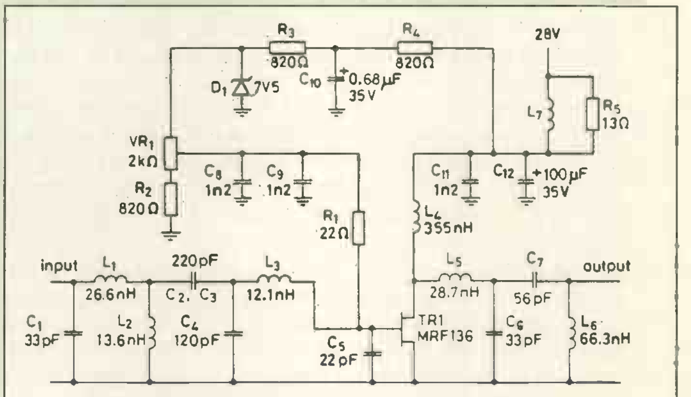
Target performance for the amplifier to be designed, constructed and tested was, over the frequency band 118-175MHz, a nominal output power of 15W, maximum flat gain and low input vswr. It also had to be stable over all operating conditions and for a load vswr of 3 or less. Design had to be built round a *MRF136* mosfet in a single-ended configuration, with a supply of 28V and quiescent current of 25mA.

Their design strategy starts with estimation of a suitable drain-load impedance, choice of stabilisation elements and the synthesis of output- and input-matching networks. *Libra* is used to calculate the impedance which the input network must transform, and to optimise the entire rf-amplifier circuit.

Towards the end of optimisation, the matching network capacitors are fixed at the nearest preferred values. Remaining variable components are then reoptimised.

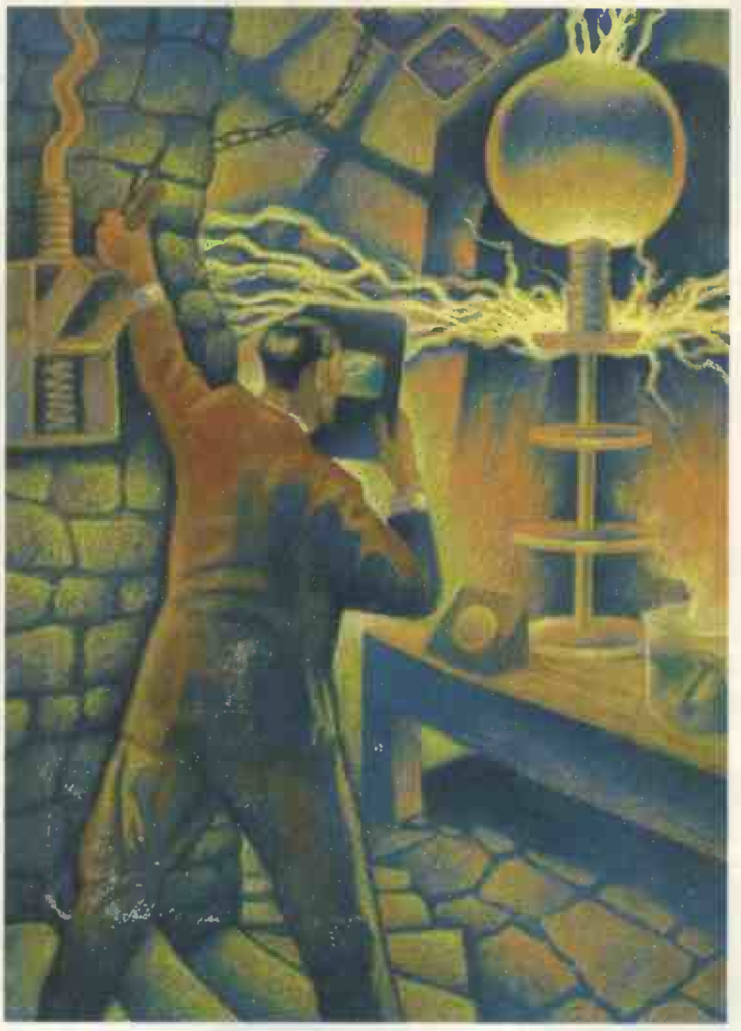
Performance of the constructed amplifier is said to be good at frequencies below 150MHz. But above this, because adjustment of the primary characteristics and parasitics of components to match the computer simulation models is not practical, there is a difference between simulated and actual results. The proposed solution is to select physical components and measure them both individually and as complete networks.

The researchers point out that the technique is more efficient and powerful than empirical optimisation of a physical circuit where only one set of values can be changed at a time. As a result it should produce rf power amplifiers with better performance than those designed by empirical methods alone. ■



Amplifier with dc-bias network produced as a result of a first non-linear optimisation.

To see a Tesla coil sending out showers of arcs in all directions at close hand is an unforgettable experience. Malcolm Watts describes the attributes and design of one such coil.



Like lightning?

Just over a century ago, Nikola Tesla was inventing devices that would revolutionise the use of electricity. Possibly his greatest invention was the ac electric motor, enabling the use of an ac distribution system for powering industrial machinery and lighting and heating systems.

Of all his inventions the Tesla coil is the most fascinating. Apart from the spectacular discharges – used for lightning effects in Hollywood – the coil is a very compact high-voltage generator. The availability of cheap eht rectifiers, such as the *BY700* from Philips series makes output rectification of low to medium power devices a real possibility.

Safety hazard
The Tesla coil described here is potentially lethal. Do not attempt to build or use it unless you fully understand the dangers of extremely high-voltages – and even then, follow carefully the safety notes on page 195.

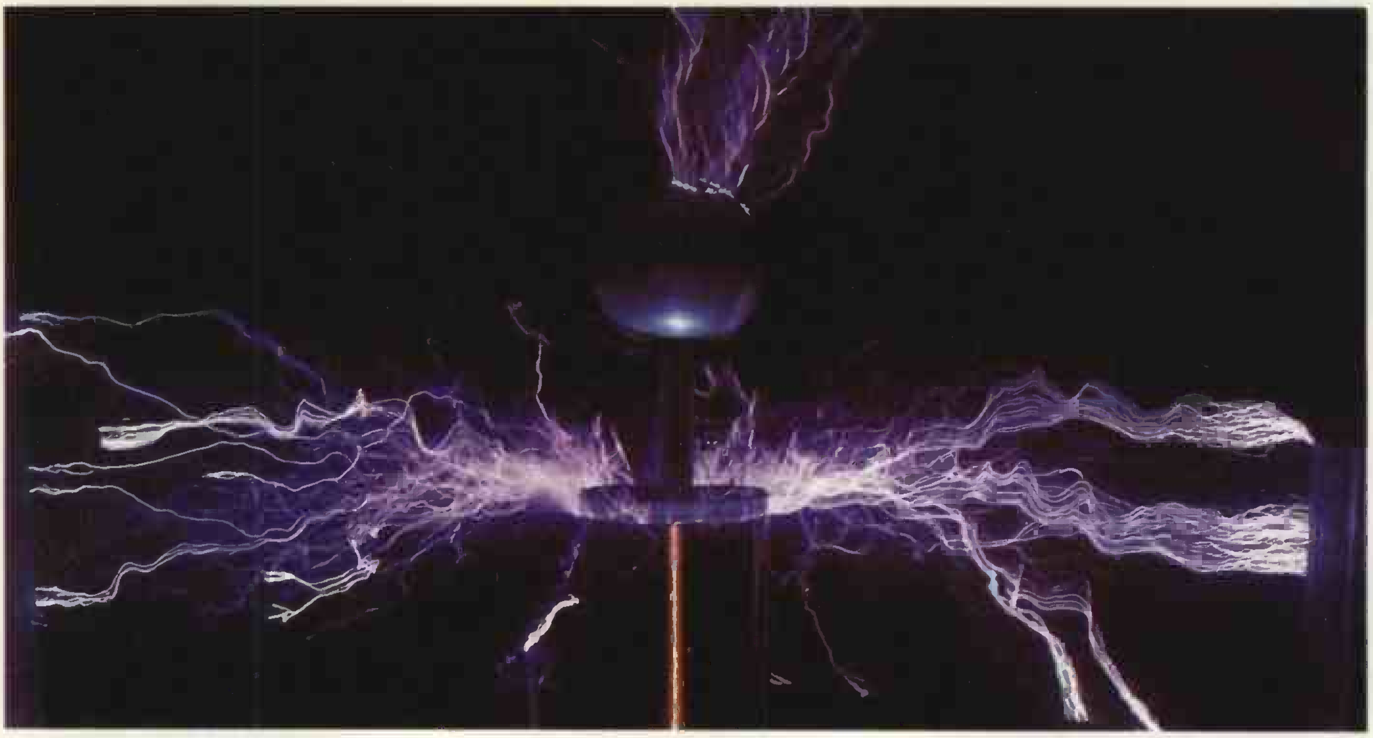
Air-cored windings

The Tesla coil is simply a transformer with primary and secondary windings but that is where the similarities with a 'normal' iron-cored transformer end. The windings are air-cored and comparatively loosely coupled. This implies a large amount of magnetic flux leakage and because of this the 'coil' behaves more like a current source than a voltage source. In fact, it has a finite internal impedance determined by physical parameters that will enable it to deliver maximum power to a load, in this case the spark, only when the load matches this impedance.

Energy not dissipated in sparks will be dissipated as electromagnetic radiation and heating of the primary components. Badly tuned and incorrectly coupled tesla coils have given the devices a notorious reputation as wideband emitters of emi. The design procedure presented here should minimise this problem.

Feeding large amounts of power into the arrangement, shown in **Fig. 1**, generates voltage gradients in the kilovolts-per-turn range across the secondary coil – hence the single-layer winding. Peak voltage at the top of the secondary coil is obtained just prior to sparks breaking out. As soon as sparks appear, a load is effectively placed across the secondary, causing the output voltage to drop as output current rises.

Spark behaviour is governed by convection and loading. In a pulse-driven coil, bursts of sparks are produced at twice the mains fre-



quency and feed the arc so formed which rises by convection. When the arc, behaving as a resistor formed from hot ionised gases, becomes stretched, current and ionisation are reduced to the point where it extinguishes and a new one forms. It follows that power consumption and loading of the system is far from constant and harmonic output will depend on the degree of variation.

Quantifying the peak output voltage is not a simple matter of turns ratio as with a standard transformer. Spark length, unlike a dc or low frequency situation, is not an accurate indicator of voltage since repetitive ionisation of the discharge channel causes successive sparks to reach progressively further until primary and secondary currents are at maximum values. Spark length is probably a more accurate indicator of system power throughput. For example, an arc over 150mm long can be generated by a low voltage high current arc welder. Experiments show that lengths of 300mm per joule of energy stored in the primary capacitor can be reached in a well built device.

How the Tesla coil works

The system shown in Fig. 1 may be recognisable to those of you familiar with spark-gap transmission systems. The major difference is the lack of an aerial which is, in effect, entirely wound on the secondary.

The top of the secondary coil connects to a metal terminal possessing some capacitance. This enhances series-resonant behaviour. See references 1&5, pages 45-50, for an excellent description of this. The ratio of inductance to capacitance in the secondary system also determines the voltage to current ratio of the output.

Primary and secondary coils are initially over-coupled in order to throughput a decent amount of power. In order for the system to approach critical coupling with minimum bandwidth and maximum power transfer it is important that the coil is driven hard enough to produce heavy sparking. Loading of the secondary causes the Q of the entire system to drop sharply which causes primary-secondary coupling to tend to the critical value.[2,3,5,6] This action also minimises radiation since power is being expended in sparks.

The primary circuit couples pulses of power to the low impedance end of the secondary coil and must resonate at the same frequency as the secondary for maximum power transfer. While the secondary is tuned by its self-capacitance and terminal capacitance, the primary coil is tuned by the primary capacitor which also serves as an energy store.

The primary capacitor is charged by the step-up transformer on every half-cycle of the mains. When its voltage causes the spark gap to fire, this energy is dumped into the primary coil and hence into a rapidly changing magnetic field that also 'cuts' the secondary coil. The speed with which the field builds up is determined by capacitor voltage and the value of inductance, including mutual secondary inductance, that the capacitor is discharged into. The faster this rate-of-change, the greater the induced emf in the secondary. As an example of the very high power levels involved, a capacitor dumping one joule of energy in ten microseconds produces a peak power of 100kW. The net effect is to transfer energy from a relatively large primary capacitor to a very small secondary capacitance.

It is essential that the resonant frequency of the primary circuit matches that of the secondary circuit as closely as possible. The primary capacitor, which virtually dictates the quality of the system, should have a minimum internal inductance and resistance, i.e. low esr for good coil performance. Old oil-filled capacitors such as those used for smoothing in transmitters work, but their current delivery capability is hampered by unsuitable internal construction, namely the thin lead wires. Their major virtue is the self-healing of the dielectric after breakdown.

Suitable initial primary-secondary coupling is required for best results. Finally, the primary chokes, whose job it is to disconnect the step-up transformer from the primary circuit when the spark gap fires, should have an impedance considerably higher than the impedance of L_P at the resonant frequency. They isolate spark-gap transients from the

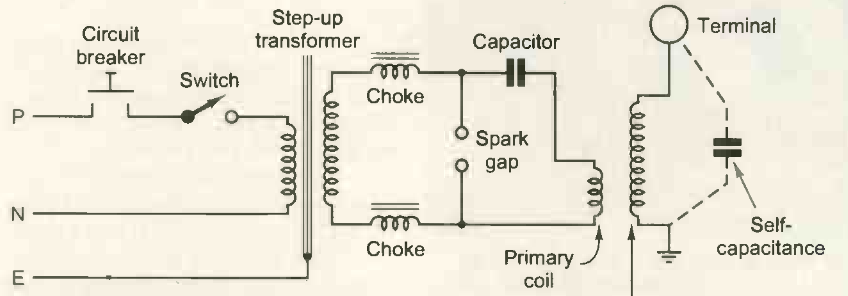


Fig. 1. Schematic of the Tesla coil generator showing primary and secondary tuned circuits.

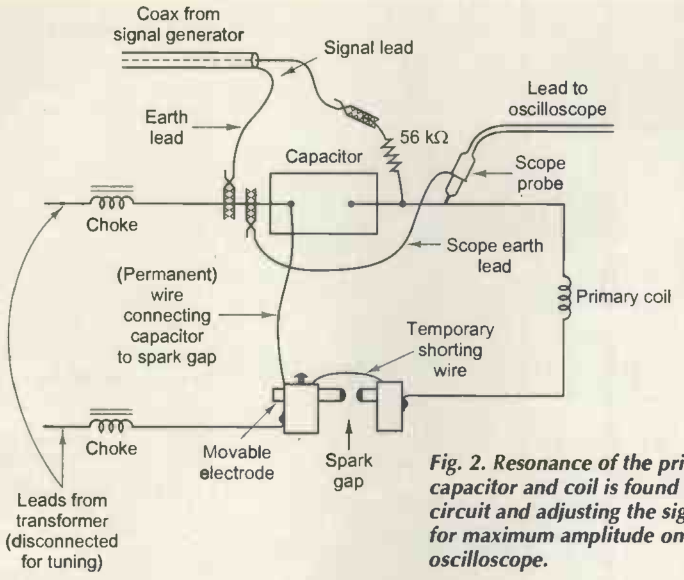


Fig. 2. Resonance of the primary capacitor and coil is found by closing the circuit and adjusting the signal generator for maximum amplitude on the oscilloscope.

mains and should be 'lossy' at high frequencies to dissipate unwanted rf as heat.

Tesla coil design

The type of coil detailed in this article is a pulse-driven two coil system. Other configurations may be driven by a power oscillator running at the resonant frequency. You can end-feed a secondary coil alone from a very low impedance oscillator for example. Other designs include additional windings.

Pulse-driven coils develop by far the highest outputs for a given input due to the trick of storing energy over a relatively long duration of 5ms and releasing it in a short period of 10μs. The recommended starting point is to choose a secondary coil height based on required spark distance. Power feeds to short high-power coils need special protection. The philosophy taken here is that simplicity is best and will leave fewer things to go wrong.

With a suitable height for the secondary selected, a former with a diameter that is somewhere between 0.2 to 0.5 times this is recommended. Secondary formers may be anything from pvc pipe to dried and varnished cardboard tubes. Rigid plastic tubing yields the lowest losses next to air except pvc which has the highest dielectric losses amongst plastics in high frequency power situations[1]. Some varnishes may also exhibit losses.

The self capacitance of the coil is calculated using Medhurst's formula, $C_{SEC}=HD$pF, where D is the coil diameter in centimetres and H is a value depending on the height to diameter ratio of the coil. Values of H range from 0.5 for *height/diameter*=2 to 0.81 for *height/diameter*=5. This formula holds if one end of the coil is connected to ground or low impedance and in close proximity to a grounded surface such as ferro-concrete. This is handy since the self capacitance is independent of the type of winding as long as it is single layer.

The next decision is whether to close-wind or space wind the coil. Close wound coils have the highest inductance but do not generally exhibit higher Q's than spaced windings because of the proximity effect[3,7] and extra wire resistance. The biggest plus to close-wound coils is that lower resonant frequency allows higher primary L/C ratios so that the Q's of the two coils can approach equality. The benefit of spaced winding is that less heavy varnishing is required. For close-wound coils, glazed-enamel magnet wire must be

The finished Tesla coil ht generator can be seen here with custom-built capacitor, and coil former containing both primary and secondary coils.

Essential equations

Terminal capacitance for a sphere

$$\frac{d^2}{7250}\text{pF}$$

where d is the sphere diameter in mm and the sphere is mounted about $d/25$ above the secondary coil. For a toroidal terminal, capacitance is,

$$\frac{(d_1-d_2)d_2}{3000}\text{pF}$$

where d_1 is the outside diameter of the ring, d_2 is the diameter of the tube and the terminal is mounted above the coil. All dimensions are millimetres.

Wheeler's formula for inductance

$$L=\frac{r^2n^2}{9r+10h}\mu\text{H}$$

where n is number of turns, r is coil radius in inches, and h is coil height in inches.

The recommended minimum wire diameter is,

$$\frac{200}{\sqrt{f}}\text{mm}$$

which is three times the skin depth at frequency f, where f is,

$$f=\frac{1}{2\pi\sqrt{LC}}\text{Hz}$$

L is secondary inductance in henries and C is the sum of coil self-capacitance and the terminal capacitance.

Since C can be calculated first, L and f values are flexible.

Primary coil design

Minimum height per turn should be,

$$0.07V_C+\frac{D_{wire}}{25}\text{mm}$$

where V_C is peak capacitor voltage in the primary components to minimise connection wiring as this will add about 0.5μH per foot and contribute to power losses, mostly in the form of em radiation. Hence it may be that a non-integral number of turns is required.

Calculating peak output voltage

$$V_{out}=\frac{Q_i\cdot V_C}{1+\left(\frac{1}{k}\cdot\sqrt{\frac{L_{primary}}{L_{secondary}}}\right)}\text{Volts}$$

where k is the initial coupling constant is bandwidth of coupled circuits divided by the resonant frequency, and V_c is peak capacitor voltage.

used because of the high voltage between turns. Self-fluxing wire may be used for spaced windings although varnish solvents may cause this type of insulation to run.

For small secondaries, litz wire is the best choice since it allows a large number of turns with high inductance to be achieved with close-wound small diameter wire while minimising skin and proximity effects.

At this stage, r and h are known. A suggested order here is to choose a frequency, calculate the wire size, calculate L from the resonance formula then calculate n using Wheeler's formula in the panel on the left, finally checking that the turns will fit into the height. Several tries may be needed until everything fits. At this stage, the resonant frequency for the primary is known and secondary construction may proceed.

A terminal that overhangs the secondary coil by a significant amount will prevent sparks issuing from the top turn of the coil. Conversely, a terminal significantly smaller in diameter than the secondary coil may be used, but discharge current will be reduced due to the lower capacitance.

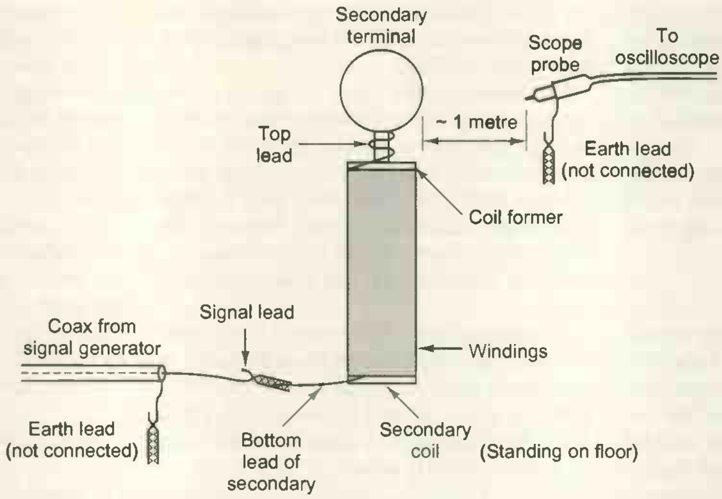


Fig. 3. Critical coupling achieves maximum power transfer. When checking the primary-secondary coupling and tuning the coil, ensure that the earth leads are not connected.

Primary coil design

Although any one of several different geometries, e.g. helical, pancake, may be used, this paper presents a method for helical design since there is less guesswork involved in arriving at satisfactory coupling and it is easy to build. The only real benefit of spirals and saucer-shaped windings is additional clearance from the top of the secondary coil.

Using the value for f obtained for the secondary coil, primary inductance is set if a suitable capacitor is available according to the resonance formula. If a choice of capacitor can be made, choose a value to suit the available transformer below.

Design the primary for a minimum diameter about twice that of the secondary for good clearance and choose coil height to be between 5 and 10 percent of the height of the secondary. For a high-Q primary with small capacitance and relatively high inductance, say 30μH, coil height should tend to the smaller value to avoid severe overcoupling. This is where a saucer-shaped coil is a better choice. This guide will give a reasonable degree of coupling between primary and secondary.

With inductance certain and height and diameter roughly known, use Wheeler's formula to calculate the number of turns.

Tuning and setting up

Tuning accuracy results in minimum electromagnetic radiation. The secondary and primary circuits must be tuned independently to the same frequency. If test equipment is unavailable, the system should be approximately correct if the design procedure has been followed.

Secondary resonance check

Stand the secondary coil upright on the floor at least a metre clear of any metal and the primary circuit. Connect only the 'signal' lead of a signal generator directly to the bottom of the coil, see **Fig. 3**. Suspend an oscilloscope probe about a metre away from the top terminal, ensuring that the terminal has a good connection to the top of the secondary. Set the scope and signal generator to 100mV then tune the coil until a signal is seen. For high Q coils, this peak can appear very suddenly. Note the frequency of peak amplitude.

Tune the generator either side of the peak for 70% amplitude and calculate the difference between these two frequencies.

Use

$$Q = \frac{1}{r} \cdot \sqrt{\frac{L}{C}}$$

to calculate the effective secondary resistance r.Next subtract the internal impedance of the generator from r, then recalculate Q. This value is Q_i (initial Q). In operation, a pvc former will degrade this figure as the secondary voltage rises.

Tuning the primary coil

This must be done with the secondary coil well away from the primary.

Connect the primary capacitor to the primary coil, disconnect the power transformer from the primary circuit, and short the spark gap with a short piece of wire.

Connect the signal generator in series with a resistor of 56kΩ across the capacitor, **Fig. 2**.

Connect the oscilloscope probe to the resistor end of the capacitor and the earth lead to the spark gap side of the capacitor.

Set the oscilloscope to 100mV/division and the generator output to 5V.

Tune the generator slowly until the amplitude on the scope peaks. This is also rather sudden. Note the frequency at which peak reading on the scope occurs.

Note the 70% frequencies as before and calculate primary Q and hence Rp, the primary series resistance using

$$Q = \frac{1}{R_p} \cdot \sqrt{\frac{L}{C}}$$

Generator impedances can be neglected. However, low impedance chokes will affect this figure.

If primary frequency is higher than secondary, try:
- Increase primary capacitance.
- Tap the primary to include more turns.
- Lower the terminal on the secondary coil until frequencies match.
- Remove a few turns from the top of the secondary coil.
- Reduce secondary terminal.

If primary frequency is lower than secondary correct by:
- Tap the primary to include fewer turns
- Reduce primary capacitance.
- Raise the terminal on the secondary.
- Increase the size of the secondary terminal.

Note that mounting the terminal too close to the top of the secondary lowers secondary Q by presenting a shorted turn to the top of the coil.

The trade-off in designing for high primary inductance is that the primary capacitor must be a smaller value for the required frequency and the applied voltage must be higher for the required output which puts more demand on capacitor breakdown voltage.

The bottom turn of the primary should start at the same height as the bottom turn of the secondary and should be mounted using good quality plastic for insulation.

Power considerations

As the spark gap may have to handle current peaks of hundreds of amps, brass rods with tungsten carbide tips are recommended. Stainless steel electrodes may be used for low power coils. Fixed gaps such as that used for the example coil already discussed are poor at 'switching off' following capacitor discharge. They can cause large transformer losses, especially if the primary chokes are low inductance.

Rotary gaps significantly improve efficiency since they minimise arcing by a switching off action. Brilliance of the gap discharge can be an indicator of excessive losses and/or poor tuning and coupling.

The most suitable types of power transformer feature current-limiting through high leakage inductance. Neon-sign transformers come into this category. However, large transformers with good regulation may be used with high inductance chokes between the transformer secondary and Tesla primary circuit; the example coil has chokes totalling 1mH.

Transformer size in conjunction with capacitor value determines power throughput. As an example, a neon transformer rated for 7500V at 60mA has an internal impedance of 125kΩ and is best matched to a capacitor of the same impedance at 50Hz since,

$$X_C = \frac{1}{2\pi f C}\Omega \qquad C = 25\text{nF}.$$

Primary chokes and capacitors

Chokes may be air or ferrite cored. The magnetic circuit in the case of ferrites must not be closed or saturation may occur.

The chokes for the example coil were made by winding a single layer of glazed-enamel magnet wire on varnished cardboard tubes and then the tubes filled with ferrite rods. Core saturation of this arrangement should never be a problem.

Air-cored chokes may be designed for a specific inductance using Wheeler's formula in the panel. Ferrite cored chokes may also be designed with this formula but will be multiplied by an induction factor for the particular ferrite used.

Choke impedance may be calculated by

$$X_L = 2\pi f L\Omega$$

For estimating inductance, the best guide is that the better the regulation of the step-up transformer, the higher the inductance needed.

The best types of capacitor are high current and voltage rated such as those used for pulsed radar[9]. Details of a rugged design suit-

Constructing a low-inductance capacitor

This design was developed to overcome difficulties in obtaining high-voltage, high-current capacitors. An additional advantage is that its value may be trimmed by removing foil during tuning.

Materials needed are 0.25mm thick clear polyethylene, commonly known as polythene sheet, available from hardware outlets and a roll of 300mm wide aluminium cooking foil. The capacitor described here is rated at 7.5kV ac for intermittent use. It is intended to be used in air, but will degrade due to coronas in any air trapped inside. Internal corona will also reduce primary Q.

To make a capacitor of 100nF needs 12m of foil and 4m of 2m-wide plastic sheet. A dust-free workbench and three people are needed. Cut the foil into 1m long pieces, 300mm wide with a 45° chamfer on each corner. Cut the sheet into 12 pieces, each measuring 500mm by 1.3m. It is important that no scratches or other defects are present. Gently wipe any dust from both sides of each piece, then accurately fold it over to obtain a doubled-up piece measuring 1.3m by 250mm.

Insert a foil into each piece so that it is centred and has one side nested into the fold. Now lay one plate, made up of the foil and plastic, on a clean surface then lay the others, one at a time on top of the previous one to form a stack with alternate foils protruding from opposite edges. There should be 25mm between the folded edge of one plate and the edge of the plastic above and below it.

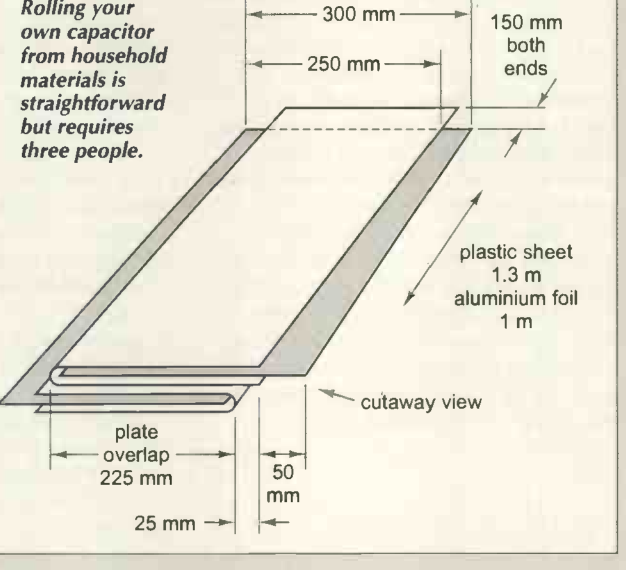
Two people should align and hold the plates in place while the third fetches the next plate, laying it on top of the growing stack. When the stack is complete, check alignment carefully and make any last minute adjustments. All folds down each side of the stack must be aligned vertically. Place a piece of 10mm thick acrylic sheet measuring 30mm by 200mm with all edges rounded across one end of the stacked plates. Roll the stack slowly and tightly around the acrylic so that it ends up in the centre.

As rolling proceeds, differences in radius of curvature will cause the plastic in different layers to try and misalign themselves. This must be allowed to happen lengthwise, but sideways misalignment must be kept to a minimum. Folded edges must not be allowed to come nearer than 15mm to any foil in an adjacent plate. When rolling is complete, hold the roll firmly and tightly wrap three pieces of 'gaffer' tape completely around it, one piece around each side and one around the middle.

The finished capacitor should be neat and tight with foils protruding out each side. Connection is made with long alligator clips clamping all foils on each side. It is a good idea to give the capacitor a test at 15kV dc. Use ear protection when discharging it and be very careful when handling it not to scrape the plastic or to come into contact with it. Check the capacitor for warmth after each 15s operation. Let it cool if the outside is warm; polyethylene melts at low temperatures.

Higher voltages may be used if additional layers of plastic are used per plate but rolling will be more difficult and the rise in working voltage may not linearly match the thickness of the plastic.

Rolling your own capacitor from household materials is straightforward but requires three people.



able for home construction are available from reference 8.

A less rugged but experimental design is presented along with this article, shown in the panel.

Coupling check

Leave the generator connected to the primary circuit, put the secondary coil inside the primary and suspend the scope probe a metre away from the secondary coil, Fig. 3. Satisfactory coupling occurs if a double-humped response is obtained with peaks 10% to 20% of f apart which equates to a coupling constant of between 0.1 and 0.2 when tuning the generator from below f to above f.

Undercoupling – peaks close together or single peak – is best corrected by increasing the L/C ratio of the primary. This should keep the primary frequency constant. Alternatively, increase the height of the primary. This will require an increase in the value of C and degrades the L/C ratio, but exposes more of the secondary to the primary. It also reduces primary/secondary clearances. Another approach is to reduce primary diameter and increase the number of turns. Any change should be followed by rechecking f and retuning if necessary.

Overcoupling is easily corrected by raising the secondary coil away from the primary a little. An overcoupled coil may exhibit flashovers between secondary turns during operation. Critical coupling – i.e. maximum power transfer – under sparking conditions requires a double-humped response when the coil is not sparking. Use the 300mm per joule criteria as a guide when operating the coil.

Firing up

Remove the spark-gap short, disconnect all instrumentation and connect the power transformer to the primary chokes.

Set the spark gap to 1mm per 5000 volts peak applied to C_p.

Ground the bottom of the secondary coil well, preferably not the mains earth.

Set up a metal discharge rod about 200mm/joule of C_p energy away from the top of the secondary coil and connect it securely to the secondary ground connection.

Connect the transformer to a mains socket with a fast-acting circuit breaker and fuse. Stand well away, and switch the mains on. If all is well, the spark gap should fire, and the secondary should send sparks to the discharge rod. If the gap doesn't fire, unplug the transformer, close the gap little, and try again.

Secondary output is peaked by widening the spark gap until it cannot be widened further without causing erratic firing. If the gap 'fizzes' and arcs, it will need widening. If problems occur that cannot be identified, consult an expert. Electrical faults can be unforgiving!

Coil example

The transformer used for this coil is capable of charging a 0.1μF capacitor to about 9000V peak. The inductance and capacitance values given may be verified by measurement and/or using the equations presented in the panel.

The coil has a space-wound secondary. Starting point for the design was the available transformer. This set a capacitor value of 0.1μF which in turn set energy storage at 4 joules. Coil height was set according to the 300mm/joule criterion to ease insulation and clearance requirements. It has 980 turns of 0.56mm glazed-enamel magnet wire wound over a 250mm diameter pvc drainpipe.

The windings were spaced by bifilar winding two lengths of the same wire, then removing the spacer wire when the winding was secured at both ends. Three coats of varnish were applied by spraying. The mean diameter of the winding is 249mm, height 1.15m, with a top terminal of diameter 229mm placed 254mm above the top of the winding.

Inductance is 44.5mH, self-capacitance is 19.1pF and terminal capacitance is 7.2pF, making total secondary capacitance of 26.3pF. The primary consists of 3 turns of 9mm outside-diameter copper pipe, wound to a diameter of 660mm, with a winding height of 80mm. Six acrylic slabs with slots cut to match winding height are used to space and support the coil.

The spark gap consists of two pieces of half-inch diameter brass rod with hemispherically-shaped tungsten carbide tips brazed on, mounted in brass blocks with one electrode being movable to allow adjustment. Primary inductance is 11.8μH. The capacitor was trimmed during tuning to match the secondary frequency of 146.5kHz. Initial secondary Q is 180 and primary Q is 50. This does not meet the 'equal Q' requirement and illustrates a trade-off made to best utilise an available transformer. Since the top terminal does not significantly 'overhang' the secondary, being the only one available at the time, sparks issue from the top turn. To prevent burning the former, the last two turns were wound over a layer of Teflon tape.

My thanks go to Russell Stevens of the Dept. of E.E., Mike Salmon and Ralph Fletcher and Paul Bryant for their help in producing this article. ■

Vital safety information

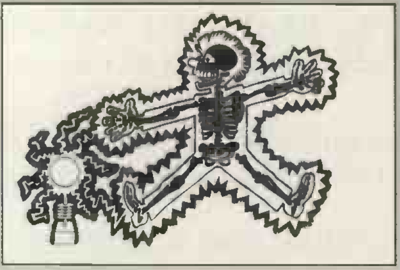
Tesla coils are both a shock and fire hazard.

Never

- Operate the coil in the presence of flammable substances, vapours or gasses.
- Earth any part of the primary circuit.
- Stare at the spark gap discharge – it has a strong ultra-violet light content.
- Attract a discharge from a large pulse-operated coil to a person. Megawatt pulses of power with a strong 100Hz content will be present in the secondary discharges of a high power coil!

Always

- Provide a terminal connected to the secondary ground for the coil to discharge to when the spark distance is unknown i.e. when firing up for the first time!
- Unplug the power transformer from the mains when adjusting the spark gap. Electrocution is guaranteed when contact is made with the primary circuit if power is accidentally applied!
- Wear shoes with thick rubber or plastic soles when experimenting.
- Use a circuit breaker between the mains and the transformer when operating the coil.
- Ensure a good ground for the secondary coil, preferably separate from the mains earth. • Wear ear protection when operating.
- Unplug the coil and short the capacitor terminals together before making adjustments.
- Check for capacitor and transformer heating before firing up repeatedly. If the transformer or capacitor feel warm, they may be HOT internally.



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Solving PROBLEMS

Although not 100% intuitive, the maths modelling package Matlab has the potential for helping electronics engineers solve a multitude of maths and graphing problems, finds Alan Brown.

Recently released as version 4.0, *Matlab* is a powerful mathematical software package featuring extensive graphics facilities, most of which are of potential interest to electronics engineers.

The package is a technical computing environment for high-performance numerical computational and visualisation. It integrates numerical analysis, matrix computation, signal processing and graphics in a convenient environment, and using it does not involve extensive programming.

Produced by *Math Works* of Massachusetts, *Matlab* is distributed in the UK by Rapid Data. The package is available for platforms including Macs, Sun workstations and pcs.

Matlab is a stand-alone package, but it can be augmented with a variety of 'toolboxes'. Each toolbox covers a specific application, such as image processing, control system design, digital signal processing, dynamic system simulation and neural network calculation. This review covers the core *Matlab* package and its likely application in electronics; the plan is to cover toolboxes in a future issue.

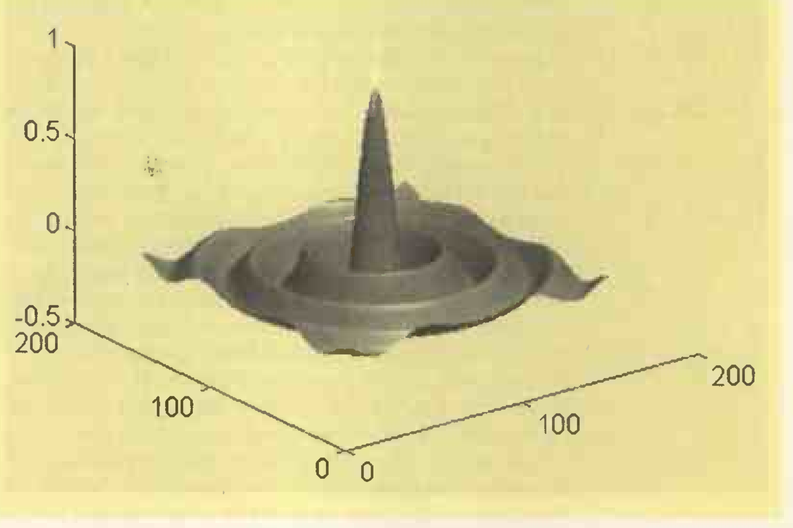


Fig. 1. The sinc function defined in a matrix and plotted as a matrix in Matlab.

Command-line editor

By combining numerical analysis, matrix computation, signal processing and graphics into a single package, *Matlab* provides a framework for problem solving without the need to resort to traditional programming.

The software operates from a command-line editor, where each line is typed in by the user. Each line can either be a stand alone operation of form part of a sequence. In the executions of its commands, the package provides for conditionals. This means that program sequences can be generated. These can be saves as so-called M-files. In fact much of *Matlab* consists of libraries of M-files which are routines made up of basic *Matlab* primitive commands.

The command-line editor is structured as a last-in first-out buffer. This allows previous commands to be recovered, but it does not allow much flexibility. If, for example, a number of command lines have been entered and you wish to modify a command several lines back then you have to loop the buffer until to reach the offending line. If subsequent lines depend on the one modified, they all have to be recovered and re-executed – which is quite messy. A screen-addressable editor would be much more useful.

All the *Matlab* functions can be evoked rom the command-line editor. One of the problems is realising just how many functions there are. The command-line editor 'help' facility is useful in this context, except you have to know what to ask for before the information is provided. Alternatively it is possible to access the traditional Windows help structure and search through the contents.

Among the features offered by *Matlab* is the opportunity to perform numerical modelling by solving differential equations. This includes coupled non-linear simultaneous differential equations. However I must point out that the choice of integration algorithms is limited to second/third order Runge Kutta or the fourth/fifth Runge Kutta Fehlberg methods. For some multi-iterative modelling requirements these methods may not provide sufficient accuracy, the user should therefore be aware of these limitations.

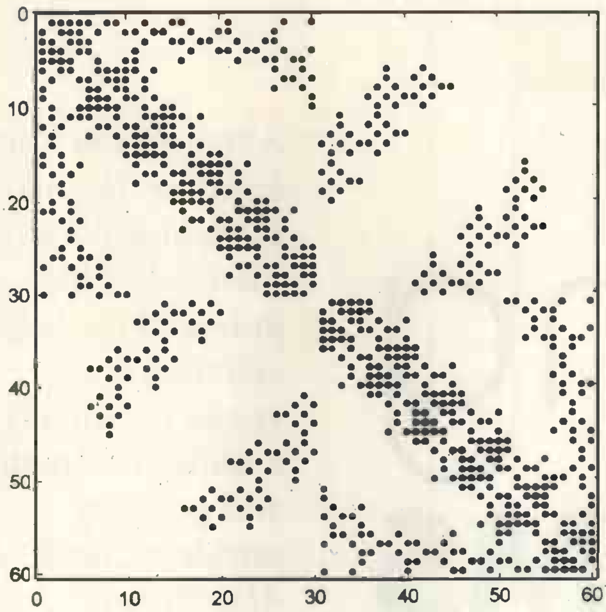


Fig. 2. An example of a sparse matrix showing the positions of data points which have finite values.

Matrix and array operations

Where *Matlab* really scores is in its options for the numerical manipulation of matrices and vectors. There are almost no type of matrix or matrix operation which are not accommodated. The range of operations is very impressive, as is the size of the matrices. **Figure 1** shows a mesh plot of the famous sinc function defined as a matrix. *Matlab* is remarkably fast in performing matrix operations. Inverting a 100 by 100 matrix is performed almost instantaneously on a 33MHz 486 pc. However to ensure that the software works at full speed the size of the matrices must not exceed the available system ram.

When *Matlab* has to resort to using virtual memory, i.e. treating the hard-disc as if it is ram, there is a punishing reduction in performance, particularly where ram capacity is limited. Many modelling operations require the use of matrices to solve set of simultaneous equations. Each floating point number requires eight bytes. Many matrices however have zero value elements. Performing operations on these wastes both memory and computation time.

To minimise wasted computation, *Matlab* has a facility for manipulating sparse matrices. An example showing the distribution of non-zero elements of a sparse matrix is shown in **Figure 2**. As a rule, the computing time needed for sparse matrix operation is proportional to the number of arithmetic operations on the non-zero elements.

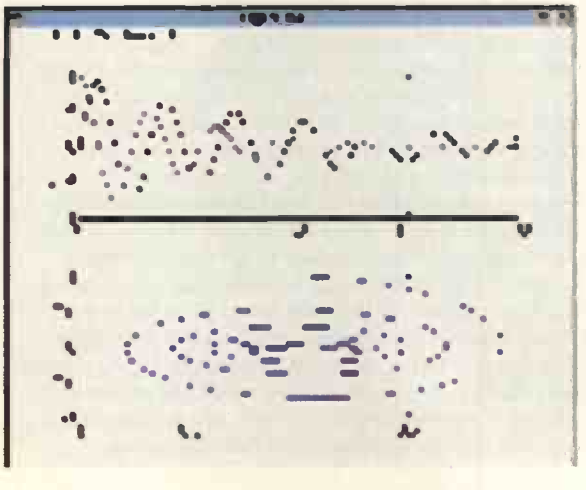


Fig. 3. An example of 2D plotting in Matlab, showing (y,t) and (y,x) graphs.

By using sparse matrices, the software cuts down on the memory usage and number crunching. To quote an example from the User's Guide, solving the full matrix equation $Dx=b$ would take 12 hours and require 128Mbyte of memory; variable D is a 4096 by 4096 matrix with 20,224 nonzero elements. Defining an equivalent sparse matrix for $Dx=b$ would take only ten seconds and 0.25Mbyte of memory, which is worthwhile considering.

Sparse matrices are frequently used in finite element analysis. In general *Matlab* does lend itself very well to solving large sets of simultaneous equation, and would serve a useful tool in analysing electronic networks which usually have many zero value elements.

Programming and control flow

Although *Matlab* uses a command-line for entering instructions, it is possible to configure instructions into a sequence and introduce conditionals. Program constructs such as FOR, WHILE, ELSEIF and ELSE can be used to implement loop operations and conditionals.

Within a FOR or WHILE loop, any of the *Matlab* commands may be used. In fact the user has the option of defining their own functions (.M files) which can be stored and called up for future operations. The program loops can be nested as required and each loop is terminated by an END instruction. To use the programming options offered by *Matlab* it does help if the user is familiar with programming languages and techniques.

2D, 3D graphics and images

One of the notable virtues of *Matlab* is its visualisation features. Much of the package is given over to graphics generation and I find the range and quality of the graphics objects very impressive. Armed with *Matlab*, I cannot see anyone ever writing their own graphics software again. Almost anything needed can be constructed from the list of graphics instructions.

An example of the 2D plotting is shown in **Fig. 3**. The user has complete control over axis scaling, labelling and plotting styles – not only in cartesian co-ordinates but also in polar plots.

There are many packages which perform as well as *Matlab* in 2D plotting but the 3D plotting facilities in *Matlab* are quite spectacular. They accommodate shading, contour plots and animation for mode simulation purposes, line trajectories and volumetric slice plots. **Figure 4** shows an example of using pseudo colour in combination with the meshgrid and contour plot. Bearing in mind that images can be imported directly into *Matlab* from a flatbed scanner, this type of process would be useful for viewing thermal profiles derived from infrared images of populated printed circuit boards under test.

It is possible for the user to exercise considerable control over the 3D plotting process, enabling the whole range of plotting features. Example are axis units, labelling and palette. But using these effectively does require a certain amount of practice.

All plots are constructed by combining *Matlab* graphic objects, of which there are nine describing all aspects of a graph or plot. Each object has a handle that is assigned to it, therefore each plot is composed of multiple objects and when designing a customised graph the user manipulates the appropriate handles.

External interfacing

Data produced within *Matlab* can be exported to other applications and there are numerous controls over file formats available. Alternatively data can be imported into the package for further processing or just for displaying graphically. The simplest instructions are load and save.

Given a array of numbers A for example, they can be saved on in an ascii file (temp.dat) on disc by using the command:

```
save temp.dat A -ascii
```

If your pc is equipped with a sound card, it is even possible to hear your data files – useful if you are processing audio signals. The software can process standard audio .WAV files generated from sound cards. This provides the opportunity for using a sound card as a conventional data acquisition card.

Additionally, there is a provision to execute external C or Fortran subroutines from within *Matlab*. These are collectively called MEX files and employ the dynamic linking facilities of Windows. This is useful if you already have written and compiled programs that you would like to run in conjunction with *Matlab*. For example, the plotting features of *Matlab* could be used to display the data generated form the external MEX file.

A word of caution. The MEX-files must be produced by a recognised 32-bit compiler which appears not to include the Microsoft Visual C++ version.

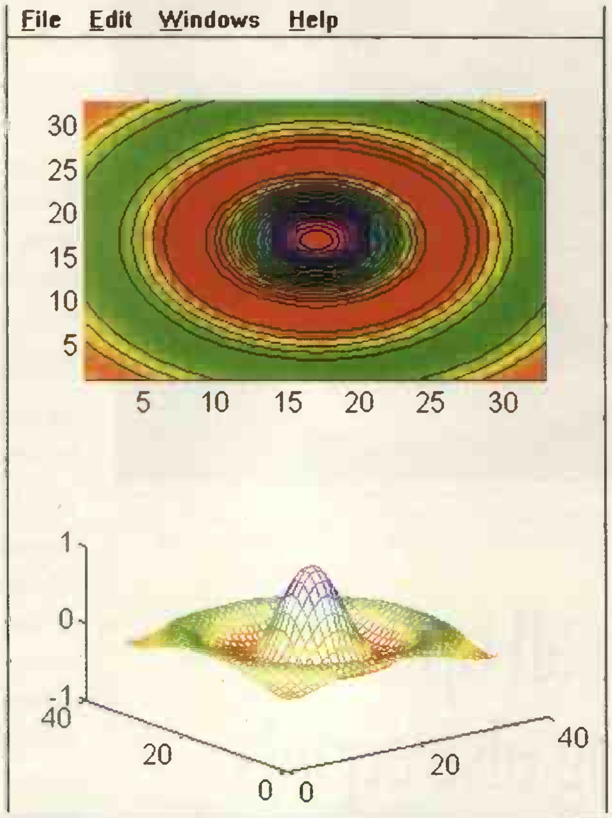


Fig. 4. An example of 3D plotting with pseudo colours. This can be used to highlight graduations.

Graphics user interface

To emphasise the interactive nature of *Matlab*, the user has the option of including control features in the display. For illustrate, when performing a model simulation it may be advantageous to adjust some of the model parameters in real-time.

To cater for such needs *Matlab* provides six control items, namely push button, check box, radio button, scroll bar (slider), drop-down list, static text field, text box and static rectangle. These make it possible to construct a screen display which is wholly controllable by the user – especially useful if the model is likely to be used by others who wish to use it interactively. Users of other Windows product will be familiar with many of these interactive feature as their use is widespread in pc software.

Figure 5 shows a display from the demonstration software supplied with *Matlab*. It indicates a typical construction of a graphical user interface. As the software becomes more widespread it is likely that highly specialised software routines (M-files) will be provided by third parties. Features of the graphical user interface should ensure that the routines are easy to drive.

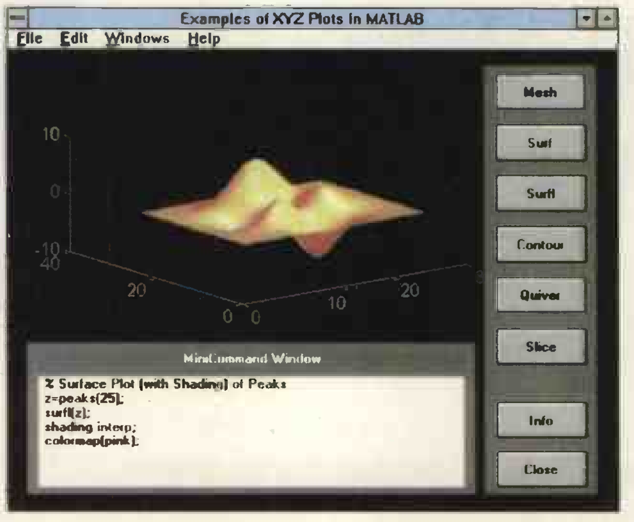


Fig. 5. An example of the use of Graphic User features for interactive processing.

User manuals

Matlab is now a few years old and the manufacturers have obviously had the opportunity to work on the user's manuals. The product comes with no fewer than five attractive manuals which on the whole are well written. The user and reference guides are essential while the other manuals relate to the external interfacing, release notes and the graphical user interfacing.

Many examples are given to illustrate the software's operation and these form a very useful learning aid. The user guide even has few colour plates to illustrate the graphics features of *Matlab*.

Conclusion

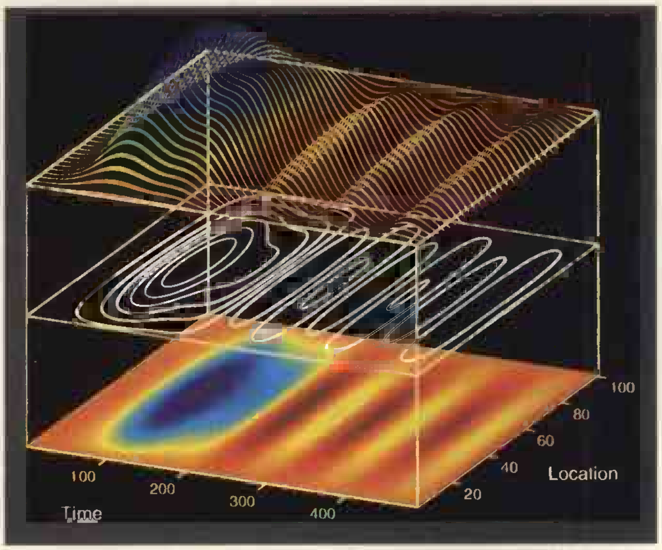
Although *Matlab* is a powerful package, the new user still faces the struggle of mastering the command-line structure of the product. It has peculiar features which are not intuitive, but once the new user is familiar with the package's basic operation they will soon feel comfortable using it for a large variety of modelling problems.

As with most things in life, to gain a working proficiency with *Matlab* a lot of practice and personal application is required. The package should be viewed as an environment which can be customised to perform specific simulation and modelling functions.

Matlab's toolboxes make the package particularly attractive to electronics engineers. These include a link simulator, a control system tool and image and signal processing options. ■

Recommended minimum system requirements
486DX 33 MHz PC
8 Mbyte System RAM
SVGA Graphics monitor with graphics accelerator card
Windows 3.1

Availability
Rapid Data Ltd., Crescent House, Crescent Road, Worthing, West Sussex BN11 5RW. Telephone: 01903-202819, fax: 01903-820762. Price of the base *Matlab* package is £1440 and the educational version is £475. These prices include documentation and full technical support, but exclude £15 for delivery and VAT.



Three views of bending forces on a magnetic levitation train guideway. Analysis was one part of an 80th-order differential equation modeled with MATLAB. Data courtesy of Grumman Corp.

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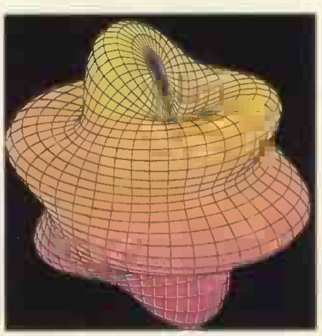

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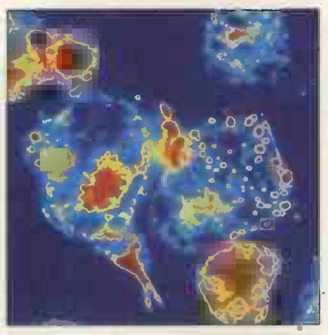

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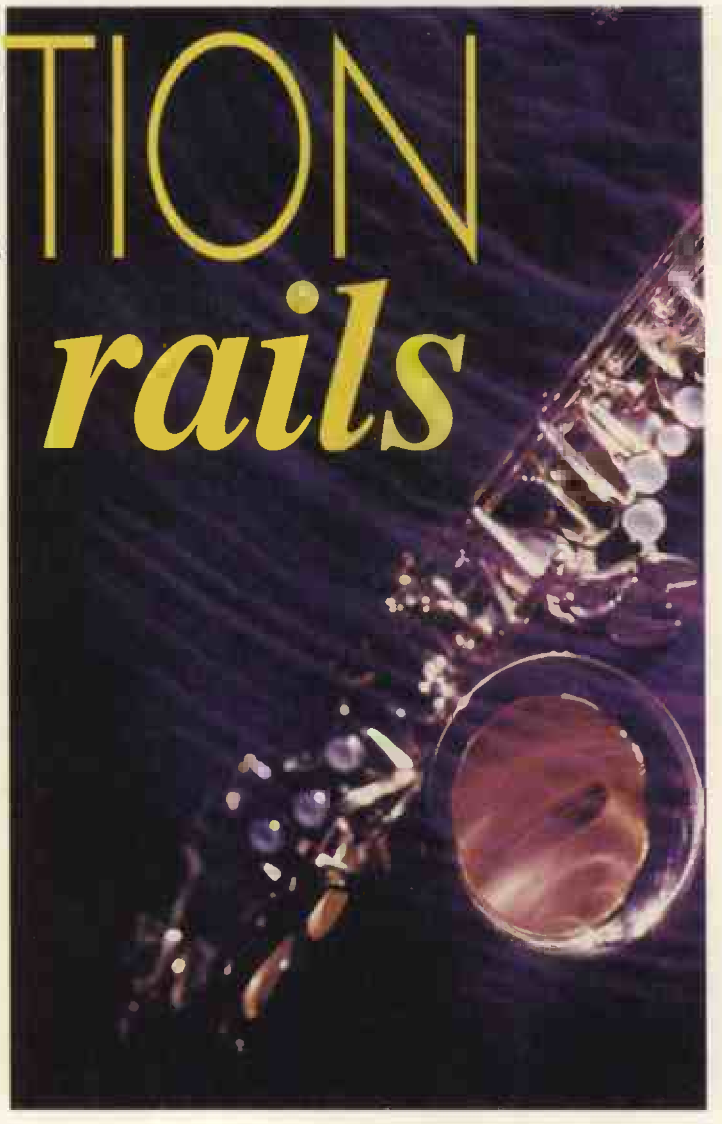
CIRCLE NO. 109 ON REPLY CARD

DISTORTION off the rails



***Douglas Self** exposes the subtle ways that power supply disturbances can intrude into an audio amplifier and demonstrates that electronically regulated supplies are not necessary for good performance.*

There has been much discussion recently about the importance of power-supply rejection in audio amplifiers, particularly with regard to its possible effects on distortion.[1,2]

I have – I hope – shown in my previous article[3] that regulated power supplies are unnecessary for exemplary thd performance. I want to confirm this by examining just how supply-rail disturbances affect the output, and by looking at the ways in which this rail-injection can be eliminated. My aim is not just the production of hum-free amplifiers, but also to show that there is nothing mysterious in power-supply effects, no matter what subjectivists may say. As before, this study addresses many kinds of amplifier – not just those intended to drive loudspeakers.

The adverse effects of inadequate power-supply rejection ratio, psrr, in a typical class-B power amplifier with a simple unregulated supply, may be two-fold:

- A proportion of the 100Hz ripple on the rails will appear at the output, degrading the noise/hum performance. Most people find this much more disturbing than the equivalent amount of distortion.

- The rails carry a signal-related component, due to their finite impedance. In a Class-B amplifier this is in the form of half-wave pulses, as the output current is drawn from the two supply rails alternately; if this enters the signal path it will degrade thd seriously.

The second possibility – generation of distortion from rail injection – can be arranged to cause very little trouble in practice, at least for the conventional amplifier architectures I have so far discussed. The most common defect seems to be misconnected rail-bypass capacitors, which add copious ripple and distortion to the signal if their return lines share the signal ground; this was denoted No 5 in my list of distortion mechanisms.[1]

This must not be confused with distortion caused by *inductive* coupling of half-wave supply currents into the signal path- this effect is wholly unrelated and is completely determined by the care put into physical layout; I labelled this Distortion No 6

Assuming these bypass capacitors are connected correctly, with a separate ground return, ripple and distortion can only enter the amplifier directly through the circuitry. It is my experience that if the amplifier is made ripple-proof under load, then it is proof against distortion-components from the rails as well; this bold statement does however require a couple of qualifications.

Firstly, the output must be ripple-free under load, i.e. with a substantial ripple amplitude on the rails. If a

Class-B amplifier is measured for ripple output when quiescent, there will be a very low amplitude on the supply rails and the measurement may be very good. But this gives no assurance that hum will not be added to the signal when the amplifier is operating and drawing significant current from the reservoir capacitors.

Spectrum analysis could be used to sort the ripple from the signal under drive, but it is simpler to leave the amplifier undriven and artificially provoke ripple on the supply rails by loading them with a sizable power resistor; in my work I have standardised on drawing 1A. One rail at a time can be loaded; since the rail rejection mechanisms are quite different for positive and negative rails, this is a great advantage.

Drawing 1A from the negative rail of the typical power amplifier in **Fig. 1** degraded the measured ripple output from –88dBu (mostly noise) to –80dBu.

Secondly, I assume that any rail filtering arrangements will work with constant or increasing effectiveness as frequency increases; this is clearly true for resistor-capacitor filtering, but is by no means certain for 'electronic' decoupling such as the negative-feedback current-source biasing used in the design in reference 4. These will show declining effectiveness with frequency as internal loop-gains fall. If 100Hz components are below the noise in the thd residual, it can usually be assumed that disturbances at higher frequencies will also be invisible, and not contributing to the total distortion.

To start with some facts proven by experiment. I took a power amplifier – similar to Fig. 1 – powered by an unregulated supply on the same pcb. The significance of this proximity will become clear in a moment. Driving power was 140W rms into 8Ω at 1kHz. The supply was a conventional bridge rectifier feeding 10,000μF reservoir capacity per rail.

Amplitude of the 100Hz rail ripple under these conditions was 1V pk-pk. Superimposed on this were the expected half-wave pulses at signal frequency. Measured at the pcb track just before the supply-rail fuse, their amplitude was approximately 100mV pk-pk. This doubled to 200mV on the downstream side of the fuse – the tiny resistance of a 6.3A slow-blow fuse is sufficient to double this aspect of the power supply rejection ratio problem, and so the fine details of pcb layout and power supply wiring may well have a major effect. The 100Hz ripple amplitude is of course unchanged by the fuse resistance.

From this it is clear that improving the 'transmitting' end of the problem is likely to

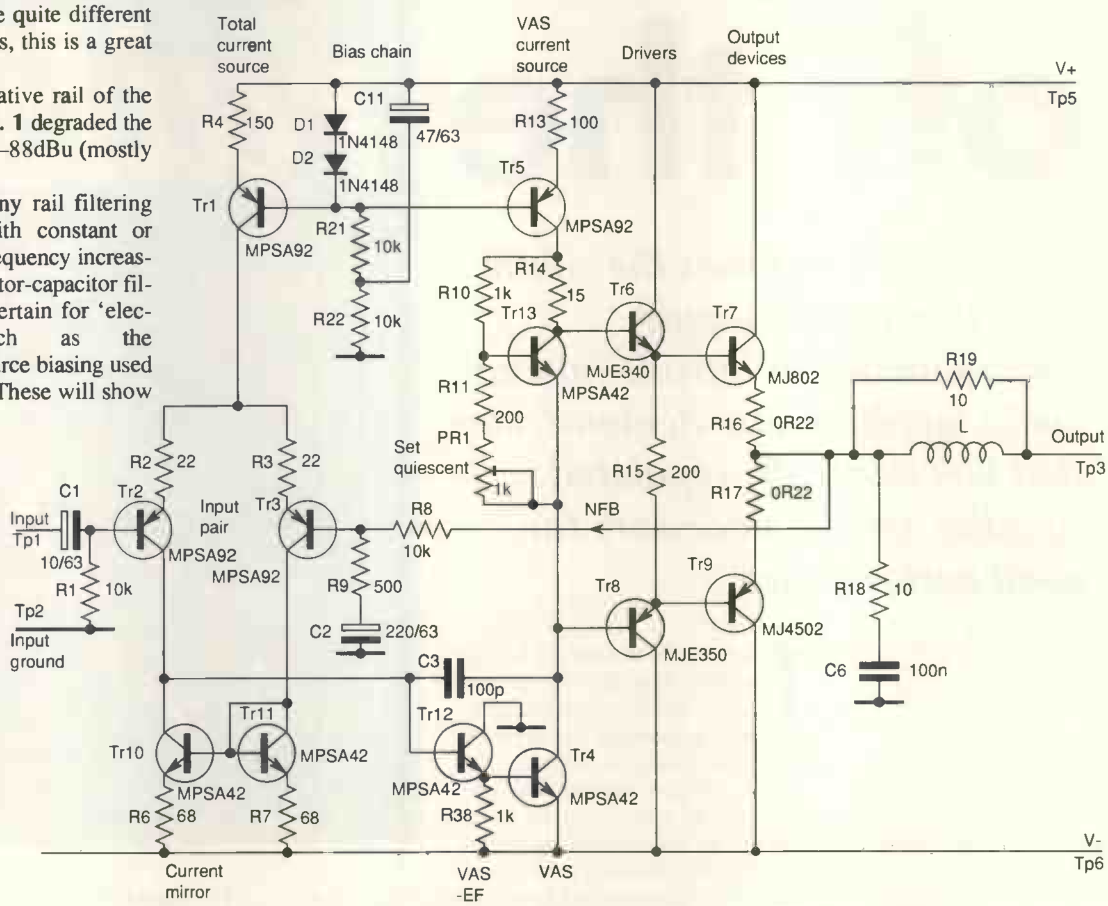


Fig. 1. In this generic power amplifier, diode biasing is used for the input tail and voltage amplifier source.

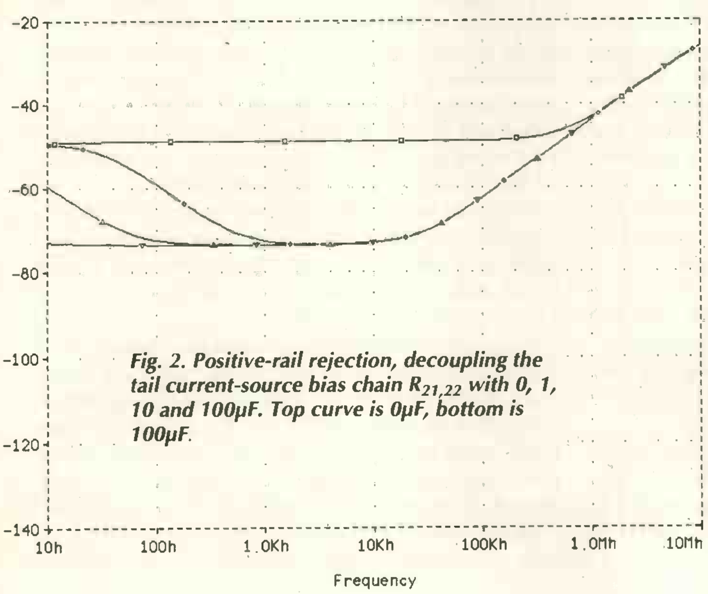


Fig. 2. Positive-rail rejection, decoupling the tail current-source bias chain $R_{21,22}$ with 0, 1, 10 and 100μF. Top curve is 0μF, bottom is 100μF.

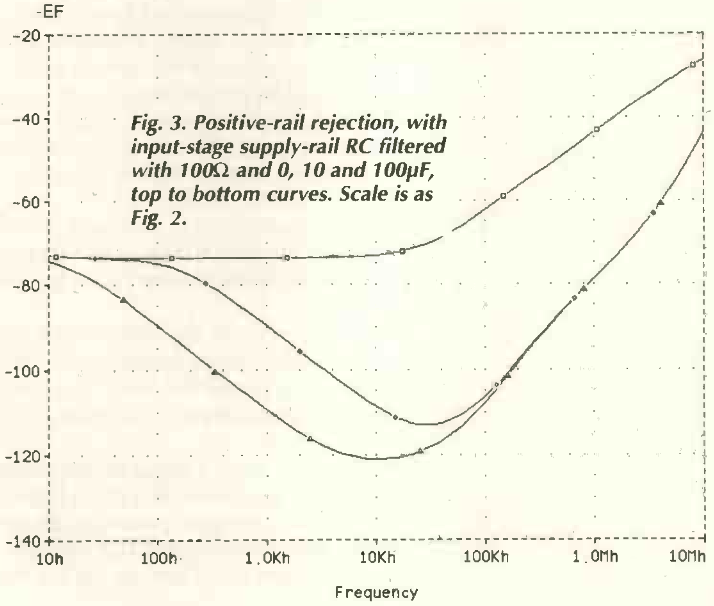


Fig. 3. Positive-rail rejection, with input-stage supply-rail RC filtered with 100Ω and 0, 10 and 100μF, top to bottom curves. Scale is as Fig. 2.

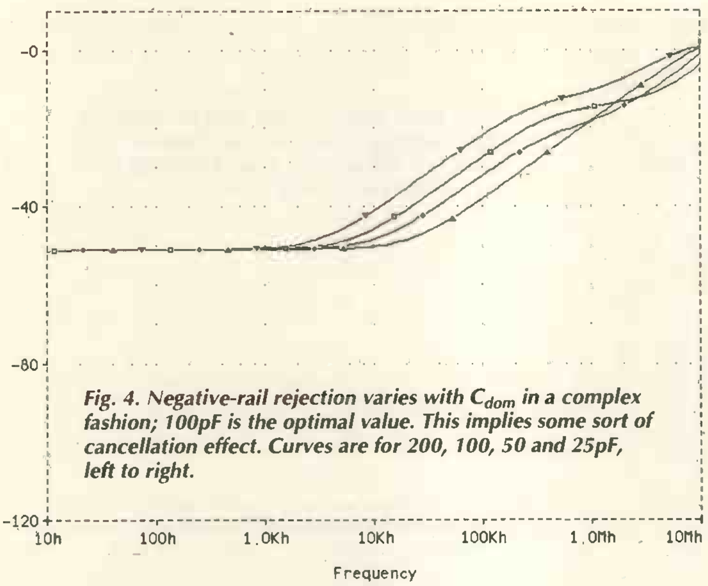


Fig. 4. Negative-rail rejection varies with C_{dom} in a complex fashion; 100pF is the optimal value. This implies some sort of cancellation effect. Curves are for 200, 100, 50 and 25pF, left to right.

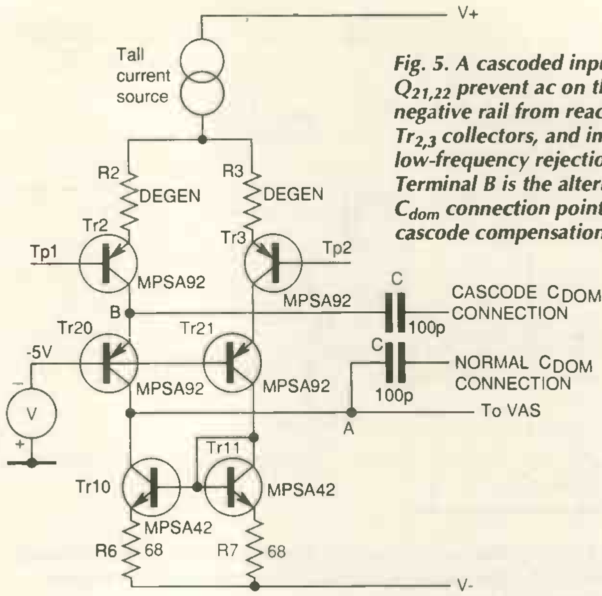


Fig. 5. A cascoded input stage; $Q_{21,22}$ prevent ac on the negative rail from reaching $Tr_{2,3}$ collectors, and improve low-frequency rejection. Terminal B is the alternative C_{dom} connection point for cascode compensation.

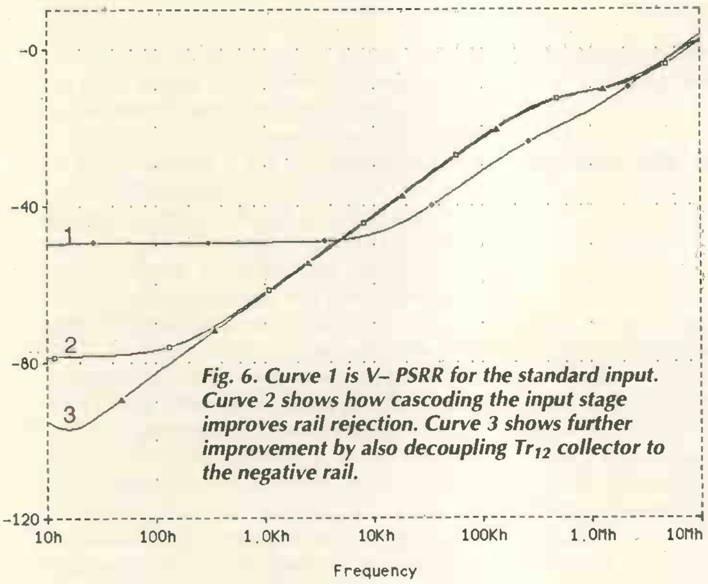


Fig. 6. Curve 1 is V– PSRR for the standard input. Curve 2 shows how cascoding the input stage improves rail rejection. Curve 3 shows further improvement by also decoupling Tr_{12} collector to the negative rail.

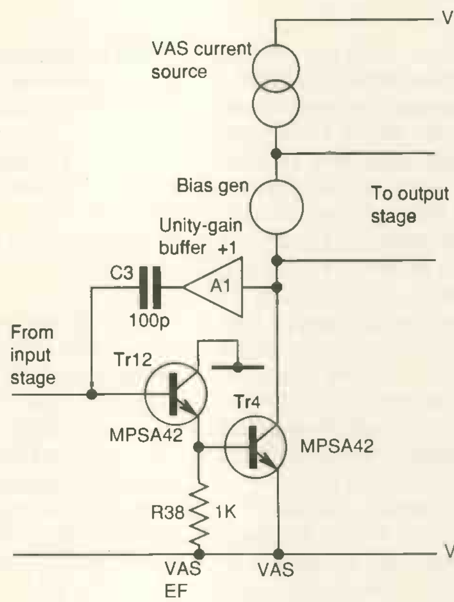


Fig. 7. Adding C_{dom} buffer A_1 to prevent any possibility of signal entering directly from the negative rail.

be difficult and expensive, involving extra-heavy wire etc, to minimise the resistance between the reservoirs and the amplifier. It is much cheaper and easier to attack the 'receiving' end, by improving the power amp's psrr. The same applies to 100Hz ripple; the only way to reduce it is to increase reservoir capacity, and this is expensive.

Designing for rejection

Firstly, ensure there is a negligible ripple component in the noise output of the quiescent amplifier. This should be easy, as the supply ripple will be minimal; any 50Hz components are probably due to magnetic induction from the transformer, and must be removed first by attention to physical layout.

Secondly, the thd residual is examined under full drive; the ripple components here are obvious as they slide along the distortion waveform – assuming that the scope is synchronised to the test signal. As another general rule, if an amplifier is made visually free of ripple-synchronous artifacts on the thd residual, then it will not suffer detectable distortion from the supply rails.

In a discrete power amplifier, supply rejection ratio is usually best dealt with by *RC* filtering. However, this will be ineffective against vlf components below 50Hz, resulting from short-term mains voltage variations reflected in the supply rails.

A design relying wholly on *RC* filtering might have low ac ripple figures, but would show irregular jumps and twitches of the thd residual; hence the use of constant-current sources in the input tail and voltage amplification stage to establish operating conditions more firmly.

The usual op-amp definition of supply rejection ratio is the decibel loss between the supply rail and the effective differential signal at the inputs, giving a figure independent of closed-loop gain. However, here I use the decibel loss between rail and output, in a standard non-inverting configuration with a closed loop gain of 26.4dB. This is the gain of the amplifier circuit in Fig. 1 and reference 4, and allows decibel figures to be directly related to test gear readings.

Looking at Fig. 1, we must assume that any connection to either supply rail is a possible entry point for ripple injection. Fortunately, the two rails can be treated independently; their behaviour is quite different.

The positive rail

On the positive rail, injection points that must be taken seriously are the input-pair tail and the voltage amplifier stage collector load. There is little temptation to use a simple resistor tail for the input; the cost saving is negligible and the ripple performance inadequate – even with a decoupled mid-point. A practical value for such a tail-resistor would be 22kΩ, which in Spice simulation gives a low-frequency supply rejection ratio of –120dB for an undegenerated differential pair with current-mirror.

Replacing this tail resistor with the usual current source improves this to –164dB, assuming the source has a clean bias voltage.

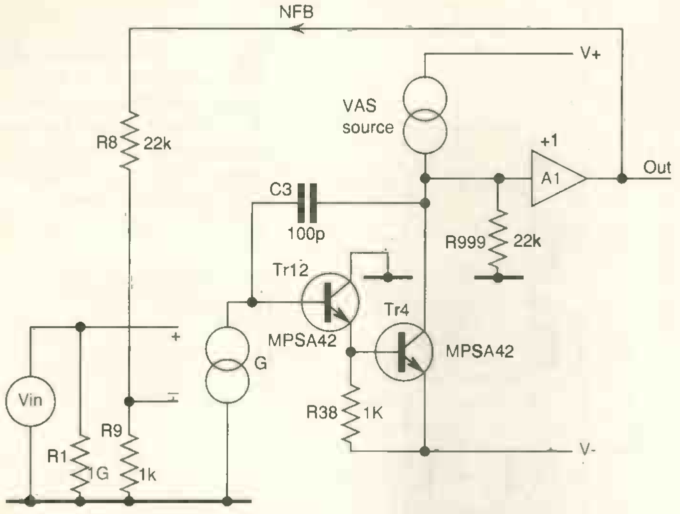


Fig. 8. A conceptual Spice model for negative-rail rejection, with only the voltage amplifier made from real components. R_{999} represents voltage-amplifier loading.

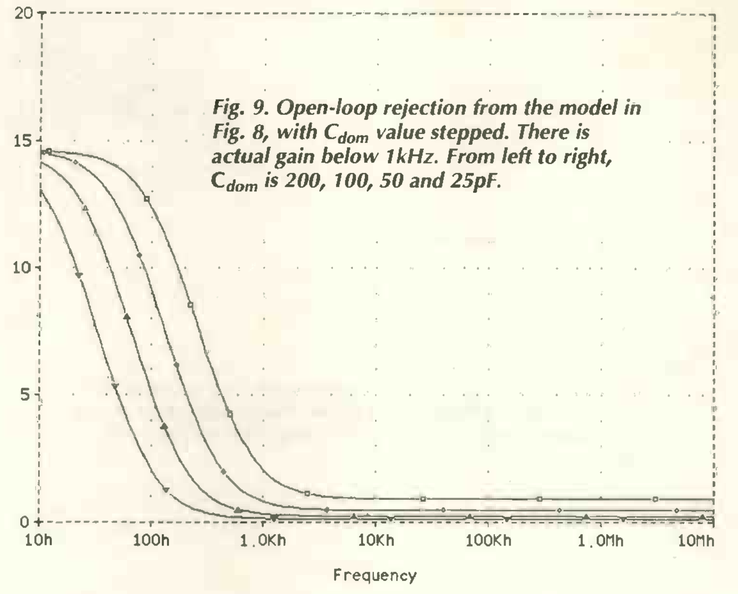


Fig. 9. Open-loop rejection from the model in Fig. 8, with C_{dom} value stepped. There is actual gain below 1kHz. From left to right, C_{dom} is 200, 100, 50 and 25pF.

The improvement of 44dB is directly attributable to the greater output impedance of a current source compared with a tail resistor. With the values shown this is 4.6MΩ, and 4.6MΩ/22kΩ is 46dB, which is a very reasonable agreement. Since the rail signal is unlikely to exceed +10dBu, this would result in a maximum output ripple of –154dBu.

Ripple excluded, the measured noise floor of the amplifier of Fig. 1 was –94.2dBu for *EIN*=–121.4dBu. This is mostly Johnson noise from the emitter degeneration resistors and the global negative feedback network. Tail ripple contribution would therefore be 60dB below the noise, where I think it is safe to neglect it.

In reality, the tail-source bias voltage will not be perfect; it will be derived from the positive rail, with ripple hopefully excluded. The classic method involves a pair of silicon diodes; led biasing provides excellent temperature compensation, but such accuracy in setting dc conditions is probably unnecessary.

It may be desirable to bias the voltage amplifier stage collector current-source from the same voltage. This rules out anything above a volt or two. A 10V zener might be appropriate for biasing the tail-source – given suitable precautions against noise generation – but this would seriously curtail the positive voltage amplifier-stage voltage swing.

The negative-feedback source-biasing system in reference 4 provides a better basic supply rejection than diodes, at the expense of some beta-dependence. It is not quite as good as an led, but the lower voltage generated is more suitable for biasing a voltage amplifier stage source. These differences become academic if the bias chain mid-point is filtered with 47µF to the positive rail, as **Table 1** shows; this is C_{11} in Fig. 1.

As another example, the amplifier in Fig. 1 with diode-biasing and no bias-chain filtering gives an output ripple of –74dBu; with 47µF filtering this improves to –92dBu, and 220µF drops the reading to below the noise floor.

Figure 2 shows Pspice simulation of Fig. 1, with a 0dB sine wave superimposed on the positive rail only. A large decoupling capacitor, such as 100µF, improves low-frequency supply rejection by about 20dB. This should drop the residual ripple below the noise. However, there remains another frequency-insensitive mechanism at about –70 dB.

Table 1. Decoupling the midpoint of the current-source bias chain improves rejection.

	No decouple	47µF decouple
2 diodes	–65dB	–87dB
LED	–77dB	–86dB
NFB low β	–74dB	–86dB
NFB high β	–77dB	–86dB

The study of supply rejection ratio greatly resembles the peeling of onions – there is layer after layer, and often tears... There also remains an high-frequency injection route, starting at about 100kHz in Fig. 2. This is unaffected by the bias-chain decoupling.

Rather than digging deeper into the precise mechanisms of the next layer, it is simplest to *RC* filter the positive supply to the input pair only as a few volts lost here are of no consequence. It makes very little difference if the source of the voltage amplifier stage is decoupled or not.

Figure 3 shows the very beneficial effect of this at middle frequencies, where the ear is most sensitive to ripple components.

Negative rail considerations

The negative rail is the major route for injection, and difficult to analyse. The well-tried Wolf-Fence approach is to divide the problem in half. In this case, the Fence is erected by applying *RC* filtering to the small-signal section – i.e. input current-mirror and voltage amplifier stage emitter – leaving the unity-gain output stage fully exposed to rail ripple.

Output ripple disappears, indicating that our wolf is getting in via the voltage amplifier stage or the bottom of the input pair, or both, and the output stage is effectively immune. We can do no more fencing of this kind, for the mirror has to be at the same dc potential as the voltage amplifier stage.

Spice simulation of Fig. 1 with a 1V (0dBV) ac signal on the negative rail gives the rejection ratio curves in **Fig. 4**, with C_{dom} stepped in value. As before there are two regimes, one flat at –50dB, and one rising at 6dB/octave, implying at least two separate injection mechanisms. This suspicion is powerfully reinforced because as C_{dom} is increased, the high-frequency rejection around 100kHz improves to a maximum and then degrades again. This means that there is an optimum value for C_{dom} of about 100pF, indicating some sort of cancellation effect. In the case of the positive rail, the value of C_{dom} made very little difference.

A primary low-frequency ripple injection mechanism is the Early effect in the input-pair transistors. This determines the –50dB low-frequency floor of curve 1 in **Fig. 6**, for the standard input circuit. (This is as per Fig. 4 with C_{dom} at 100pF.)

To remove this effect, a cascode structure can be added to the input stage, as in **Fig. 5**. This holds the V_{ce} of the input pair at a constant 5V, and gives curve 2 in Fig. **6**. The low-frequency floor is now 30dB lower, although high-frequency rejection is slightly worse. The response to the value of C_{dom} is now monotonic; simply a matter of more C_{dom}, less rejection. This is a good indication that one of two partly-cancelling injection mechanisms has been deactivated.

There is a deep subtlety hidden here. It is natural to assume that Early effect in the input pair is changing the signal current fed from the input stage to the voltage amplifier, but it is not so; this current is in fact completely unaltered. What *is* changed is the integrity of the feedback subtraction performed by the input pair; modulating the V_{ce} of $Tr_{1,2}$ causing the output to alter at low frequency by global feedback action. Varying the amount of Early effect in $Tr_{1,2}$ by modifying *VAF* (Early inter-

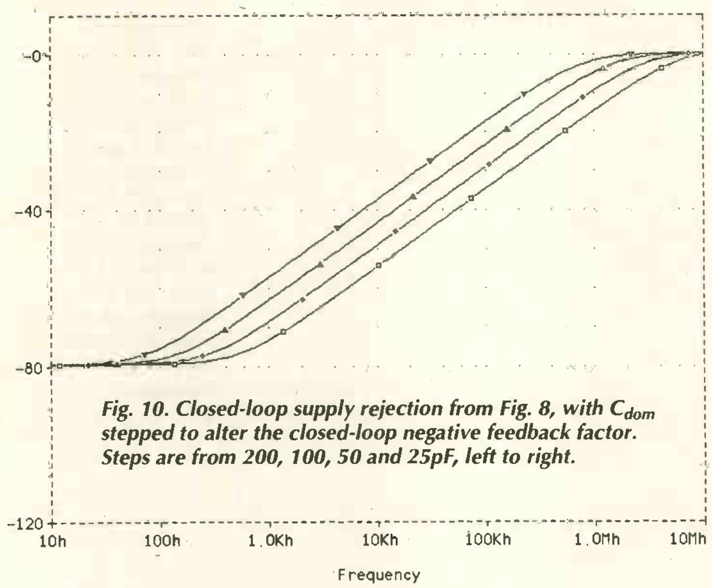


Fig. 10. Closed-loop supply rejection from Fig. 8, with C_{dom} stepped to alter the closed-loop negative feedback factor. Steps are from 200, 100, 50 and 25pF, left to right.

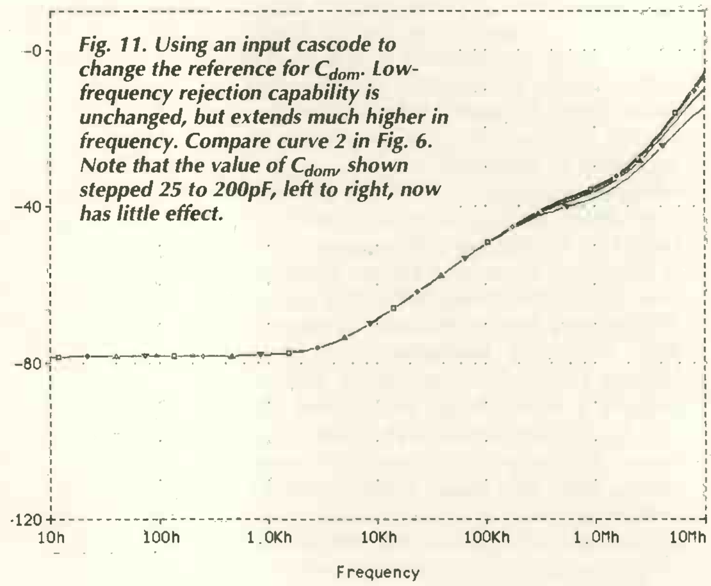


Fig. 11. Using an input cascode to change the reference for C_{dom}. Low-frequency rejection capability is unchanged, but extends much higher in frequency. Compare curve 2 in Fig. 6. Note that the value of C_{dom}, shown stepped 25 to 200pF, left to right, now has little effect.

cept voltage) in the PSpice transistor model alters the floor height for curve 1; the worst injection is with the lowest *VAF* (i.e. V_{ce} has maximum effect on I_c which makes sense).

We still have a low-frequency floor, though it is now at –80 rather than –50dB. Extensive experimentation showed that this is getting in via the collector supply of Tr_{12}, the *VAS* beta-enhancer, modulating V_{ce} and adding a signal to the inner *VAS* loop by Early effect once more. This is easily suppressed by decoupling Tr_{12} collector to the negative rail, and the low-frequency floor drops to about –95 dB, where I think we can leave it for the time being, curve 3 in Fig. 6.

Having peeled two layers from the low frequency rejection ratio onion, something needs to be done about the rising injection with frequency above 100Hz. Looking again at Fig. 1, the *VAS* immediately attracts attention as an entry route.

It is often glibly stated that such stages suffer from ripple fed in directly through C_{dom}, which certainly looks a prime suspect, connected as it is from negative rail to the V_{AS} collector. However, this bold statement is untrue. In simulation it is possible to insert an ideal unity-gain buffer between the voltage amplifier stage collector and C_{dom}, without stability problems, see A_1 in **Fig. 7**, and this prevents all direct signal flow from negative rail to voltage amplifier stage collector through C_{dom}; the rejection ratio is completely unchanged.

Capacitance C_{dom} has been eliminated as a direct conduit for ripple injection, but the rejection ratio remains very sensitive to its value. In fact the negative feedback factor available is the determining factor in suppressing negative-rail ripple-injection, and the two quantities are often numerically equal across the audio band.

The conventional amplifier architecture we are examining inevitably has the voltage amplifier stage sitting on one supply rail; full voltage swing would otherwise be impossible. As a result, the voltage amplifier stage input must be referenced to the negative rail, and it is very likely that this change of reference from ground to the negative rail is the basic source of injection. At first sight, it is hard to work out just what the voltage amplifier stage collector signal *is* referenced to, since this circuit node consists of two transistor collectors facing each other, with nothing to determine where it sits; the answer is that the global negative feedback references it to ground.

Consider an amplifier reduced to the conceptual model in **Fig. 8**. It has a real voltage amplifier stage combined with a perfect transconductance stage *G*, and unity-gain output buffer A_1. The voltage amplifier stage beta-enhancer Tr_{12} proves to have a powerful effect on low-frequency rejection ratio.

To start with, global negative feedback is temporarily removed, and a dc input voltage is critically set to keep the amplifier in the active region – an easy trick in simulation. As frequency increases, local negative feedback through C_{dom} becomes steadily more effective, and the impedance at the voltage amplifier stage collector falls.[5] Therefore the voltage amplifier stage collector becomes more and more closely bound to the ac on the negative rail, until at a sufficiently high frequency of typically 10kHz, the rejection ratio converges on 0dB, and everything on the negative rail couples straight through at unity gain, **Fig. 9**.

There is an extra complication here; the $Tr_{12,4}$ combination actually shows *gain* from the negative rail to the output at low frequencies; this is due to Early effect, mostly in Tr_{12}. If this transistor was omitted, the low-frequency open-loop gain drops to about –6dB.

Reconnecting the global negative feedback, **Fig. 10**, shows a good emulation of supply rejection for the complete amplifier in Fig. 6. The 10-15dB open loop rise is flattened out by the global negative feedback, and no trace of it can be seen in Fig. 10.

Now the feedback attempts to determine the amplifier output via the voltage-amplifier collector, and if this control was perfect the rejection ratio would be infinite. It is not, because the negative feedback factor is finite, and falls with rising frequency, so rejection ratio deteriorates at exactly the same rate as the open-loop gain falls. This can be seen on many op-amp specification sheets, where negative-rail rejection falls off from the dominant-pole frequency, assuming conventional op-amp design with a voltage amplifier on the negative side.

Clearly a high global negative feedback factor at low frequency is vital to keep out negative-rail disturbances. In reference 5, I rather tendentiously suggested that apparent open-loop bandwidth could be extended quite remarkably, without changing the amount of feedback at high-frequency, where it matters. This is done by reducing low-frequency loop gain. A high-value resistor R_{nfb} in parallel with C_{dom} does the trick. What I did not say was that a high global negative feedback factor at low frequencies is also invaluable for keeping hum out – a point overlooked by those advocating low negative feedback as a matter of faith rather than reason. **Table 2** shows how reducing global feedback by decreasing the value of R_{nfb} degraded ripple rejection in a real amplifier.

Table 2. Reducing global feedback by decreasing the value of R_{nfb} degraded ripple rejection in a real amplifier.

R_{nfb}	Ripple out
None	–83.3dBu
470kΩ	–85.0dBu
200kΩ	–80.1dBu
100kΩ	–73.9dBu

Having understood the negative-rail rejection ratio mechanism, in a just world our reward would be a new and elegant way of preventing such ripple injection. Such a method indeed exists, though I believe it has never been applied to power amplifiers.[6,7] The trick is to change the reference, as far as

C_{dom} is concerned, to ground. Figure 5 shows that cascode-compensation can be implemented simply by connecting C_{dom} to point B rather than the usual voltage amplifier base connection at A. **Figure 11** demonstrates that this is effective, rejection ratio at 1kHz improving by about 20dB.

Elegant or not, the simplest way to reduce ripple below the noise floor still seems to be brute-force *RC* filtering of the negative supply to the input mirror and voltage amplifier, removing the disturbances before they enter. It may be crude, but it is effective, as shown in **Fig. 12**. Good low-frequency rejection requires a large *RC* time-constant, and the response at dc is naturally unimproved. The real snag is that the necessary voltage drop across *R* directly reduces amplifier output swing. Since the magic number of watts depends on voltage squared, it can make a surprising difference to the raw commercial numbers – though not, of course, to perceived loudness.

With the circuit values shown, 10Ω is about the maximum tolerable value; even this gives a measurable reduction in output. The accompanying *C* should be at least 220μF, and a higher value is desirable if every trace of ripple is to be removed. ■

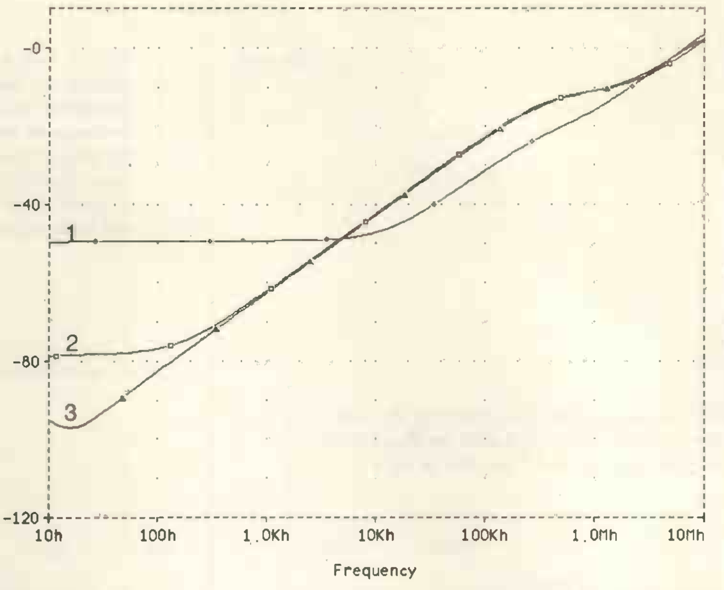


Fig. 12. RC filtering of the negative rail is effective at medium frequencies, but less so at low frequencies, even with 1000μF of filtering. Resistance R is 10Ω. Top to bottom, curves show 0, 10, 100 and 1000μF decoupling on V–.

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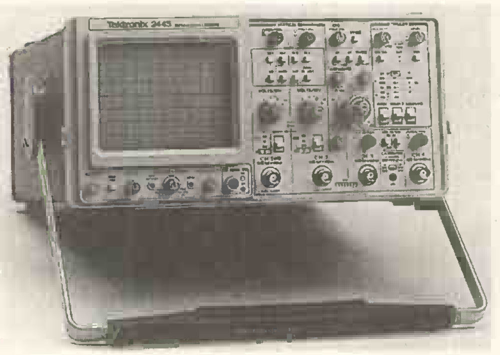

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Winning designs

Three winning power control circuits representing the best of many excellent entries submitted for the International Rectifier design competition.

These circuits are three of the six winners of the International Rectifier design competition.

Three further winners are MS Nagaraj with a power inverter, Clyve Caines, with his electroluminescent lamp driver, and Neoklis Kyriazis, who submitted a 1-to-3-phase motor converter. We hope to publish these designs in the near future.

If you submitted a design and are not among the winners, don't despair. Your design will automatically be submitted for consideration in our circuit ideas section.

Each of the winners of the International Rectifier design competition will receive one of six prizes worth from £100 to £750. The prizes are electronics engineer's design kits comprising semiconductors selected from the large IR range, together with storage facilities.

First prize

Power modulator

By modulating the duty cycle of the *IR2151*'s oscillator, it is possible to vary the power output delivered into any load by the output pair.

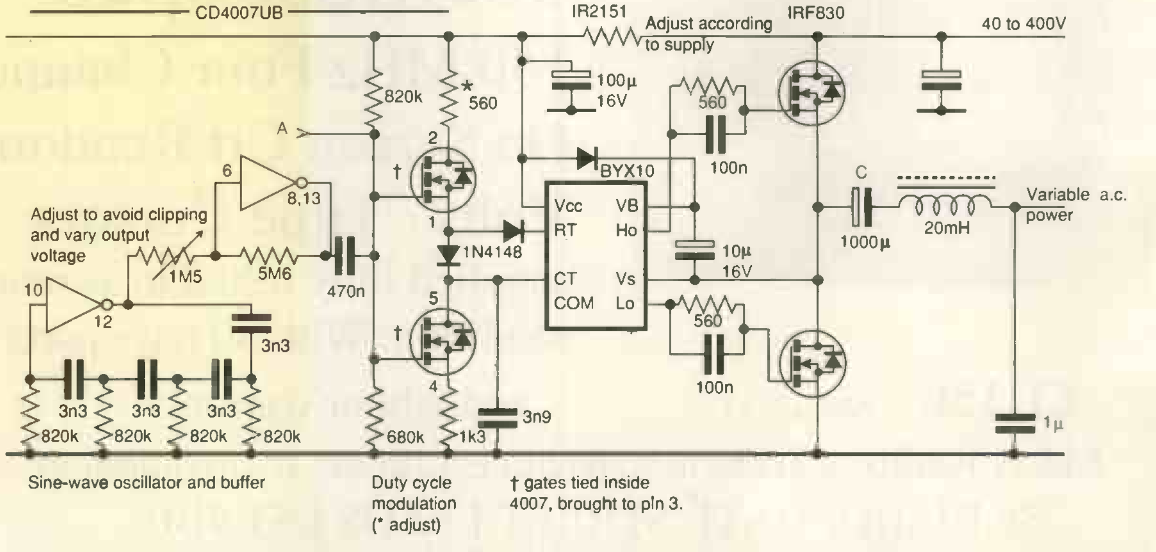
Setting the duty cycle to a little less than 50% and modulating up to a maximum of 50% on peaks, it is possible to transmit audio-frequency information via fluorescent tubes driven by this circuit. This opens up the possibility for a wireless and rf-free paging or communication system in a building, without the need for expensive power modulators.

With the output fed from a lower level of 45V or so, it is possible to produce a class-D-type amplifier capable of driving a loudspeaker. The 20mH inductor helps demodulation.

Being able to modulate the output also opens up the possibility of producing a kind of solid-state variac capable of delivering variable ac. Output could be up to about $^1/_3$ of the output-stage supply voltage, and frequency could be set to any value within the limits of the device. One potential application is testing 400Hz aircraft equipment.

For simplicity, the additional circuitry has been designed around a *4007* unbuffered cmos logic part.

A. Zeimacki
Rotherham
Yorkshire



Power modulator has many uses – from inter-building wireless signalling to loudspeaker driving. For af communications applications, omit the oscillator/buffer and feed audio/data at A. Where low-frequency power is needed, use a large capacitor on the output.

Second prize
Switch-mode power supply

In this application, the typical half bridge circuit around the *IR2151* is not loaded with a fluorescent lamp, but with a transformer. The secondary of the transformer is rectified and filtered, giving output of 27.6V at 3.25A, which equates to around 90W. The voltage chosen for the circuit shown is that of a 24V nominal lead-acid battery, but it can of course be varied.

Circuitry controlling the output voltage is a typical *TL431* application. The optocoupler however is a dual type with bipolar output connection. Control of the output voltage is possible by variation of the switching frequency.

The single *IR2151*'s timing resistance comprises two 3kΩ fixed resistors, potentiometer P_1 and the outputs of the optocouplers. Minimum switching frequency is adjusted via P_1 just above the resonant frequency of the resonant circuit formed by L_1/C_2. Resonant frequency of this circuit is about 50kHz.

At full load, switching frequency is about 57kHz. Current in L_1 is very close to sinusoidal, and switching of the mosfets is close to zero current. Snubber capacitor C_1 gives further improvement. Switching losses are very low. Full load efficiency is about 89%. Signals occurring at full load are shown in the left-hand plot.

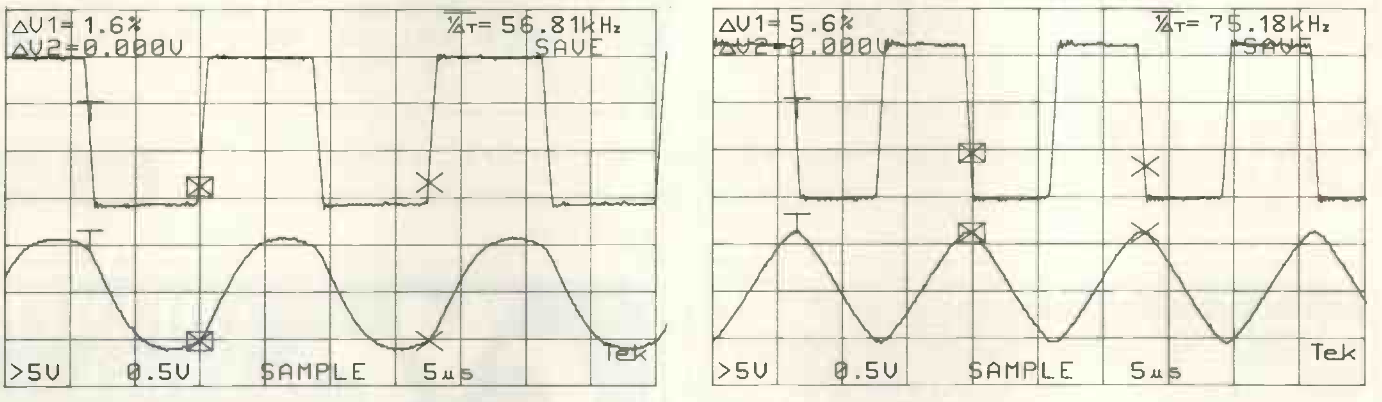
At no load the switching frequency is increased by the control circuit to about 75kHz. Switching of the mosfets is at maximum current, but because of snubber C_1 switching losses remain low. No load input power is about 5W, and signals occurring are shown in the right-hand plot.

Both at full load and at no load, resonant current is about the same, i.e. around 2.5A pk-pk). At about half load, 45W the resonant current is at its maximum of about 3A.

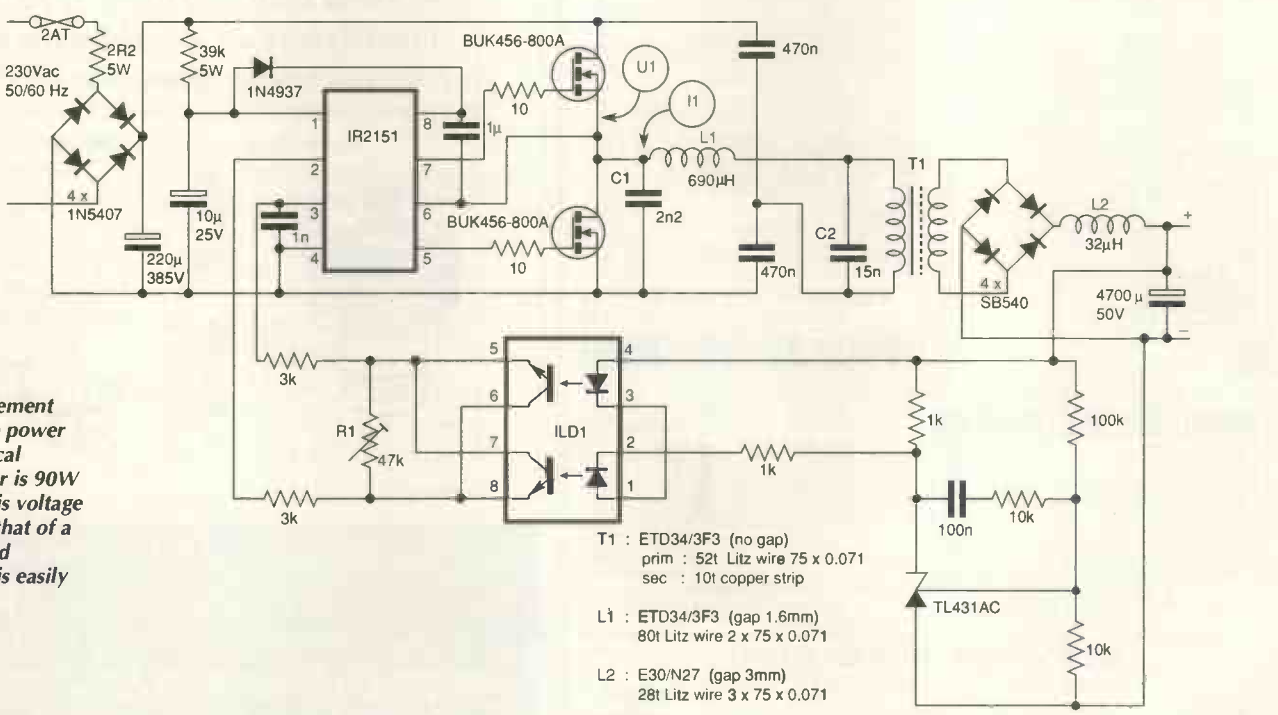
Short circuiting the output brings the switching frequency to its minimum of 57kHz. In this case inductor L_1 gives a very efficient current limit. Inductor current in L_1 is about 2A pk-pk. Short-circuit output current is typically less than 4A dc while full load output current is typically 3.25A dc.

Because of the soft switching of the mosfets and the low harmonics, generated EMI is low, compared to other topologies. Not that not all components are ideal; their selection was based mainly on availability.

P. Van der Pol
Alblasserdam
The Netherlands



Plots representing signals produced by the easy-to-implement switch-mode supply. Plot on the left shows 90W full-load conditions, while the one on the right represents no-load conditions. Both plots depict U1 on the top trace and I1 on the bottom. Scales are 100V/div for U1 and 1A/div for I1.



Easy-to-implement switch-mode power supply. Typical output power is 90W at 27.6V. This voltage is typical of that of a 24V lead-acid battery, but is easily altered.

Third prize

Single-phase speed control for induction motors

Electronic variable speed control for induction motors can be implemented with the simple inverter shown in **Fig. 1**. Using an inverter, the motor can be made to run at speeds both above and below the fixed speed obtained by driving the motor directly from mains frequency.

Section a of the *LM324* op-amp splits the low-voltage supply rail to produce a mid point. Potentiometer Vr_1 is the speed control. Components R_3 and C_3 form the starting circuit, providing gradual acceleration.

Sections b and c of the op-amp form a voltage-controlled oscillator while section c produces a square wave output. Section b integrates the current into pin 5. This is positive or negative depending on the output of the schmitt trigger, section c.

The triangle at the output of section b is shaped by R_{11} and diodes $D_{4,5}$ produce an approximate sine wave. Frequency with the components shown goes up to about 70Hz.

Op-amp section d acts as a modulator of the duty cycle of the oscillator within the *IR2151*. Modulation of the *IR2151* is sensitive to component tolerances and the zener voltage, so negative feedback is used to stabilise the output. The duty cycle at the RT output on pin 2 is filtered with R_{14} and C_7 and fed back to section d of the *LM324* via R_{13}.

Current fed into the timing capacitor C_8 changes the charge and discharge ramps and modulates the duty cycle. Capacitor C_6 and R_{20} are fitted for stability.

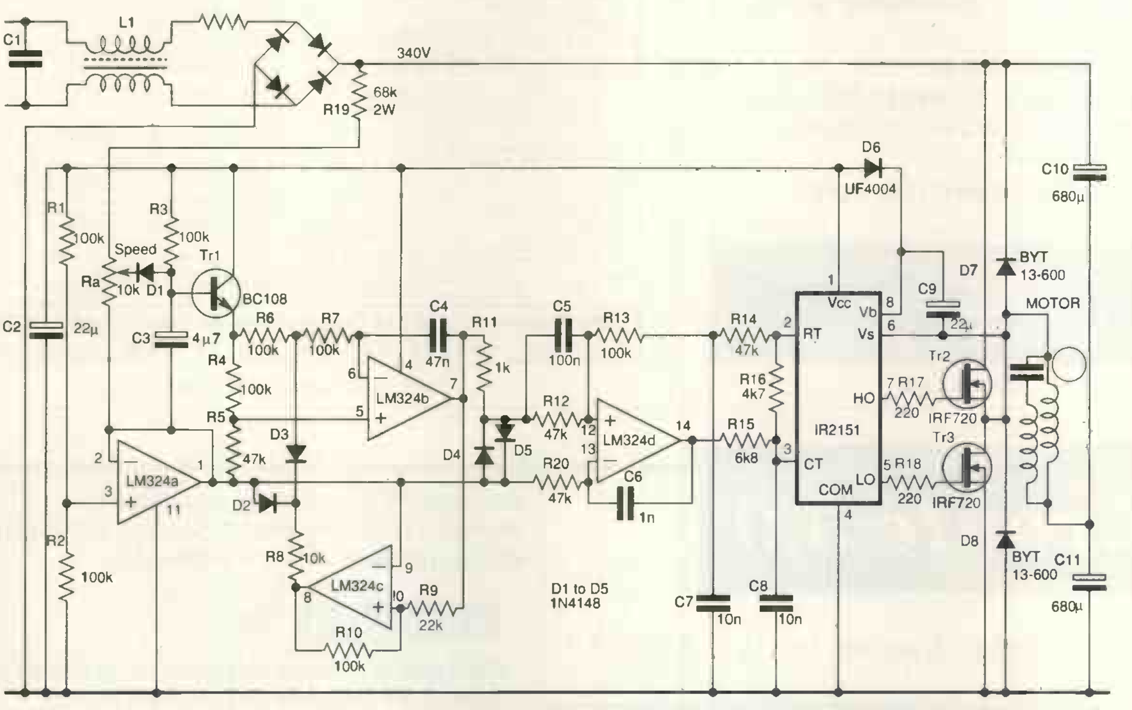
Induction motors need a voltage drive increasing with frequency. Capacitor C_5 provides this function with a linear voltage frequency slope. Resistor R_{12} gives a voltage at low speed to boost the motor starting. These components will require change for different motors and loads. Resistor R_{13} set the output, R_{12} sets the low speed boost and C_4 the frequency range of CO.

Series resistors R_{17} and R_{18} determine the switching speed of the output fets. Too high a value will give slow switching while too small a value pulls a high peak current through diodes $D_{7,8}$, creating interference.

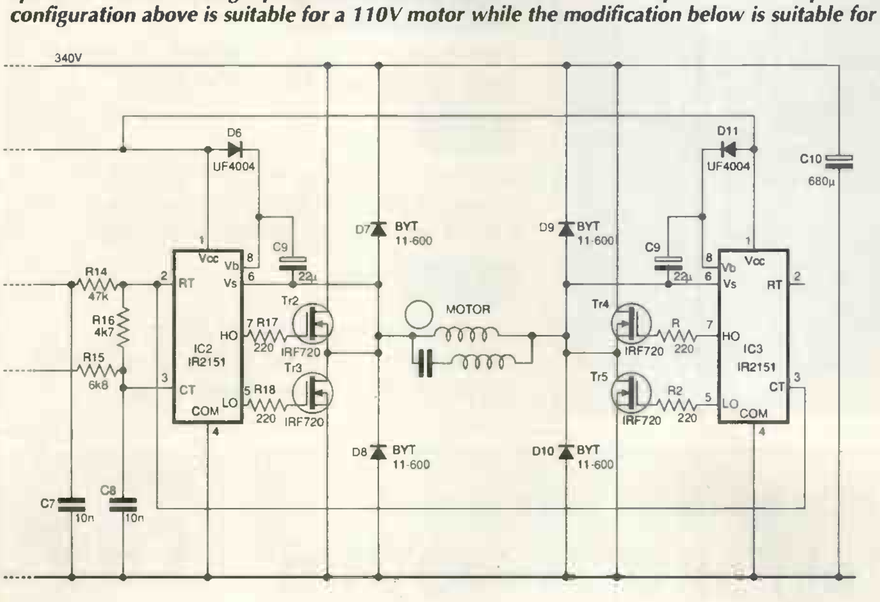
As shown, the circuit is ideal for driving a 110V capacitor motor: peak rms output voltage is 120V. A 240V motor will rotate on this but if 240V output is required a second *IR2151* needs to be added. This can be driven from the RT output of the first *IR2151* as there is a phase inversion between RT and CT.

As well as working from rectified mains, the circuit can be powered from a low voltage and drive an induction motor via a step up transformer. In this case the fets should be lower voltage, higher current types, such as the *IRF224*, and the values of $C_{10,11}$ should be increased by a factor of say 1 to cope with the higher ripple current.

It may be possible to dispense with the diodes $D_{7,8}$ as the diodes within the IR mosfets are characterised. This will increase dissipation and a higher value of R_{17} and R_{18} may be required. Resistor R_{19} will need reducing to allow for the lower supply voltage. ■



Speed controller for single-phase induction motors has soft start. Components shown produce up to 70Hz. The configuration above is suitable for a 110V motor while the modification below is suitable for 240V motors.



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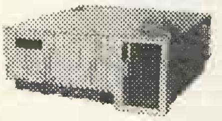

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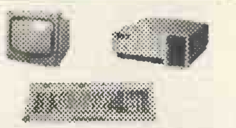

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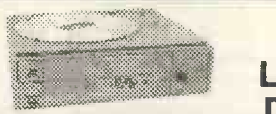

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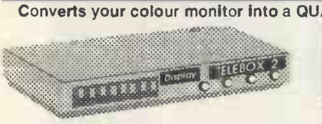

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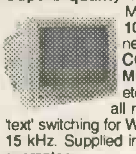

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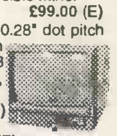

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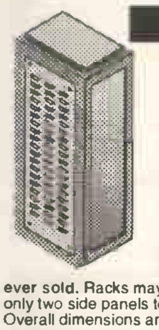

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CIRCLE NO. 123 ON REPLY CARD

SINEWAVES step by step

Ian Hickman explains the fundamentals of sine waves across series LCR circuits

Following one of my earlier articles, Peter Dawe wrote in '*Sine Post*', *Letters, EW+WW*, February, pp.149, with a query about the initiation of sinusoidal waveforms in a simple series *L*, *C*, *r* circuit, by the application of a step function of voltage to the circuit. Why should the voltage waveforms developed across the 'lossy' inductor and the 'ideal' capacitor always be sinusoidal?

He goes on, in a sense, to answer his own question by acknowledging that sinewaves drop out in the solution of a second order differential equation describing the system. But he seeks a more intuitive insight based on the physical properties of inductors and capacitors. It would be based on the inductor property of terminal voltage proportional to the rate of change of flux linkage and capacitors property of terminal current proportional to the rate of change of stored charge.

Such an intuitive explanation is indeed possible, as will appear in what follows, though some readers may point out that it simply amounts to solving the differential equations by stealth. And to make the explanation simpler, the series circuit shown in **Fig. 1.** assumes that *r*, the resistance of the inductor is negligible and its Q infinite – an assumption which is also applied to the capacitor. Furthermore, the transient has been applied not as a voltage step function in series with the circuit, but as a unit current impulse or delta function in parallel with it, again in the interests of keeping everything as simple as possible for the purposes of explanation.

After all, both a step function and a delta function are bony, angular, non-repetitive shapes and about as far as one can get from a repetitive curvaceous sinewave.

Generator *G* in Fig. **1a**, with its 1000MΩ internal resistance, provides an almost perfect delta function, dumping a current of 1×10^9 amps into whatever circuit it is connected to, for a period of a nanosecond. Thus the delivered charge is one coulomb, assuming the voltage in the target circuit remains negligible compared to 1×10^{18} volts – not a difficult condition to meet. (The circuit values may seem a little impractical, but for the purposes of explanation, the theorist has the advantage over the practical man that he can chose whatever values he likes. Furthermore, his circuits always work, serenely untroubled by parasitics or other unforeseen contingencies, but unfortunately only on paper.)

As our interest centres on the response of the circuit to the transient stimulation – rather than on the transient itself, let t=0 be the moment immediately following the delta function, when the generator voltage has just returned to zero. During the preceding one nanosecond it will have dumped a charge of one coulomb into capacitor *C*, whose terminal voltage will have risen linearly over that nanosecond from zero to 1V.

Left connected, the generator would then draw a current, albeit tiny, from the capacitor so switch *S* has been included in the circuit.

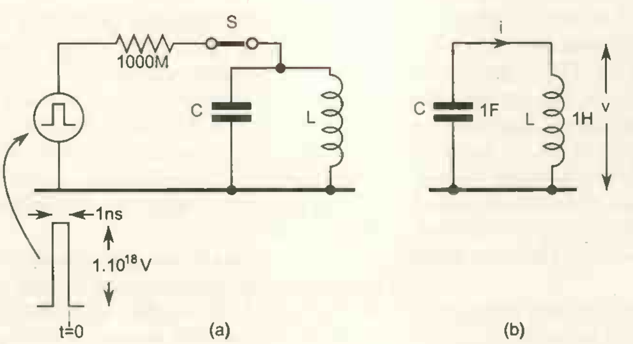


Fig. 1a. LC circuit with delta function excitation. In b), i and v represent voltage and current at time t=0 and after, see Fig. 2.

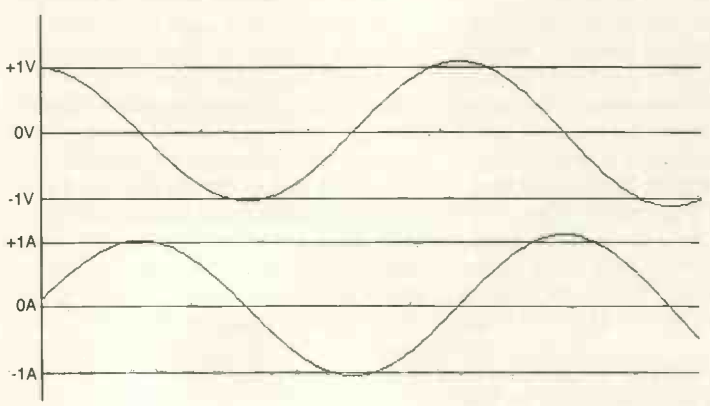


Fig. 2. Behaviour of the circuit of Fig. 1b according to the program listing on the next page.

```
100 REM ww_sine.bas
110 SCREEN 2: REM selects turbobas screen
120 WINDOW (0,399)-(399,0) REM redefines window to read l to r, bottom to top
121 h=2: REM sets horizontal step size
122 u=300: REM sets vertical position for voltage display
123 w=100: REM sets vertical position for current display
124 s=50: REM sets display vertical scale
126 i1=0: v1=1: REM initialises current i Amps and voltage v Volts
127 t=0.05: REM sets time step to 50ms
130 FOR j=0 TO 199
140 i2=i1+v1*t: REM calculates new value of current
160 v2=v1-(i1+i2)/2*t: REM adjusts capacitor v for average i drawn off
180 LINE (h*j, u+s*v1)-(h*(j+1), u+s*v2): REM plots lines joining present vals
185 LINE (h*j, w+s*i1)-(h*(j+1), w+s*i2): REM of v and i to previous ones
190 i1=i2: REM redefines present new value of current as next old value
191 v1=v2: REM as above, for voltage
200 NEXT
210 REM draw scale lines
220 LINE (0, 0)-(0, 399)
250 LINE (0, u)-(399, u): LINE (0, w)-(399, w)
280 LINE (0, u+50)-(399, u+50): LINE (0, w+50)-(399, w+50)
300 LINE (0, u-50)-(399, u-50): LINE (0, w-50)-(399, w-50)
```

Listing 1. **Turbobasic** ***program to plot*** **LC** ***response to a delta function (unit impulse). It should run under*** **Turbobasic** ***on almost any pc XT/AT***

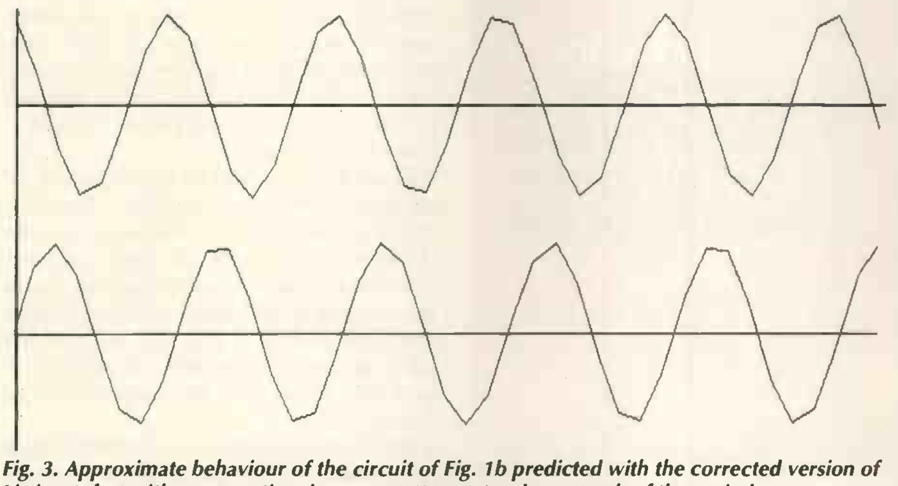
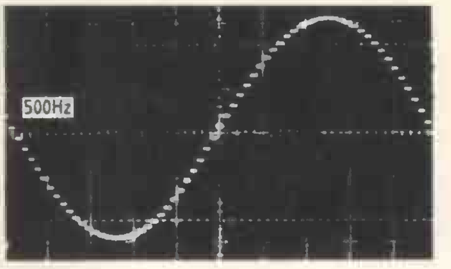
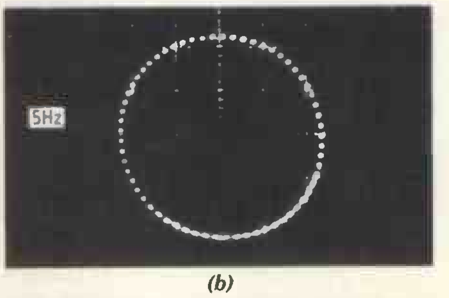


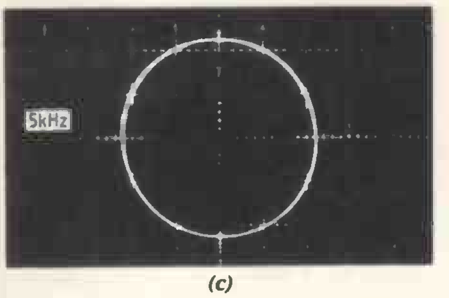
Fig. 3. Approximate behaviour of the circuit of Fig. 1b predicted with the corrected version of Listing 1, but with a coarse time increment Δt, greater than a tenth of the period.



(a)



(b)



(c)

Fig. 4. (a) 500Hz sinewave output of quadrature oscillator in ref. 2. Photographs (b) and (c) are 5Hz and 5kHz Lissajous figures produced by the in-phase and quadrature outputs of the oscillator.

We can open this at our leisure just after t=0. The other point to clarify concerns inductor L at t=0. Voltage across it is 1V, so the rate of increase of current through it is one amp per second. But at t=0, the current itself can be assumed with negligible error to be zero, despite the fact that there was some voltage across it during the one nanosecond period preceding t=0. This is because the impressed volt-second product was negligible, see panel on page 260.

All these seemingly nit-picking preliminaries are necessary not only to satisfy the rigour of the mathematical purists, but also for the more important practical reason of ensuring that we get the right answers. So at t=0, there is zero current in the inductor and the instantaneous voltage v across the capacitor is 1V. Voltage across the inductor is proportional to the rate of change of flux linkage, and here – unlike the case of a generator – there is no externally applied flux. So the flux is simply proportional, at every instant, to the current through the inductor. Hence the rate of increase of current i in Fig. **1b** at t=0 is one amp per second.

As i increases, it will draw charge from the capacitor, resulting in a reduction of the voltage across the capacitor, according to the equations in the panel. If the voltage across the capacitor is regarded as the cause and the current through the inductor as the effect, then here, the effect is seen in its turn directly to affect the cause. This makes it difficult to see what is going on, which is why methods for solving differential equations were invented in the first place.

However, you can see approximately what is going on by imagining the voltage across the capacitor held constant for a short period Δt while you see what happens to the current in the inductor. Now make a correction to the capacitor voltage to allow for the charge drawn off during that instant.

The *Turbobasic* program shown in **Listing 1** does just this, and should be largely self explanatory. It calculates out and plots the circuit voltage and current as shown in **Fig. 2.** It shows the voltage and current indeed to be sinusoidal, and given that the 199 steps of 50ms shown correspond to just over 1.5 cycles, this ties up with the 0.159Hz predicted as the resonant frequency by the formula,

$$f=\frac{1}{2\pi\sqrt{LC}}$$

But the gradually increasing amplitude indicates a circuit with a Q of greater than infinity: there must be something wrong somewhere! Line 140 calculates the increase in the currrent over a short period Δt. It assumes that capacitor voltage remains constant the whiles and line 160 then adjusts the capacitor voltage, assuming the charge drawn off during Δt is equal to the average current during that period times Δt. Using average current like this would be acceptable if, to be fair, average voltage $(v_1 + v_2)/2$ had been used in line 140, instead of

Continued on page 260...

the sign of IC_1's output ramp. The Schottky diodes are bound to IC_5 non inverting input, ensuring its clean recovery from overdrive.

Sine of the times

The *AD639* trigonometric function generator, biased via IC_4, converts IC_1's triangular output into a sine wave, trace d. To avoid output distortion, you must supply the *AD639* with a triangular wave that does not vary in amplitude. At higher frequencies, delays in the IC_1 integrator switching loop result in late turn-on and turn-off of Q_1. Unless you minimise these delays, the triangle amplitude will increase with frequency and cause the distortion level to increase. IC_5, the Q_4/Q_5 level shifter, and Q_1 generate a total delay of 14ns. This small delay, combined with the 22pF feed-forward network at IC_5's input, keeps distortion to just 0.40% over the entire 1MHz range. At 100kHz, the distortion is typically less than 0.2%. The 8pF capacitor in Q_1's source line minimises the effects of gate-source charge transfer, which occurs whenever Q_1 switches. Without this capacitor, a sharp spike would occur at the triangle peaks, increasing distortion. Fets Q_2 and Q_3 compensate for the temperature-dependent on-resistance of Q_1 and keep the $+2I/-I$ relationship constant with temperature.

This circuit responds very rapidly to input changes - something most sine-wave generators cannot do. Figure 8b trace A shows what happens when the input switches between two levels. Op-amp IC_1's triangle output in trace b shifts frequency immediately, with no glitches or poor dynamics. Sine output, trace C, reflecting this action, is similarly clean.

To adjust this circuit, apply 10.00V and trim the 100Ω potentiometer for a symmetrical triangle output at IC_1. Next, apply 100μV and trim the 100kΩ potentiometer for triangle symmetry. Then, apply 10.00V again and trim the 1kΩ frequency-trim adjustment for a 1MHz. Finally, adjust the distortion-trim potentiometers for minimum distortion as measured on a distortion analyser, Fig. 8c trace E. You may have to re-adjust the other potentiometers slightly to achieve the lowest possible distortion. If you won't operate the circuit below 100Hz, you can delete the IC_2 basic-dc stabilisation state. If you make this change, you should ground the positive input of op-amp IC_1.

Many of the filter and oscillator circuits presented here are simple as well as useful. Their simplicity shows that clever circuit designers often take a minimalist approach. When you speak or write, you are more likely to get your point across if you use short words that are familiar to your audience. So it is with circuits. The simplest design that does the job usually costs the least and operates more reliably than complex alternatives ■

QUAD SPEED RS232

Sending more data down a serial data channel normally involves an increase in bandwidth. Guruprasanna and Lanka Kumar show that psk provides a way to achieve four times the bit rate without increasing the baud figure.

The aim of the design detailed here is to provide increased transmission rate for a given channel bandwidth when binary data is being transmitted. The method is based on 16-ary phase-shift keying, psk, where a phase shift of multiple of 22.5° is produced corresponding to every block of four bits. Thus the baud rate is a quarter of the bit rate. The reference phase is also transmitted with the psk output to provide coherent detection.

Bit rate versus baud rate

In data communication, digital data needs to be transmitted in binary form between one end of the RS232 link, called DTE, and the other end, designated DCE. These abbreviations represent data terminating equipment and data circuit-terminating equipment.

Distance between DTE and DCE is limited by noise introduced, and by the ensuing distortion over the propagation line. This, along with channel bandwidth, puts a limitation on the bit rate of base-band transmission.

With multiple psk, the baud rate is less than the bit rate. So, for a given distance between the transmitter and receiver, the bit rate can be increased. In 16-ary psk, the bit rate can be raised by a factor of four for a given transmitter-receiver distance, and for a given channel bandwidth, compared to binary psk.

RS-232C is the standard connector used between the DTE and DCE to transmit binary data. Transmission rate is increased by adding a 16-ary psk transmitter in the DTE and 16-ary psk receiver at the DCE end, **Fig. 1**. Output of the DTE connects to the 'transmit' pin at one end of the link, and to the psk receiver following the receiver pin at the other end, in the DCE.

Voltage level shifters are used at both ends, so as to conform to RS-232C specifications. Thus, RS-232C ,which permits a maximum of 20kbit/s can be made to carry 80kbit/s over the permissible distance of 15m.

Unassigned pins of RS-232C can be used along with correct voltage level shifting to transmit various control signals that are required for coherent detection.

Transmitter details

The block diagram of the transmitter is given in **Fig. 2**. If a clock is available, as in a DTE, then this can be multiplied by two using a pll, and used as the main clock. The serial bits come in at the rate of the clock pulses of the DTE. These pulses feed to a serial-to-parallel shift register, **Fig. 3**. A latch pulse is given to the four bit latch, whose inputs are the shift register outputs. The latch pulse is generated after every four bit period so that the four bit word is latched periodically.

Outputs of the latch feed the 16:1 multiplexer-select lines. This multiplexer is in turn fed with the 16 carrier phases. The main clock, which is twice the frequency of the bit

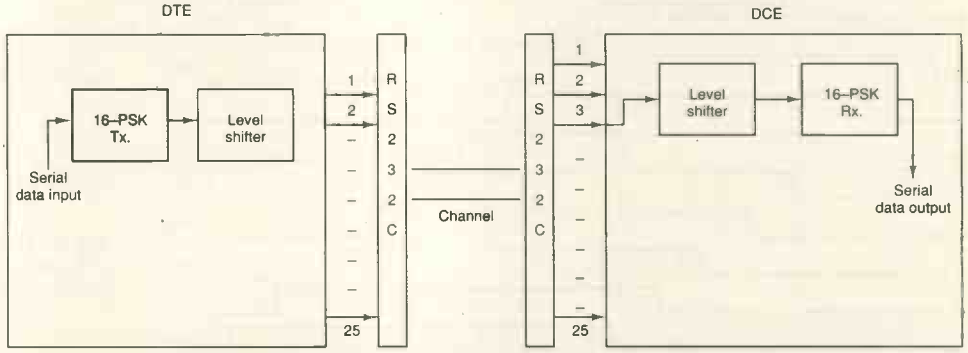


Fig. 1. Using 16-ary psk and spare pins on the 25 way D type connector, four times the baud rate can be achieved for a serial link.

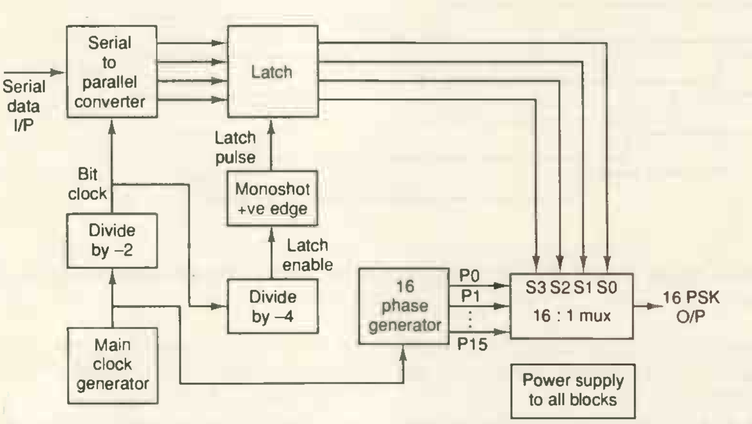


Fig. 2. In the psk transmitter, a 16:1 multiplexer switches in one of 16 different carrier phases, coded from the serial data.

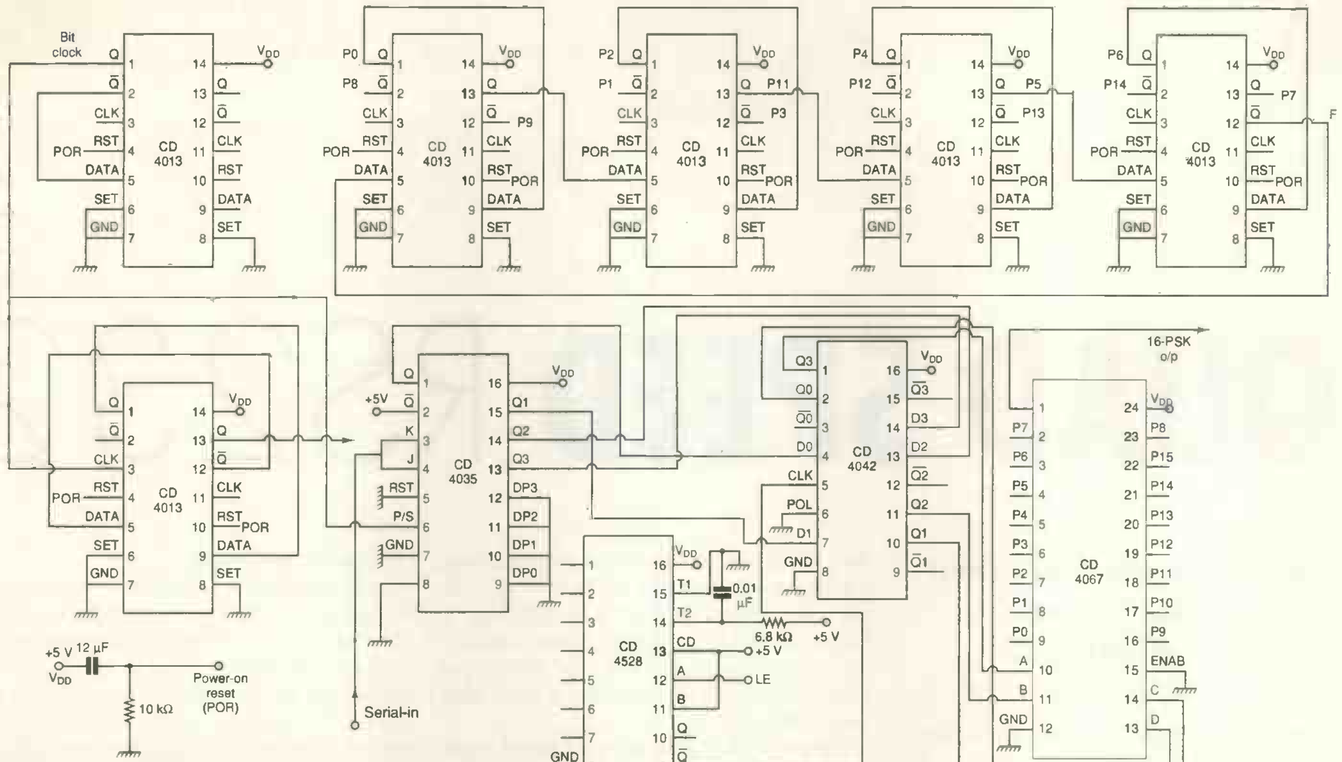


Fig. 3. Serial data is shifted into the register and periodically latched to provide the multiplexer select lines.

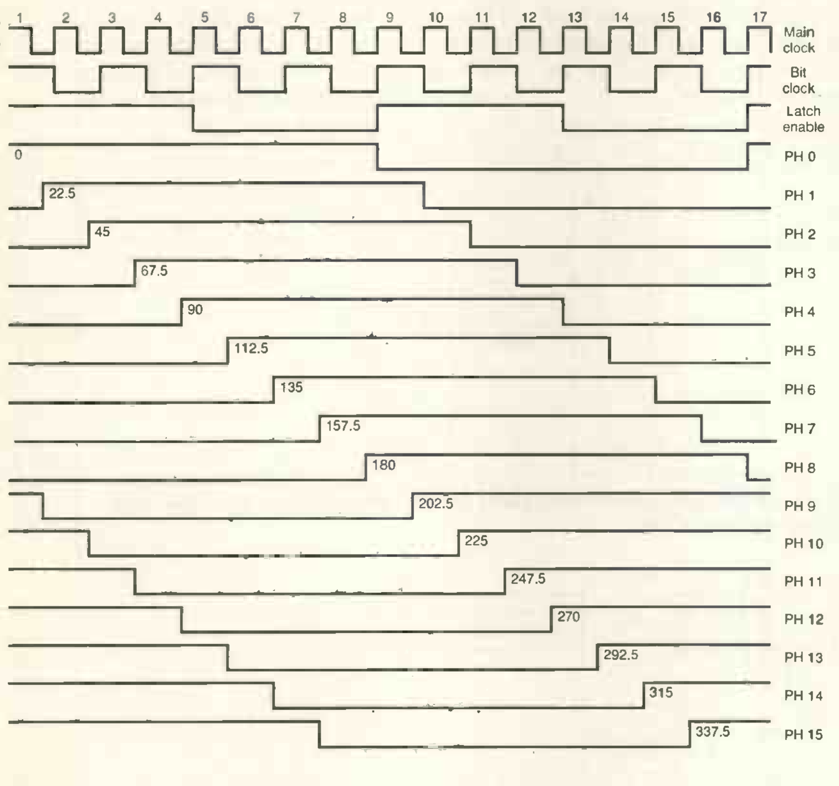


Fig. 4. Phase signals separated by 22.5° are seen in this timing diagram synchronised to the main clock.

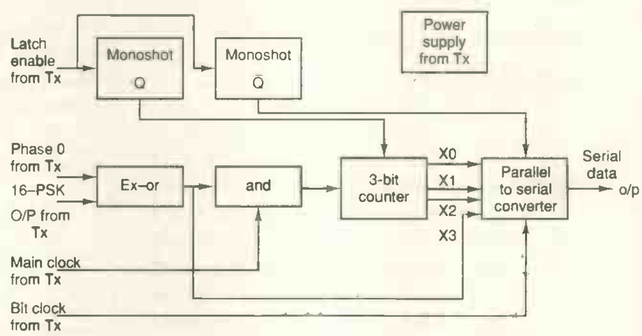


Fig. 5. At the receiving end, after level shifting, exclusive-OR gates combine the psk output and the reference phase to extract the original data.

clock from the DTE, is divided by a factor of 16 using 8 D-type bistable devices in the twisted ring mode.

Figure 4 shows the outputs of the bistable devices. In this circuit, we have assumed that the main clock is available, and this divided by two is taken as the bit clock. The 16 phases are fed to the multiplexer inputs. Thus, depending on the four-bit word stored at the latch outputs one particular phase out of the 16 is output by the multiplexer for a period of four bit-clock cycles.

The carrier is transmitted for a duration equal to half its period, which is sufficient to enable the receiver to detect the phase. The latch holds the select lines at the same combination for this duration and the psk output is then obtained from the multiplexer.

The 16-psk output along with the reference phase is sent to the receiver. An unassigned pin of the connector is used to transmit the reference phase and the 16-psk output is transmitted on the 'transmit' pin. Voltage level shifters are used when there is a need for interface between the psk circuit and connector.

The receiver generates the various other signals, for example the main clock which is derived from *reference phase× 16 bit clock* and the latch enable waveform which is equal to *bit clock/4*.

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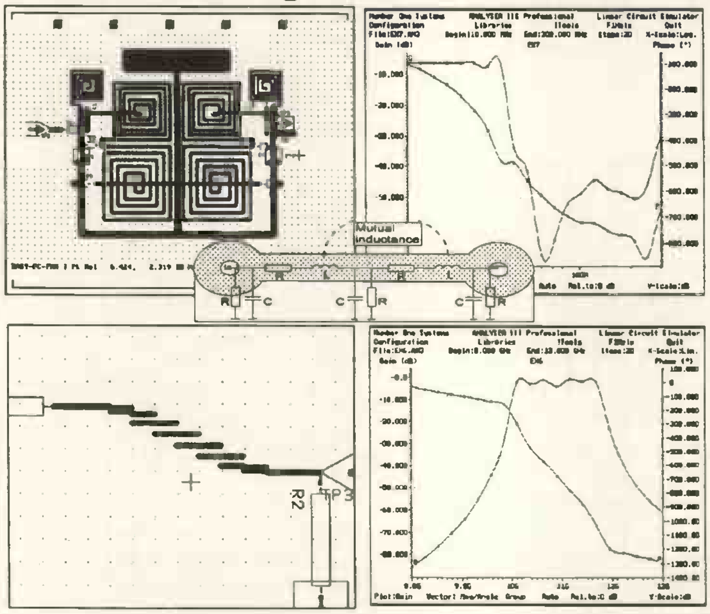

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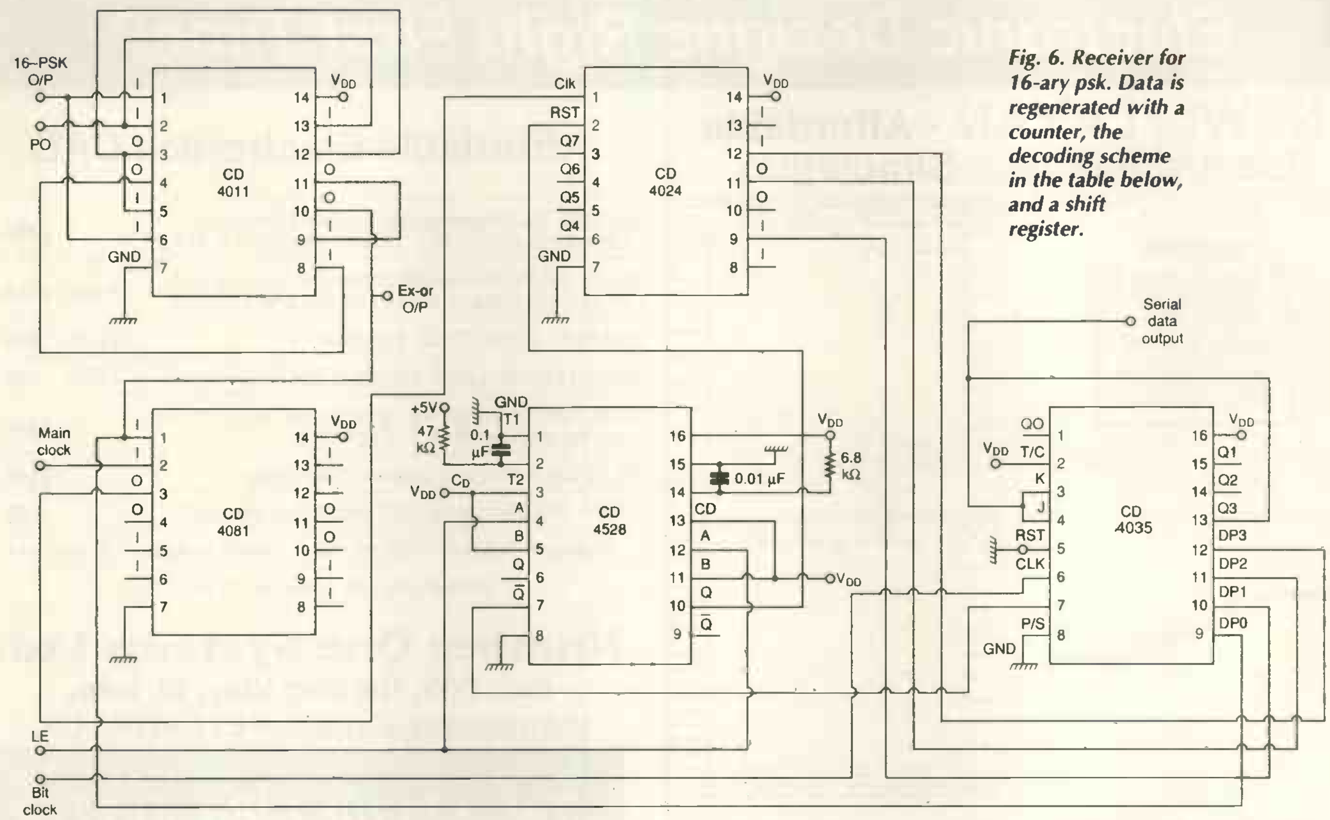

Fig. 6. Receiver for 16-ary psk. Data is regenerated with a counter, the decoding scheme in the table below, and a shift register.

Table 1. Bit pattern generated at the output of the exclusive-OR gate.

phase degrees	Y_7	Y_6	Y_5	Y_4	Y_3	Y_2	Y_1	Y_0	X_3	X_2	X_1	X_0
	bits at the exclusive-OR output								decoded message			
0	0	0	0	0	0	0	0	0	0	0	0	0
22.5	1	0	0	0	0	0	0	0	0	0	0	1
45	1	1	0	0	0	0	0	0	0	0	1	0
67.5	1	1	1	0	0	0	0	0	0	0	1	1
90	1	1	1	1	0	0	0	0	0	1	0	0
112.5	1	1	1	1	1	0	0	0	0	1	0	1
145	1	1	1	1	1	1	0	0	0	1	1	0
167.5	1	1	1	1	1	1	1	0	0	1	1	1
180	1	1	1	1	1	1	1	1	1	0	0	0
202.5	0	1	1	1	1	1	1	1	1	1	1	1
225	0	0	1	1	1	1	1	1	1	1	1	0
247.5	0	0	0	1	1	1	1	1	1	1	0	1
270	0	0	0	0	1	1	1	1	1	1	0	0
292.5	0	0	0	0	0	1	1	1	1	0	1	1
315	0	0	0	0	0	0	1	1	1	0	1	0
337.5	0	0	0	0	0	0	0	1	1	0	0	1

Receiver details

Figures 5 and **6** show the receiver block and circuit diagrams. The psk output is exclusive-ORed with the reference phase and this is ANDed with the main clock to get 8bit values of the exclusive-OR output. The AND gate converts the continuous waveform into pulses. These pulses are fed to a three bit counter.

At the end of eight main clock cycles, counter outputs contain three of the four transmitted bits. The fourth is the same as the last bit of the exclusive-OR output. **Table 1** gives the bit pattern for each transmitted phase and the corresponding four-bit combination which selects that particular phase. The decoding scheme using the counter and the last exclusive-OR bit is obtained by the inspection of this table. The four bits thus obtained are fed to the parallel inputs of the shift register which loads the bits once every eight main clock cycles. Once loaded, the bits are shifted out at the bit clock rate to eventually become the input bit stream at the transmitter.

In this circuit, the main clock is divided by two and this is taken as the bit clock. This implies that the transmitted phase has to be held for a duration of four bit periods. If the bit clock is the same as the main clock, then the transmitted carrier phase has to be held for a duration equal to eight bit clock cycles. This requires increased hardware for coherent detection because the transmitter has to store the incoming four bits for a period of four clock cycles, after transmitting the carrier phase corresponding to the preceding four bits.

Waveforms at the various positions in the transmitter and receiver for an input bit stream are given in **Fig. 7**.

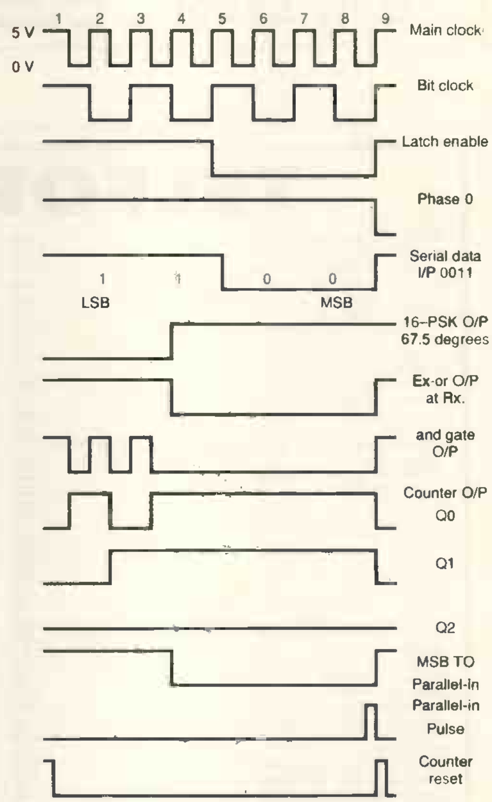

Fig. 7. Timing of the transmitter and receiver together shows the transmitted phase held for four bit clock cycles.

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Digital filters adapt

Adaptive digital filters are invaluable for maximising signal-to-noise performance in a variable electronic environment Allen Brown discusses their implementation.

An important application for digital signal processing is adaptive filters. Digital adaptive filters can be very effective for eliminating unwanted signals or noise. As a result they are finding numerous uses as adaptive line equalisers, in noise cancellation systems and for echo reduction in telecomms systems.

Digital adaptive filters are by no means restricted to telecomms systems, but the thrust of their development has been closely allied to it. Availability of the digital signal processor, dsps, has also contributed greatly to the widespread use of adaptive fitters. Almost all manufactures of dsps supply application notes on implementing adaptive filters on their products.

To gain an insight into the operation of adaptive filters it is useful to have an understanding of fixed gain digital filters.

Fixed-gain digital filters

Most fixed-gain digital filters are broadly grouped into two camps. The first is the finite impulse response, fir, filter which only depends on the current and previous input samples. The second is the infinite impulse response (iir) filter which operates on the current and previous input and output samples. These are expressed mathematically; the fir filter,

$$y(n) = \sum_{m=0}^{m=M-1} b_m x(n-m) \qquad \ldots 1$$

and the iir filter,

$$y(n) = \sum_{m=0}^{m=M-1} b_m x(n-m) - \sum_{k=1}^{k=K-1} a_k y(n-m) \qquad \ldots 2$$

In these expressions $\{x(n)\}$ represents the digitised set of samples from the input signal, $\{b_m\}$ the weightings placed (gain factors or coefficients) on the current and previous input values and $\{a_k\}$ is the respective weightings on the previous output values. The filter characteristics depend upon the coefficients $\{a_k\}$ and $\{b_m\}$.

The finite-impulse response filter is always well behaved and is unconditionally stable because its output is only dependent upon the curent and previous input samples $\{x(n)\}$. Whereas the fir filter with its feedback or recursive components $\{y(n)\}$ is not unconditionally stable. Careful control must therefore be exercised on the choice of $\{a_k\}$ to ensure that the filters remains stable. Digital filter design is extensively covered in the literature and a very readable text is by Ifeachor & Jevis[1]. During the operation of these type of filters the coefficients remain fixed. In an adaptive filter however this is not the case, the coefficients actually change between the arrival of each input sample x(n). Because of the stability problem and the ease of implementation, many adaptive digital filters are based on the non-recursive fir model.

Each digital filter will have its own transfer characteristics and this has the effect of reshaping the input signal. The filter inpulse response actually convolves with the input signal. In the case of a fir filter the reshaping process is static, in an adaptive filter the filter profile is dynamic which means that the reshaping is also dynamic and this changes within the sampling interval (time between the arrival of the digitised samples).

Adaptive digital filters

An adaptive digital filter derived from a fir filter design can be represented schematically as shown in **Fig. 1**. Each gain element is a multiplication process and the samples the enter

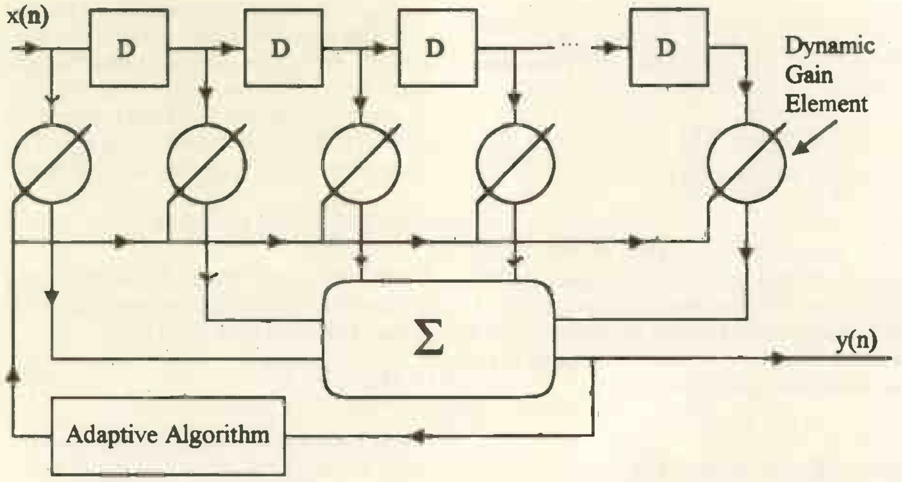

Fig. 1. Adaptive digital filter showing the dynamic gain elements.

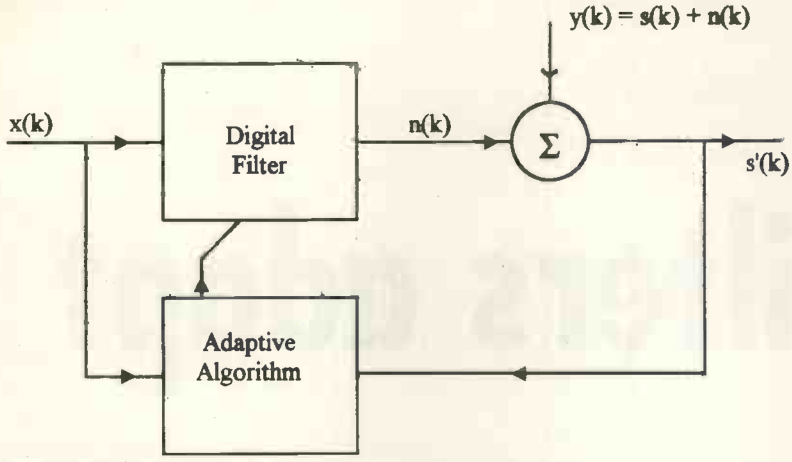

Fig. 2. Schematic of a noise cancellation system.

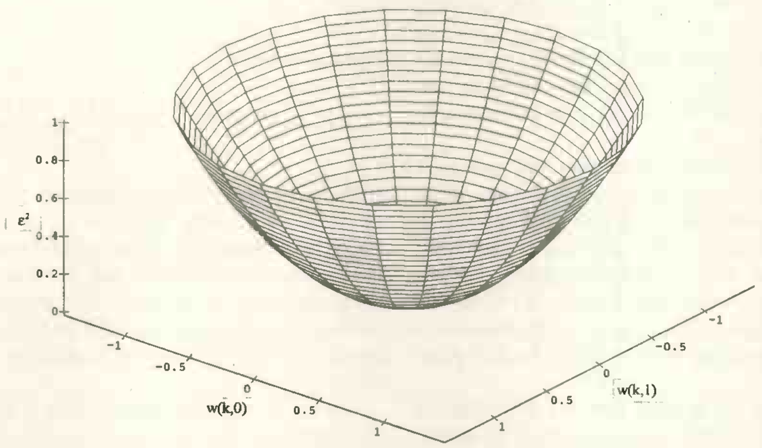

Fig. 3. Performance surface of an adaptive filter. Minimum error is at the base of the bowl.

the delay elements, marked D. The significance, or weighting, placed on each gain element will change dynamically according to the required shaping of the adaptive filter. What has not been mentioned is the method which selects the required gain on each filter element. This is achieved by implementing an adaptive algorithm. Consider the schematic of a system shown on **Fig. 2** for cancelling noise. When samples a derived from a system, each sample will consist of signal proper s(k) desired signal which has noise n(k) superimposed upon it to give y(k), so that,

$$y(k) = s(k) + n(k) \qquad ...3$$

What is required is a method for removing the n(k) component from y(k). As shown in Fig. 2, if {x(k)} is a sequence of samples which is in some way correlated to {n(k)} (the background noise for example), then its spectrum can be reshaped by the adaptive filter to produce the sequence {$\hat{n}$(k)} which matches the noise contamination {n(k)} on the desired signal {s(k)}. Therefore by subtracting one from the other it is possible to extricate {s(k)} from {y(k)} to obtain the noise free desired signal. The question remains how are the filter coefficients adjusted to perform the task? Answer – by means of an algorithm.

Adaptive algorithms

Many adaptive algorithms use a mathematical technique that is frequently encountered in curve fitting. Consider the noise (or an unwanted signal component) on the measured signal y(k) and a processed version of the noise $\hat{n}$(k) generated by the adaptive filter. The difference between them is called the error signal,

$$\varepsilon_k = n(k) - \hat{n}(k) \qquad ...4$$

From Fig. 2, you can see that the output of the system s′(k) is related to ε_k by,

$$s'(k) = y(k) - \hat{n}(k)$$
$$= s(k) + n(k) - \hat{n}(k)$$
$$= s(k) + \varepsilon_k$$

This shows how the output of the system {s′(k)} related to the desired signal {s(k)}. What is required is a method of minimising εk and this performed though the use of the least squares technique. By taking the square of ε^2_k the result will always be positive,

$$\varepsilon_k^2 = (n(k) - \hat{n}(k))^2 \qquad ...5$$

The total error squared ε^2 for a sequence of data containing K values is given by,

$$\varepsilon^2 = \sum_{k=0}^{k=K-1} [n(k) - \hat{n}(k)]^2 \qquad ...6$$

The signal $\hat{n}$(k) comes out of the digital filter and can therefore be expressed as,

$$\hat{n}(k) = \sum_{m=0}^{m=M-1} w(k,m)x(k-m) = \mathbf{W}^T\mathbf{X}(k) \qquad ...7$$

since the adaptive is acting like a finite impulse response filter. In Eq:7 {w(k,m)} are the coefficients of the adaptive filter which are represented as a vector **W** (see appendix). The T refers to the transpose operation which makes **W** into a row vector. **X**(k) is the vector whose elements are the M previous inputs to the filter. The objective of any adaptive algorithm is to minimise the value of ε. The filter, in effect, reshapes the signal {x(k)} to produce {$\hat{n}(k)$} to match {n(k)}. When Eq:7 is substituted into Eq:6 and expanded, the result is,

$$\varepsilon^2 = \sum_{k=0}^{k=K-1} n^2(k) - 2\sum_{k=0}^{k=K-1} n(k)\mathbf{W} + \sum_{k=0}^{k=K-1} \mathbf{W}^T\mathbf{X}(k)\mathbf{X}^T(k)\mathbf{W} \qquad ...8$$

The first term on the right-hand side is a measure of the noise power and is expressed as the variance σ^2 of the noise. Now introduce the autocorrelation vector **R** and the variable **P** which are defined as,

$$\mathbf{R} = \sum_{k=0}^{k=K-1} \mathbf{X}(k)\mathbf{X}^T(k) \qquad ...9a$$

and,

$$\mathbf{P} = \sum_{k=0}^{k=K-1} n(k)\mathbf{X}(k) \qquad ...9b$$

Eq:8 then becomes,

$$\varepsilon^2 = \sigma^2 - 2\mathbf{PW} + \mathbf{W}^T\mathbf{RW} \qquad ...10$$

Equation 10 is actually a quadratic in **W** which gives rise to a minimum. This means that when ε is plotted against **W** a multidimensional bowl is generated. **Figure 3** shows the case when **W** only has two elements in it. This multidimensional bowl is known as the performance surface and represents all possible values of **W**. Each set of coefficients in the vector **W** corresponds to a point on the performance surface. What is needed is a method of finding the value of the vector **W** that minimise ε^2. Remember that the vector **W** contains the coefficients of the filter and they have to be adjusted so that the it reshapes the signal {x(k)} to make it look like {n(k)}. This is achieved by differentiating Eq:10 to give,

$$\nabla\varepsilon^2 = \frac{d\varepsilon^2}{d\mathbf{W}} = -2\mathbf{P} + 2\mathbf{RW} \qquad ...11$$

At the minimum point on the surface – the bottom of the bowl – the gradient vector $\nabla\varepsilon^2$ is zero. Therefore from Eq: 11,

$$\mathbf{W}_{opt} = \mathbf{R}^{-1}\mathbf{P} \qquad ...12$$

This is known as the Wiener-Hopf equation[2] which actually contains a set of simultaneous equations, all of which must be solved during the sampling period of the process. Wopt is the vector which contains the optimum set of filter coefficients which ensures the best reshaping on {x(k)} to give {n(k)}. Obtaining Wopt is not an easy task since it involves a calculation to determine **R** and its inverse. However various algorithms have been developed for calculating an estimate of Wopt.

Least mean squares algorithm

The basis of the least mean squares (lms) algorithm is that the next value of the vector **W** (the set of coefficients for the adaptive filter which change every sample interval) is a modification of the current value. This can be expressed as,

$$\mathbf{W}_{k+1} = \mathbf{W}_k - \mu\nabla_k\varepsilon_k^2 \qquad ...13$$

The reasoning behind this lies in the belief that the shortest route from any point on the performance surface to the minimum point is alone the path of *steepest descent*. In order to determine the value of the gradient vector in Eq:13 it is necessary to consider Eq:6 but instead of the total squared error only the *k*th error value is evaluated, this gives,

$$\varepsilon_2^k = [n(k) - \mathbf{W}_k^T\mathbf{X}(k)]^2 \qquad ...14$$

Differentiating Eq:14 with respect to **W** gives,

$$\nabla_k\varepsilon_k^2 = 2[n(k) - \mathbf{W}_k^T\mathbf{X}(k)]\nabla_k\{n(k) - \mathbf{W}_k^T\mathbf{X}(k)\} \qquad ...15$$

Substituting for ε^2_k, Eq:15 becomes,

$$\nabla_k\varepsilon_k^2 = -2\varepsilon_k^2\nabla_k\{\mathbf{W}_k^T\mathbf{X}(k)\} \qquad ...16$$

which reduces to (see the appendix),

$$\nabla_k\varepsilon_k^2 = -2\varepsilon_k\mathbf{X}(k) \qquad ...17$$

Eq:13 (the lms Algorithm) becomes,

$$\mathbf{W}_{k+1} = \mathbf{W}_k + 2\mu\varepsilon_k\mathbf{X}(k) \qquad ...18$$

The lms algorithm is implemented by simply updating each filter coefficient accordingly,

$$w(k+1,m) = w(k,m) + 2\mu\varepsilon_k x(k-m) \qquad ...19$$

So far no mention has been made of the μ factor in the lms equations. In fact, m controls the rate of convergence and the stability of the algorithm and usually lies in the range, $0<\mu<1$. The rate of convergence relates to how long (how many samples) it takes for the algorithm to converge onto the desired signal. In practice the vector $\mathbf{W}_k$ never actually reaches the theoretical optimum value of the Weiner-Hopf equation (Eq:12) but tends to fluctuate about it.

Implementing the lms algorithm

As an example, consider the application of an adaptive filter for noise cancellation. In this case the error signal is the difference between the measured signal {y(k)}, which has the desired signal burried in it, and the output from the filter unit. In this case two signal sources are required. The first containing the desired signal contaminated by the noise and the second from a nearby location that only has noise which is correlated with the noise contaminating the desired signal. In a number or applications the contamination may not be noise, it may simply be an unwanted signal component. For example when trying to detect foetal heart beats, the unwanted signal component would be the mother's heart beat. Implementing the basic lms Algorithm for this applicatio can be achieved in four steps.

- Set the initial parameters to an arbitrary

Table 1. Least-mean-squares noise cancellation algorithm written in C. Performance of this routine is shown in Fig. 4.

```
/*  LMS Adaptive filter for noise cancellation   */
/* The data is read from and
written to RAM-DISC (drive d:)
sig_1.dat is the noise source and
sig_2.dat is the desired signal contaminated by noise */

#include<stdio.h>
#include<math.h>
#define mu  0.1
#define M   50  /* Number of taps in filter */
#define K   512 /* Number of data points    */
float x[M+1], w[M+1], y, desire, n ;
int k,m ;
char data_a[80] ;

main()
{
       FILE *signal_1, *signal_2, *out_sig, *noise ;
       signal_1 = fopen ("d:\\sig_1.dat", "r")  ;
       signal_2 = fopen ("d:\\sig_2.dat", "r")  ;
       out_sig  = fopen ("d:\\output.dat", "w") ;
       noise    = fopen ("d:\\noise.dat",  "w") ;

       /* Zero the filter coefficients */
       for( m= 0 ; m <= M ; m++)
              w[m] = 0;

       /* Perform filter operation (Eq:7) */
       for( k=0 ; k <= K-1 ; k++)
       {
         n = 0 ;

         fgets(data_a,40,signal_2);
         x[0] = atof(data_a) ;
         for( m=0 ; m <= M-1 ; m++)
              n = n + w[m] * x[m] ;

         /* Calculate the error estimate */
         fgets(data_a,40,signal_1);
         y = atof(data_a) ;
         desire = y - n ;
         fprintf( out_sig, "%f \n", desire);
         fprintf( noise,   "%f \n", n );
         /* Update coefficients (Eq:19) */
         for( m=M-1 ; m >= 0 ; m—)
         {
            w[m] = w[m] + 2*mu*desire*x[m] ;
             if (m != 0 )
              x[m] = x[m-1];
          }
       }
  fclose (signal_1);
  fclose (signal_2);
  fclose (out_sig);
  fclose (noise);
}
```

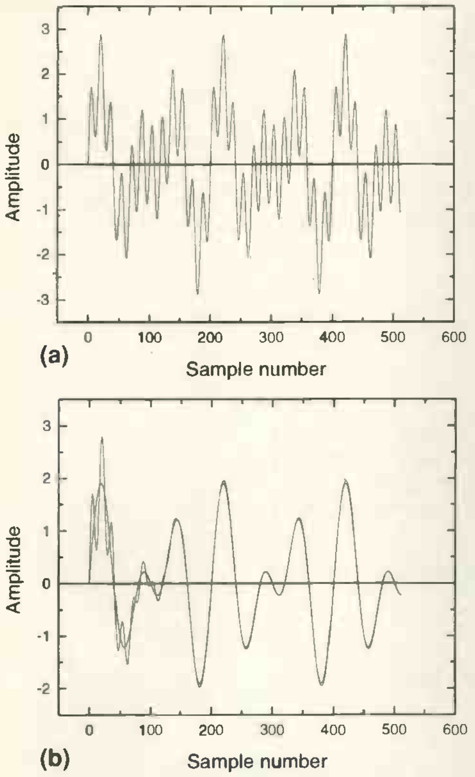

Fig. 4. Performance of the lms algorithm. Plot a) is an input signal contaminated with noise while b) shows convergence of the adaptive filter output on to the desired signal.

value, zero is convenient for this purpose and zero the memory locations for $\{x\}$. Choose a value for μ.

- Input samples x(k) and y(k) and start the filter to calculate the estimate of $\hat{n}(k)$ using Eq:7.
- Calculate an error value $\varepsilon_k=y(k)-\hat{n}(k)$ and output this (the desired signal).
- Update the filter coefficients using Eq:19 and continue.

Table 1 is a C program of the lms algorithm coded for noise cancellation. **Figure 4** shows an example of the performance of the algorithm. Figure **4a** shows the noise contaminated desired signal $\{y(k)\}$. Figure **4b** shows the actual desired signal and the output from the adaptive filter. It can be seen that after 120 samples the adaptive filter output matches that of the desired signal. The convergence could be improved by increasing the value of μ. Of course in practice one would not have access to the desired signal, after all this is what the adaptive filter is trying to exercise.

One of the drawbacks of the lms algorithm is its poor performance when used with signals that are regarded as very non-stationary (signals whose amplitude and frequency are rapidly changing). For non-stationary signals the performance surface shifts round in the **W** plane and the algorithm not only has to track the position of the surface but also the minimum point within it.

The lms algorithm has been coded in assembly language for a number of digital signal processors. For the Texas Instruments *TMS320C25* (fixed point) and *TMS320C30* (floating point), look at references 3 and 4 respectively. Although only the lms algorithm has been mentioned here, there are many other algorithms such as the recursive least squares, rls, algorithm – and that is another story. ■

References

1. EC Ifeachor & BW Jervis, *Digital signal processing - a practical approach*, Addison-Wesley (1993).
2. S Haykin, *Adaptive filter theory* (Second Edition), Prentice-Hall (1991) Chapt 5.
3. R. Chassaing & DW Horning, *Digital signal processing with the TMS320C25*, Wiley (1990) Ch. 8.
4. R Chassaing, *Digital signal processing with C and the TMS320C30*, Wiley (1992), Chapt 7.

Matrix representation

As an example consider the case when there are only two elements in the sum of Eq:7,

$$\hat{n}(k) = \sum_{m=0}^{m=1} w(k,m)x(k-m) = w(k,0)x(k) + w(k,1)x(k-1)$$

This can be written in a vectorial form as,

$$\hat{n}(k) = \begin{pmatrix} x(k) & x(k-1) \end{pmatrix}\begin{pmatrix} w(k,0) \\ w(k,1) \end{pmatrix} = \begin{pmatrix} w(k,0) & w(k,1) \end{pmatrix}\begin{pmatrix} x(k) \\ x(k-1) \end{pmatrix}$$

Writing the vectors **X**(k) as,

$$\mathbf{X}(k) = \begin{pmatrix} x(k) \\ x(k-1) \end{pmatrix} = \begin{pmatrix} x(k) & x(k-1) \end{pmatrix}^T$$

and W as,

$$\mathbf{W} = \begin{pmatrix} w(k,0) \\ w(k,1) \end{pmatrix} = \begin{pmatrix} w(k,0) & w(k,1) \end{pmatrix}^T$$

therefore,

$$\hat{n}(k) = \mathbf{X}^T(k)\mathbf{W} = \mathbf{W}^T\mathbf{X}(k)$$

In Eq:16 the gradient vector can be expressed as,

$$\nabla_k = \begin{pmatrix} \dfrac{\partial}{\partial w(k,0)} \\ \dfrac{\partial}{\partial w(k,1)} \\ \dfrac{\partial}{\partial w(k,M-1)} \end{pmatrix}$$

and since,

$$\mathbf{W}_k^T = \begin{pmatrix} w(k,0) & w(k,1) & . & w(k,M-1) \end{pmatrix}^T$$

then,

$$\nabla_k \mathbf{W}_k^T = 1$$

leaving,

$$\nabla_k\{\mathbf{W}_k^T\mathbf{X}(k)\} = \mathbf{X}(k)$$

since the derivative do not affect **X**(k).

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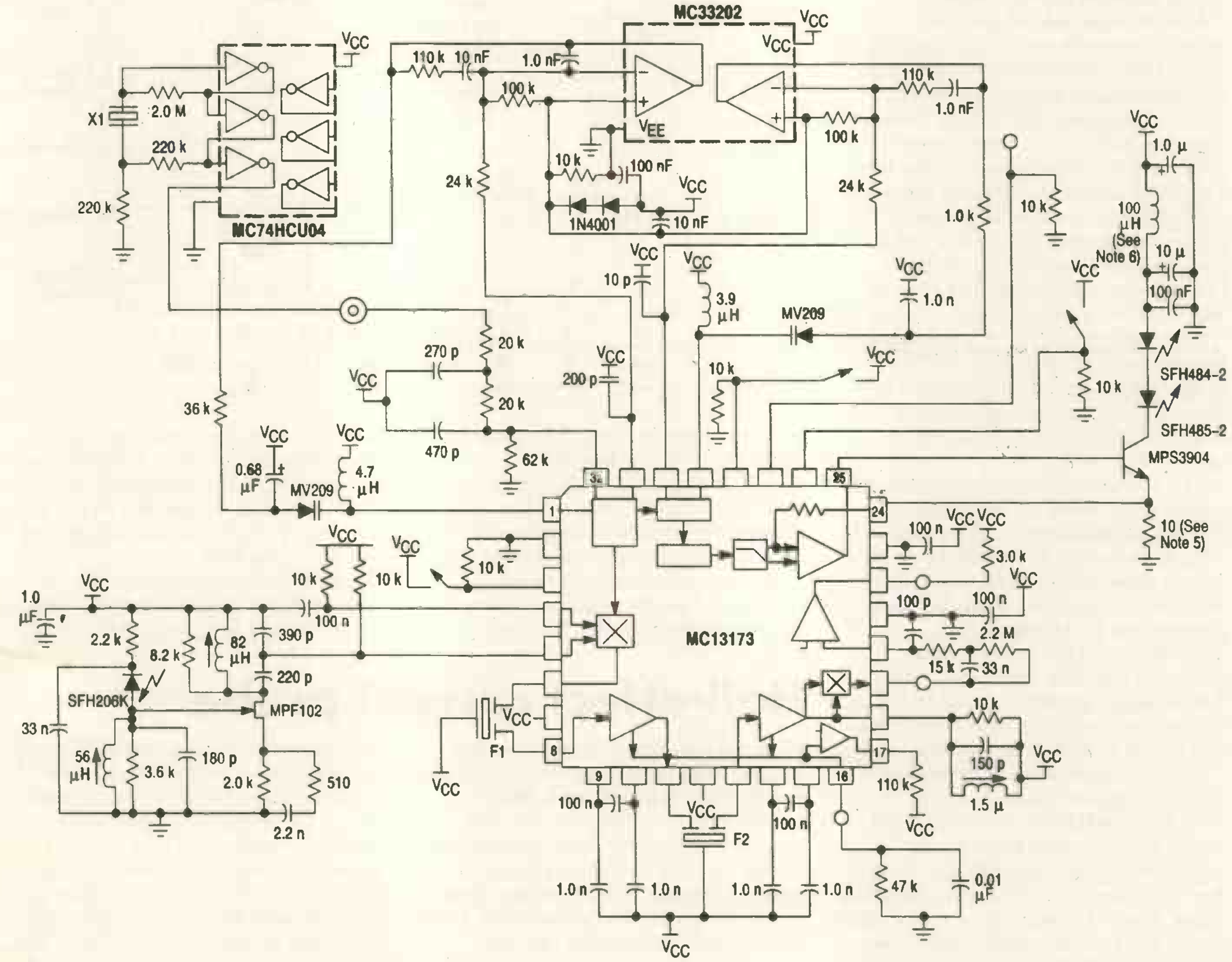
Half-duplex infrared computer link

From Motorola, the *MC13173* is a low power infrared integrated system (IRIS). It is a unique blend of a split IF wideband fm receiver and a specialized infrared led transmitter. The device is designed to provide communications between portable computers via a half duplex infrared link at data rates up to 200kb/s. The receiver comprises a mixer, IF amplifier and limiter with two cascaded filters, data slicer and received signal strength and carrier detect functions. The transmitter section includes a frequency synthesizer, fsk modulator, harmonic low pass filter and an infrared led driver. The transmitter can operate in two modes – on-off pulsing for remote control or fsk modulation at 1.4MHz for point to point digital communications. All are controlled via a six line interface bus. Over 70dB of rssi range is achievable with split IF for improved filtering.

For power conservation, individual circuit blocks can be powered down when not in use. Typical applications include cordless



1) F1 and F2 – 10.7MHz ceramic filter, Toko *107MA-AE-10* (360kHz), Toko *107M0 AE-10* (330kHz) or equivalent.
2) Tunable shielded inductors:
56µH Toko *A119ANS-T1042Z* 1.0µH Toko *292KNS-T1372Z*
82µH Toko *A119ANS-T1044Z* 1.5µH Toko *292KNS-T1373Z*
3) Crystal X1 – 32.768kHz C-type tuning-fork crystal. Digikey part No *SE 3201* or equivalent.
4) LEDs and detectors *SFH484-2*, *SFH 485-2* and *SFH206K* are made by Siemens.
5) Optimum bias resistors depends on the leds used.
6) May be fixed or tunable.

Using the transceiver IC for data communications. The* MC13173 *provides a complete led driver stage and power saving modes. A pcb foil for the application circuit is given in data sheet MC13173D from Motorola.

communication from keyboard to computer and data link between portable computers.

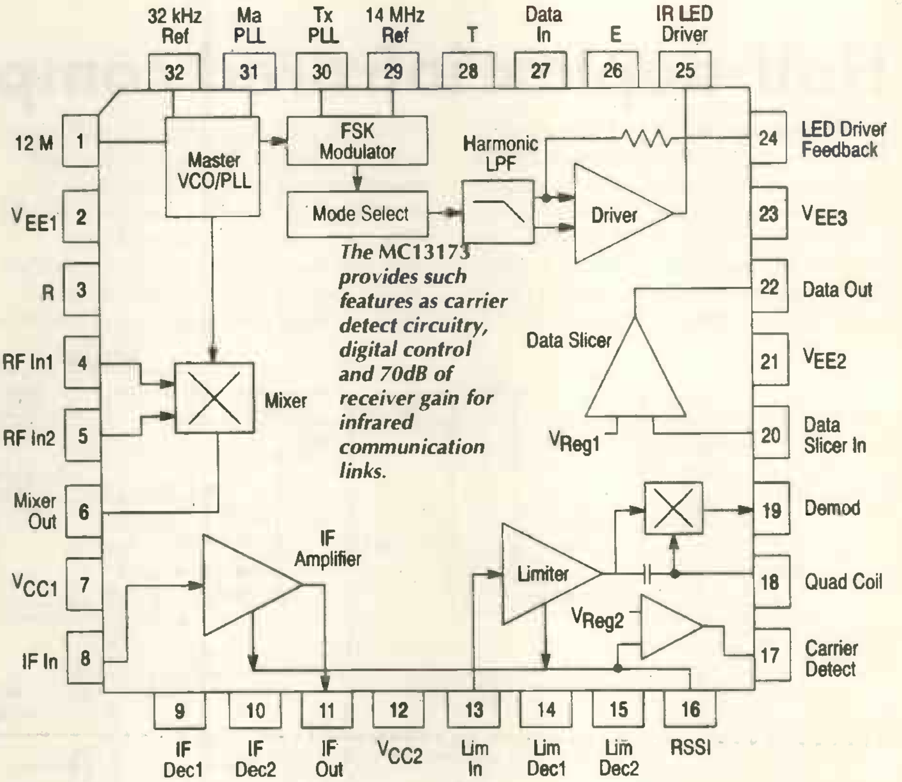
The six line 3.3V digital interface bus includes three control pins, data in and data out pins and a carrier detect pin. The master vco/pll provides the reference frequency for the fsk modulator and the LO frequency for the receiver down converter. With a 32.768kHz input frequency to the master vco on pin 1, the LO frequency for the receiver will be at 12.075MHz. The reference frequency for the fsk modulator will be at approximately 1.1MHz.

A double balanced four quadrant multiplier/mixer is provided so that it can be driven either differentially or single-ended by connecting the unused input to the positive supply rail. The buffered output is internally loaded for an output impedance of 330Ω for use with a standard ceramic filter. The first IF amplifier section has internal dc feedback and external input decoupling for improved symmetry and stability with a total gain of approximately 40dB. The fixed internal input impedance is 330Ω for use with a 10.7MHz ceramic filter.

Output of the IF amplifier is buffered and the impedance is 330Ω. When a signal is being received, a current proportional to the log of the received signal amplitude is output on rssi, derived by summing the currents from the IF and limiting amplifier stages. An external resistor sets the output voltage range. When the rssi level exceeds the carrier detect threshold at approximately 1.2V dc, the carrier detect output goes high. A large resistor may be added externally between the comparator output and the positive input for hysteresis.

The demodulator is a conventional quadrature type with an external *LC* tank driven through an internal 5pF capacitor. A buffered output gives an impedance of less than 1kΩ at an average dc level of around 1.1V. The data slicer is designed to square up the data signal. It is self centering at about 1.1V, and clips at about 0.75V and 1.45V. There is a short time constant for large peak-to-peak voltage swings or when there is a change in dc level at the detector output. The time constant is longer for small signals or for continuous bits of the same polarity which drift close to the threshold voltage.

The *MC13173* uses a dual modulus pll to fsk modulate the baseband digital input signal, producing the necessary logic high and low frequencies for transmission. The transmit frequency for a logic high is 1.427MHz, and the frequency for a low is 1.317MHz with a 32.768kHz reference frequency. In communications mode, the fsk



The MC13173 provides such features as carrier detect circuitry, digital control and 70dB of receiver gain for infrared communication links.

modulator uses the reference frequency from the master vco to produce the two frequencies required for a logic high and a logic low. In the a/v mode, the fsk modulator is not used and is powered down.

A low pass filter following the fsk modulator removes undesired harmonic frequencies from the square wave output of the divider circuits in plls. The resulting sinusoidal waveforms feed a unity-gain difference amplifier which drives the base of an external transistor, modulating the ir led. In a/v mode, data is input directly into the inverting input of the op amp, and the low pass filter is not used.

***Motorola Ltd**, 88 Tanners Drive, Blakelands, Milton Keynes MK14 5BP. Tel 01908 614 614, fax 01908 618 650.*

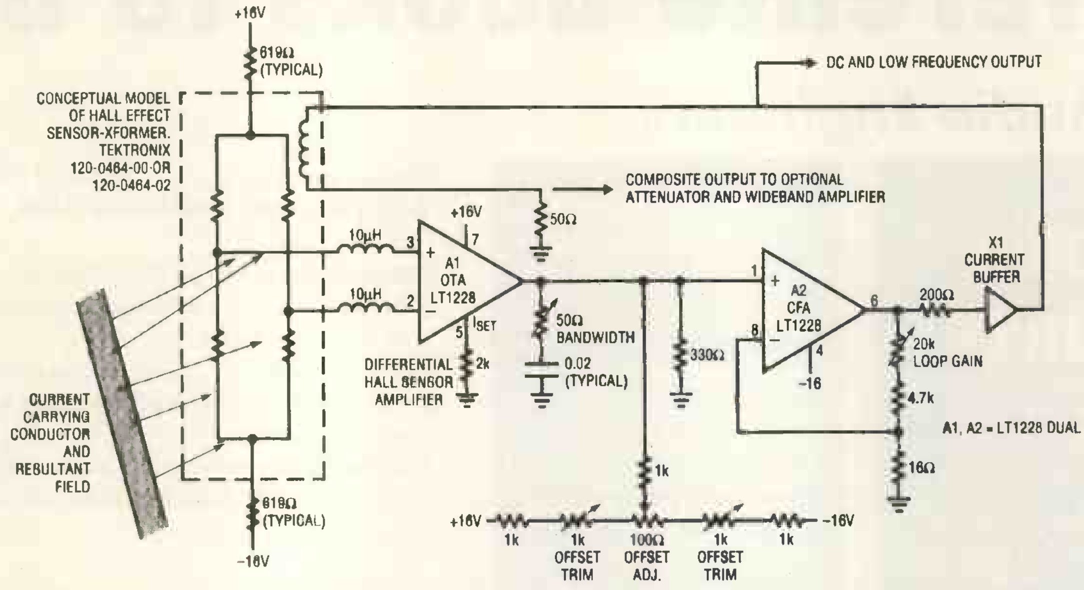
Hall-effect current probe

Current transformers are common and convenient, permitting wideband current measurement, independent of common-mode voltage considerations. The most convenient current transformers are the clip-on type, commercially sold as current probes. A problem with all simple current transformers is that they cannot sense dc and low frequency information. This was addressed in the mid-1960's with the advent of the Hall effect stabilized current probe.

The approach discussed in *Linear Technology's* application note 61 uses a Hall-effect device within the transformer core to sense dc and low frequency signals. This information is combined with the current-transformer output to form a composite dc-to-high frequency output.

Careful roll-off and gain matching of the two channels preserves amplitude accuracy at all frequencies. Additionally, the low frequency channel is operated as a 'force-balance', meaning that the low frequency amplifier's output is fed back to magnetically bias the transformer flux to zero. Thus, the Hall effect device does not have to respond linearly over wide ranges of current and the transformer core never sees dc bias; both advantageous conditions. The amount of dc and low frequency information is obtained at the amplifier's output, which corresponds to the bias needed to offset the measured current.

The diagram on page 231 shows a practical circuit. The Hall-effect transducer lies within the core of the clip-on current transformer specified. A very simplistic way to model the Hall generator is as a bridge,



A Hall effect device combines with a current measuring transformer to overcome the problem of measuring current at low frequencies down to dc.

excited by the two 619Ω resistors. The Hall generator's outputs at the midpoints of the bridge feed differential-input transconductance amplifier A1, which takes gain, with roll-off set by the 50Ω/20nF *RC* at its output. Further gain is provided by A2, packaged with Al. A current buffer provides power gain to drive the current transformer's secondary. This connection closes a flux nulling loop in the transducer core.

The offset adjustments should be set for 0V output with no current flowing in the clip-on transducer. Similarly, the loop gain and bandwidth trims should be set so that the combined high and low frequency output across the grounded 50Ω resistor has a clean step response and correct amplitude from dc to high frequencies.

Micro Call Ltd, 17 Thame Park Road, Oxon, OX9 3XD. Tel 01844 261939, fax 01844 261678.

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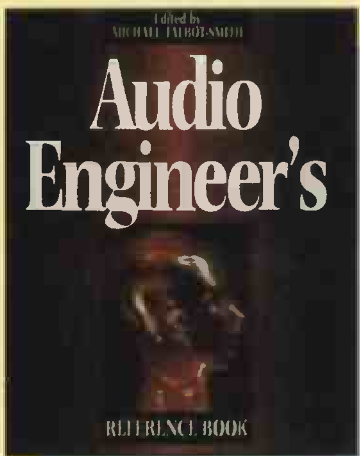

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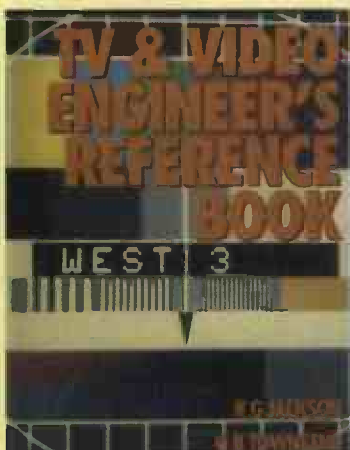

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Visualising electron disturbances

Transmission lines can be easier to understand when you consider them as electron disturbances, explains Geoffrey Billington.

The transmission of electrical pulses may be treated in a similar way to the transmission of sound through a tube of gas, or to the disturbances produced in a long spring by pushing one end backwards and forwards along the spring's length.

In fact the case of electron pulses is more complex and cannot be fully explained without considering the fields in the space between the wires, but many aspects of transmission line behaviour can be derived without any direct reference to fields. The following crude model may help you to visualise electron disturbances travelling down a wire.

Imagine a corridor, closed at both ends and tightly packed with people. Suddenly, a group of extra people is pushed in at one end. As a result of this, a surge or compression pulse passes along the corridor. No person moves very far, but the pulse travels to the other end of the corridor and may even be reflected from there when the people are squashed against the end wall and push back again.

If instead of people we have electrons in a wire, similar sorts of effects can take place. In this case we should say that a current flowed during the passage of the pulse through any part of the wire.

An alternative sequence of events would occur if a number of people were suddenly removed from one end of the corridor. People would fall into the empty space, and again a pulse would move down the corridor, but this time it would be a pulse of reduced density.

In the electrical case this would be a positively charged region which is propagated down the wire – i.e, a region where there are insufficient electrons to neutralise the immobile positive charges. Even though the positive charges do not move, the 'hole' does so.

Electrical pulses on a transmission line

Now picture a generator connected across one end of a pair of long wires. The generator produces a pulse of very short duration by 'pumping' a number of electrons out of one wire and into the other. The combined pulse consists of a positively charged region in one wire and a negatively charged region in the other; these move parallel to each other along the wires at a speed nearly equal to that of light.

It should be appreciated that although both pulses move outward from the generator, the electric currents in the wires must be taken as being in opposite directions, because the arrow indicating the current points in the direction of motion of the *positively* charged region, but in the opposite direction to the motion of the negatively charged region, **Fig. 1**.

Another important point arises here. Providing the pulse lasts a short time, it will be over and done with before the leading edge of the pulse has reached the end of the line – or to be more precise, before the leading edge of any possible reflected pulse arrives back at the generator.

During this time, the current provided by the generator is entirely independent of whatever load may be connected across the far end of the line. The generator 'sees' a resistance, whose value depends upon the wire diameter and spacing and is termed the

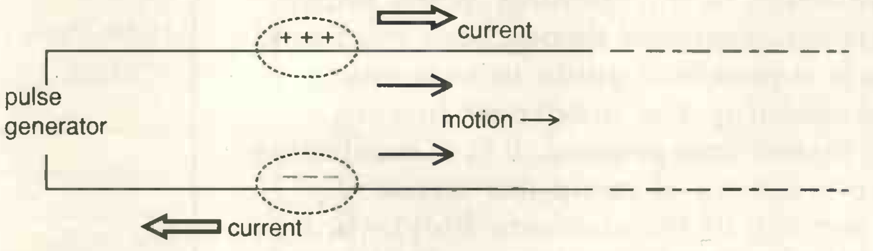

Fig. 1. Currents and motion of charged regions. A pulse generator is connected at one end of a pair of long wires and a short pulse is produced that consists of a positively charged region in one wire and a negatively charged region in the other.

'characteristic impedance' or 'characteristic resistance' (Z_0) of the line. In practice it will probably be in the region of a few tens or hundreds of ohms.

If a generator is connected to a resistor, heat energy is produced. In the case of transmission lines the energy is invested in the electric and magnetic fields associated with the pulse, but it may be calculated in the usual manner. The power supplied to the line by the generator before the arrival of any reflected wave is V^2/Z_0 watts, where V is the voltage applied to the line.

The impedance 'seen' by the generator will change on the arrival of any reflected pulse at the generator end of the line.

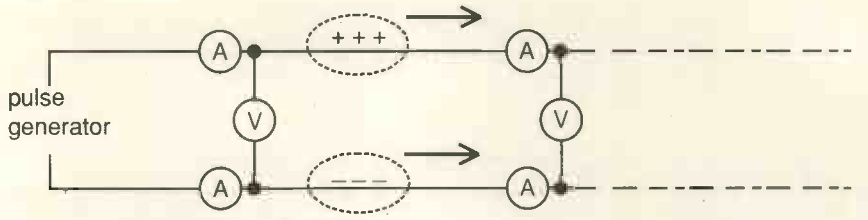

Fig. 2. Monitoring currents and voltages. Pairs of ammeters connect to the wires. As pulses pass, readings are opposite but equal.

Currents and voltages in a travelling wave

In order to give some idea about what is happening, imagine that the whole process could be slowed down. Pairs of ammeters are connected in the wires as shown in **Fig. 2**. As the pulses pass, each pair of ammeters will show a reading which will increase from zero and then return to zero again. The readings of each pair will be equal and opposite. The more distant ammeters will respond later than the nearer ones, but their readings will go through exactly the same sequence, and in the ideal case of loss-free lines there will be no reduction in the current values.

In the same way, voltmeters can be connected between opposite points on the lines. In Fig. 2 the voltmeter and ammeters monitor the same short section of a long line. If the readings are monitored during the passage of a pulse it will be found that at all times during the passage of the pulse:

Volts=Amps=Z_0.

where Z_0 is the characteristic impedance of the lines mentioned earlier; i.e, the resistance 'seen' by the generator during the production of the pulse.

Eventually the pulse arrives at the end of the line. What happens there depends upon how the line is terminated, but the most likely outcome is that a reflected pulse will move back down the line. The reflected pulse is usually weaker than the original pulse: it is never stronger.

If the end of the line is left open circuit, i.e, the line simply ends, no energy is absorbed from the pulse and a reflected pulse of the same shape and energy is obtained, albeit turned round so that the leading edge is still at the front. A perfect short circuit across the ends also acts as a perfect reflector. (see later).

If a resistor is connected across the ends, the general result is that some energy is absorbed in the resistor, the remaining energy being carried off by a weakened reflected pulse. If the terminating resistor is equal to the characteristic impedance of the lines, all the energy is absorbed and no reflection occurs.

It is perhaps surprising that it is a fairly simple matter to explain all these effects and to calculate the degree of reflection which will occur with any given terminating resistor. As we shall be dealing with pulses travelling in both directions there is one important fact which must be appreciated. If the direction of motion of a charged region is reversed, the current reverses sign but the voltage remains unchanged.

To illustrate this, suppose that a pulse is moving from left to right along a transmission line. The pulse consists of a positively charged region moving out along one wire, and a negatively charged region along the other. The passage of the pulse is imagined to be slowed down and monitored with ammeters and a voltmeter as before, **Fig. 3**. The pulse passes the meters and the deflections are noted.

Now suppose that the pulse is reflected at an open circuit and the reflected pulse, consisting of the same charged regions, approaches the meters moving in the opposite direction.

The voltmeter reading would be unaffected as the charged regions are exactly as before, but the ammeter readings would reverse due to the reversal of the motion of the charged regions.

This fact is of vital importance in what follows and throughout the article.

Two opposing pulses pass through one another

Another thing worth considering is what happens when two pulses moving in opposite directions pass through one another. Now the pulses on the lines behave like typical waves and obey the same general rules.

If you throw two stones into a pond and watch the ripples, you observe that they pass through each other and continue as if nothing had happened. To find out what happens while they are passing through one another we can apply the 'principle of superposition'. This is a useful principle which can be applied to many different sorts of wave motion.

In the case of electrical disturbances the rule simply states:

"The resultant voltage or current (at a given point and given instant) due to two waves is obtained by adding together the voltages or currents which each would produce on its own."

If the voltages or currents have opposite signs, the addition becomes a subtraction, e.g. +7 and –4 give +3. The voltages and currents due to the two waves are termed 'component' voltages and currents. The values obtained by combining the component values are termed 'resultant' values.

It is the resultant values which give the real currents and voltages; i.e, the reading which an ammeter or voltmeter would record.

Suppose the resultant value of the current is found by combining a component current of 7A flowing in one direction and a component current of 4A in the opposite direction. The actual electron flow in the wire will consist of a current of 3A. Don't imagine that opposite currents of 7A and 4A are somehow flowing in two opposing streams within the same piece of wire.

A general rule

Much of what follows is concerned with what happens when two transmission line waves travelling in opposite directions are superimposed. There is one rule which will be used frequently. It is an extension of the previous statement about reversing the motion of a charged region.

"At any point where two waves travelling in opposite directions are superimposed, either the component voltages add and the component currents subtract, or vice versa."

This is because the termination of a line always imposes some condition which the voltage and/or current must obey at this point, and it is this which determines how reflection occurs.

The open circuit termination is perhaps the simplest example of how reflection is determined. If a line is terminated with an open circuit, i.e, the wires simply finish and are not connected to anything, there is one condition which must be fulfiled. The current at the end of the line must be zero. There is nothing for it to flow into or out of.

This can be achieved if a suitable reflected wave is set up. All that is required is for a reflected pulse to flow away from the termination and to give a current which is at all times exactly equal and opposite to the current due to the forward wave at this point. This would result in continuous current cancellation at the termination.

At all instants, such a reflected wave would produce a voltage at the termination which was identical to the voltage produced by the oncoming wave – the waves travel in opposite directions, therefore the equal and opposite currents are associated with equal voltages of the same sign.

The voltages due to these two waves always add at the open circuit termination, where they give double the voltage which the oncoming pulse would produce on its own, and a reflected pulse of the same shape and energy as the oncoming pulse passes back down the line.

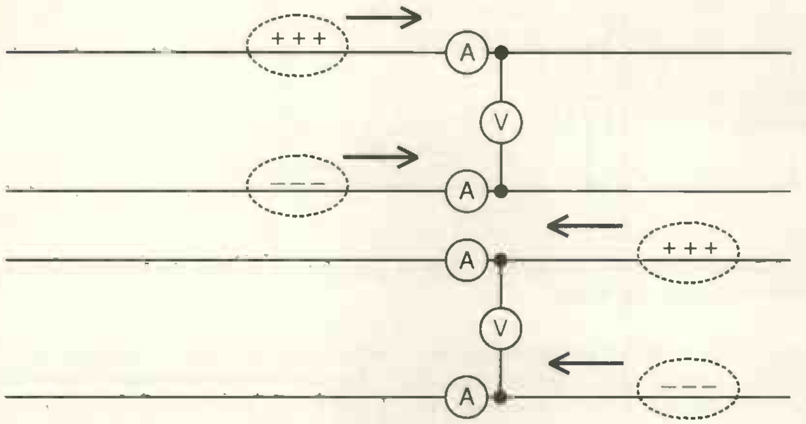

Fig. 3. Travelling waves moving in opposite directions. The sign of the current reverses but the voltage is unchanged.

Resistive terminations

If the line is terminated with a resistive load, R, this dictates that at the termination volts/amps=R.

In the oncoming pulse the ratio volts/amps is equal everywhere to Z_0, so unless it so happens that $R=Z_0$, the oncoming wave cannot fulfil the conditions dictated by the load, and reflection occurs. The oncoming and reflected waves together produce resultant values of voltage and current which fulfil the requirements of the load at all times.

Another point about a resistive termination is that it must absorb some of the energy of the oncoming pulse, so the reflected pulse must be weakened. The result is that both the current and voltage of the reflected wave are scaled down by a factor k, termed the 'reflection coefficient'

At the termination, at any given instant, the component voltage and current due to the forward wave are V and I, and the components due to the reflected wave are kV and kI.

As mentioned before, when two waves travelling in opposite directions are superimposed, either the component voltages add and the component currents subtract or vice versa. There are therefore two possible ways in which the reflection may occur:

Either resultant voltage is V+kV and resultant current is I–kI. Or resultant voltage is V–kV and resultant current is I+kI This gives two equations for R. Either:

$$R=(V+kV)/(I-kI) \text{ or } R=(V-kV)/(I+kI)$$
$$R=V(1+k)/I(1-k) \text{ or } R=V(1-k)/I(1+k)$$

But $V/I=Z_0$, so,

$$R=Z_0.(1+k)/(1-k) \text{ or } R=Z_0(1-k)/(1+k).$$

The equation $R=Z_0.(1+k)/(1-k)$ is applicable if R is greater than Z_0, and $R=Z_0.(1-k)/(1+k)$ is applicable when R is less than Z_0. Reflection occurs in different ways depending upon whether R is greater or less than Z_0.

These two equations can be rearranged to give two equations for k. If 'R' is greater than Z_0, $k=(R-Z_0)/(R+Z_0)$, and if 'R' is less than Z_0, $k=(Z_0-R)/(R+Z_0)$.

Reflection from a short circuit

For a short circuit termination R=0. Substituting this in the second equation gives:

$$k=(Z_0-0)/(Z_0+0)=Z_0/Z_0=1.$$

A short circuit also acts as a perfect reflector. This is not surprising as no power can be dissipated in a zero resistance. You may like to verify that putting $R=Z_0$ in either of the formulae gives k=0.

Both open circuit and short circuit terminations absorb no energy and act as perfect reflectors.This means that they both have a reflection coefficient, k, of unity.

If a line is terminated with a resistance equal to Z_0, there is no reflection and all the power is absorbed in the load. The line is 'matched'.

If a line is terminated with a resistance other than Z_0, reflection will occur. There are always two values of resistance which will give the same reflection coefficient, k, one value being greater than Z_0 and the other smaller. There is however a difference in the type of reflection produced by these two resistances.

Reflection with voltage inversion

Reflection by a short circuit, or by a resistance which is less than Z_0, may at first look surprising. The sign of the voltage between the lines is reversed after reflection. This means that a positively charged region moves out along one wire and returns as a negatively charged region after reflection.

The simplest way of looking at this is to say the two oppositely charged regions originally moving outward along the two wires arrive together at the termination where they may suffer some attenuation but pass through each other like water ripples and return along the opposite wires. It might be expected that the oppositely charged regions would simply neutralise one another and the pulses disappear, but inductive effects prevent this.

From this point of view you might wonder whether the process should be referred to as reflection.In fact there is a good reason for this, and equivalent effects occur with all sorts of waves.

An alternative electrical example of this sort of reflection occurs if a charged region is propagated along a single wire which terminates in a large metal block or is earthed. The pulse will overshoot at the termination, causing an oppositely charged region to return along the wire.

Multiple reflections

Having looked at reflection at various resistive terminations, we can now look at what happens to the reflected pulse when it arrives back at the generator.

Suppose that the generator is quiescent when the pulse returns. What happens then obviously depends upon the impedance offered by the generator to the returned pulse. Unless the pulse is completely absorbed, reflection must again occur. This means that our original pulse will die away by 'echoing' backwards and forwards along the line. In actual practice the decay is likely to be rapid and the pulse will become undetectable after a few reflections.

Periodic waves and the steady state

Now suppose that the generator is producing an alternating voltage of constant frequency and is still running when the reflected wave returns back.

Before the return of the reflected wave the generator 'sees' a resistance Z_0, but this impedance must change on the arrival of the reflected wave. This will, as a rule, cause a change in the amplitude and/or phase of the forward wave. We therefore appear to be faced with the daunting prospect of this process repeating over and over again, with the forward wave changing each time the reflected wave returns.

However, we can appeal to experience here. We know that conditions on the line settle into a steady state, when an unchanging wave of amplitude V volts flows away from the generator, and a reflected wave of amplitude kV flows back down the line. We also know that the time taken for this to happen is so short that it is not normally detectable.

What must happen is that a series of modifications occurs; these are in the form of decreasing 'corrections' to the forward wave, leading rapidly to the steady state, which is effectively attained after a few cycles.

No one should be too surprised by the idea of a 'settling down' period. Most ac circuits require a short time after switching on before each current cycle is identical to the previous one.

In the steady state, the result of superimposing the forward and reflected waves is to produce what is termed a 'standing wave'. This consists of a stationary pattern of alternating currents and voltages which have maximum and minimum rms values at regularly spaced fixed points on the line. ■

In a second article, Geoffrey Billington looks at standing waves in more detail.

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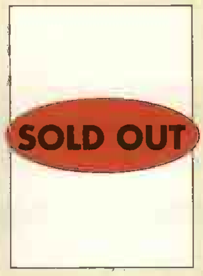

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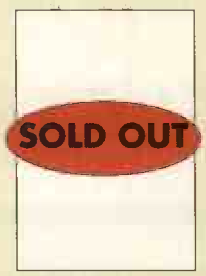

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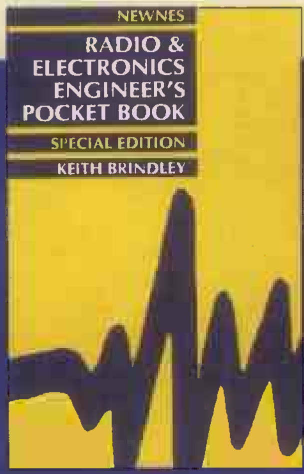

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Signed

SEND YOUR CIRCUIT IDEAS TO THE EDITOR, ELECTRONICS WORLD, QUADRANT HOUSE, THE QUADRANT, SUTTON, SURREY SM2 5AS

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

£100 WINNER

Fast power amplifier for audio

Since phase shifts in the output stage set a limit to the amount of feedback that can be applied, this power amplifier uses the fastest output stage possible with readily available components. This output stage was modified for common-source output from a proposal by Finnegan[2] and made faster by adding a differential drive to the current mirrors $Tr_{3,4}$ and $Tr_{7,8}$ to speed up charge and discharge of each mosfet's gate capacitance, allowing 50ns rise and fall times and a gain-bandwidth product of 15MHz.

Sensing the mosfet source current with R_{17} and R_{18} makes each mosfet much more linear and insensitive to temperature variations; drain current is proportional to $I_1 - I_2$ when $I_1 > I_2$ and is very linear down to a drain current of 50mA.

Without feedback, the output stage power bandwidth is 500kHz and gain is 40 with an 8Ω load. Reducing the load impedance reduces the open-loop gain, but this effect can be offset by reducing $R_{17,18}$ (or $R_{9\text{-}12}$). Some local feedback is applied to the output stage with $R_{15,16}$ and C_4 to give a gain of 3, making the output stage bandwidth 10MHz.

Overall feedback of 40dB at 10kHz is applied via the op-amp to reduce distortion. Output-stage distortion can be trimmed to around 1% by adjusting RV_1 and RV_2 in equal increments while the overall feedback is disabled, by shorting pin 2 to 6 of the op-amp, using a 1V/100Hz sinusoidal input. With overall feedback, distortion in the audio frequency range is estimated to be about 0.005%, with an overall circuit bandwidth of 1MHz and a gain of 15. Input filter R_1C_1 keeps the op-amp free from slew-rate distortion by limiting the input bandwidth to 150kHz.

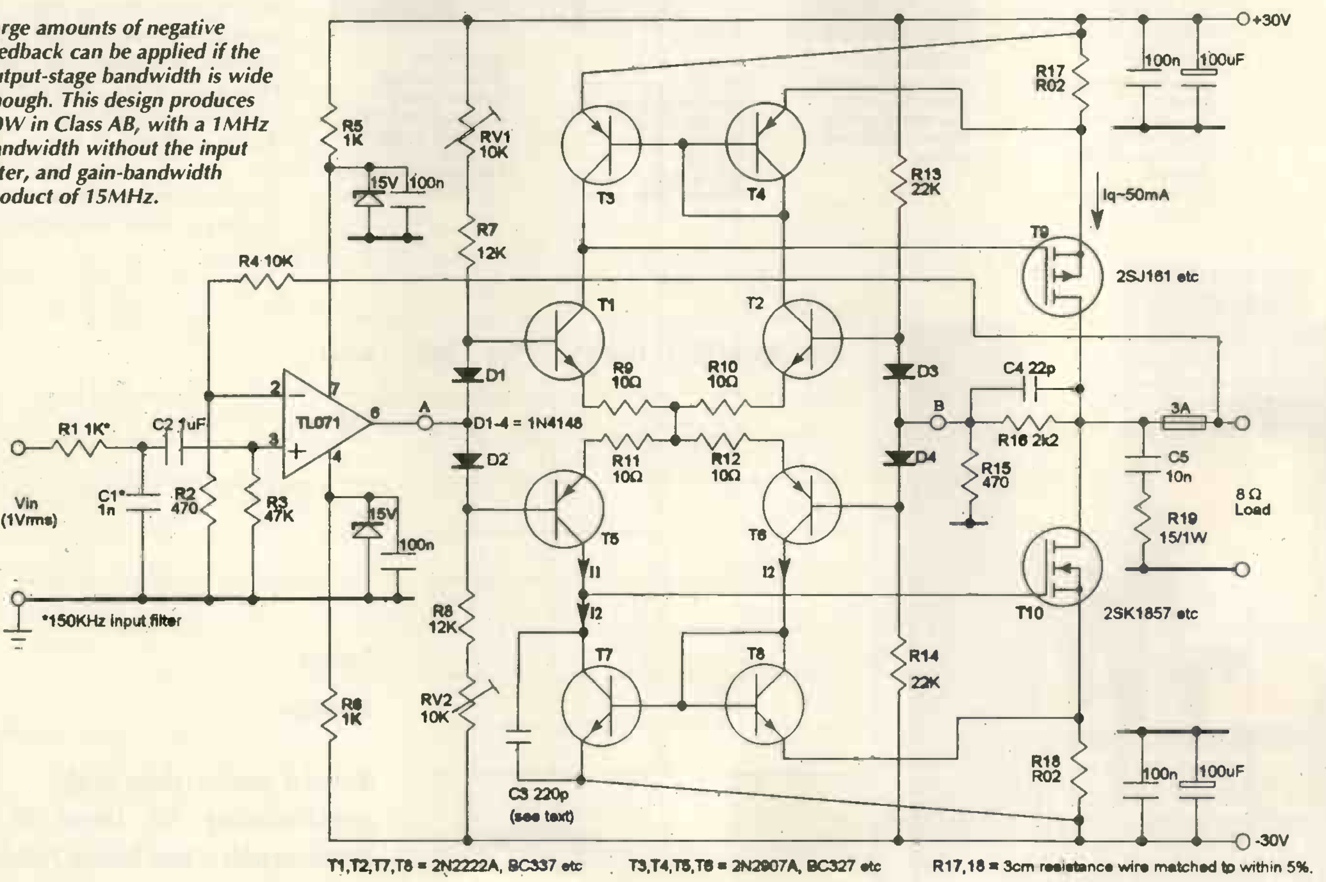
Tie mirror pairs $Tr_{3,4}$ and $Tr_{7,8}$ together for thermal stability; their emitters should be close to the sense resistors to prevent parasitic oscillation. Capacitor C_3 may need to be altered for different types of mosfets to equalise the difference between the p and n mosfet gate capacitance.

Ian Hegglun
Manawatu Polytechnic
New Zealand

References

1. Self, D., *Distortion in power amplifiers*. *EW & WW*, February, 1994, p.139.
2. Finnegan, T., *Precise power output stage*. *EW & WW*, May, 1993, p412.

Large amounts of negative feedback can be applied if the output-stage bandwidth is wide enough. This design produces 50W in Class AB, with a 1MHz bandwidth without the input filter, and gain-bandwidth product of 15MHz.



YOU COULD BE USING A 1GHz SPECTRUM ANALYSER ADAPTOR!

Got a good idea? Then this Thurlby-Thandar Instruments TSA1000 spectrum analyser adaptor could be yours.

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Recognising the importance of good design, **TTI will be giving away one of these excellent instruments every six months** to the best circuit idea published in the preceding period until further notice. **This incentive will be in addition to our £100 monthly star author's fee, together with £25 for all other ideas published.**

Our judging criteria are ingenuity and originality in the use of modern components – with simplicity particularly valued.

Peak detector benefits from high-speed op-amps

Peak detectors of the basic form shown are well known[1], but it is worth looking at the circuit using high-speed, high-gain, low-noise op-amps such as the *TLE2027/37*.

There are pitfalls. In the circuit shown, A_1 should be an uncompensated op-amp to obtain the highest speed but, since D_1 presents a highly variable impedance, R_3C_3 reduce overshoot. If this is overdone, the circuit is slowed and narrow input pulses ignored; the values given allow a 2% or 3% overshoot for a 600mV step. Capacitor C_1 prevents instability and, with R_1, limits bandwidth to 5MHz.

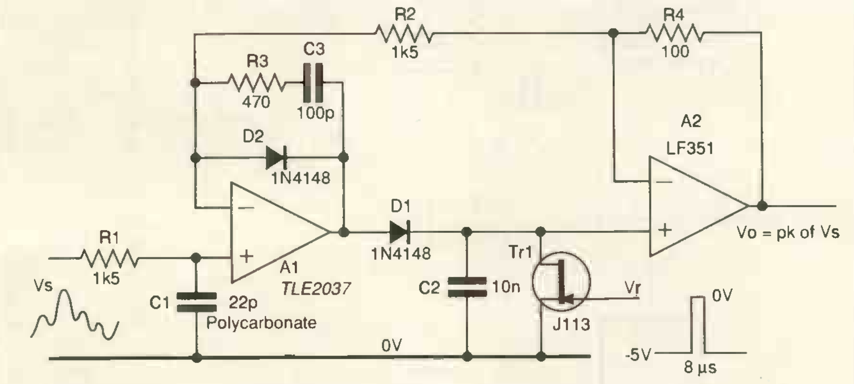
Saturation in the output stage of A_1 during parts of the signal below the voltage stored on C_2 from previous input would occur were it not for the clamp diode D_2; there is a built-in recovery circuit in the op-amp, which also helps.

Op-amp A_2 is a unity-gain buffer with low bias current to minimise droop on C_2; its slew rate is important, nevertheless; jfet Tr_1 resets the circuit when Vs is low. The value of C_2 is a compromise between fast charging and low droop. A value of 10nF gives 1V/µs charge rate at 10mA and 1nA leakage gives –100µV/ms droop.

Benefits of op-amps such as the *TLE2027* in this circuit are clear; reduced gain/bandwidth, higher noise or offset would produce less precision in peak capture. Output current is important to charge C_2 quickly - the *TLE2027* sources or sinks 35mA from ±15V supplies, slews at 7.5V/µs and has a 75MHz gain/bandwidth product, all without large offset and offset drift and stability problems during transitions.

Testing took the form of rectangular 600mV pulses at the input to determine minimum pulse width for acquisition.

A_1	pulse width	R_3	C_3
TLE2037	1.0µs	470Ω	100pF
LF351	2.0µs	1kΩ	220pF
LM741	3.3µs	1.8kΩ	470pF



Newer, faster op-amps make more precise peak detectors that handle narrower input pulses – but there are points to watch.

CJD Catto
Elsworth
Cambridgeshire

Reference
Linear Applications handbook, National Semiconductor Corporation.

£100 WINNER – No 2

Meter resolves to 0.1dB at frequencies to vhf

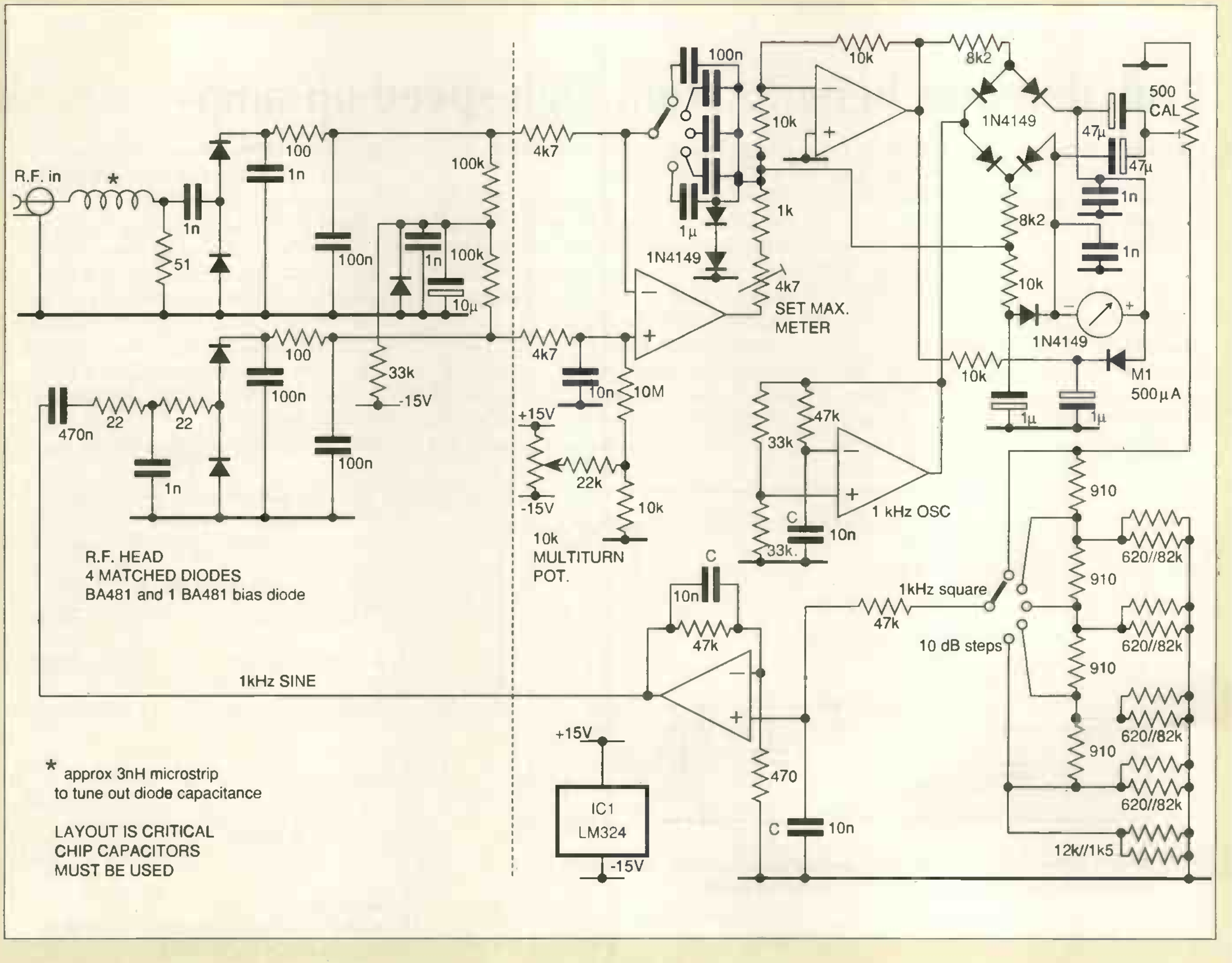
Mixer-type Schottky diodes can detect rf signals between around –35dBm and 20dBm. Response flatness is good to about 1GHz with a cheap *BA481* and better with higher frequency-response types; a general-purpose Schottky measures higher powers in a 50Ω system but with a reduced frequency response.

Since the diode output is temperature-dependent and somewhat variable between batches, the circuit shown uses two matched detectors. The second one is fed with a 1kHz sinewave, adjusted in amplitude by the error amplifier circuit until the outputs are balanced. A chopper composed of four *1N4149* diodes provides the 1kHz square, which is then formed into a sinewave by the op-amp. It also provides a dc level proportional to the amplitude. This is a measure of the rf input level and is read, in this case, on a moving-coil meter calibrated in decibels – though it could easily be digitised.

Ranges of 10dB are selected in the 1kHz drive to cover the 55dB dynamic range, and the error integrator capacitors also switched to provide the relevant time constant.

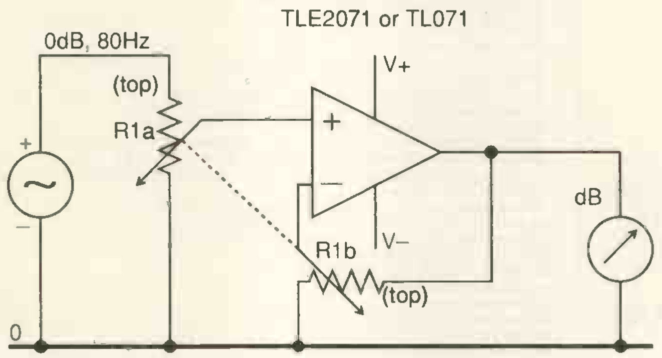
The 10kΩ multi-turn potentiometer zeroes the meter. Additional diodes on the meter are to cope with reverse drive if the integrator is zeroed too low, since the chopper provides only forward drive to the meter. The 4.7kΩ potentiometer on the integrator output sets maximum meter current. Accuracy to several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

P D Brooking
Ryde
Isle of Wight



Rf level meter using Schottky diodes is accurate into the vhf range, feedback removing the effects of temperature and diode variations.

Measuring ganged-potentiometer tracking



To select a ganged potentiometer with good tracking from a batch, this circuit avoids the need to test each one point by point. It also gives the results in decibels.

Instead of plotting the tracking characteristics of individual elements in ganged potentiometers, this circuit allows direct measurement in decibels.

In theory, attenuation introduced by R_{1a} is perfectly balanced by the gain determined by resistor R_{1b}, so that the output is constant at the input level, 0dB. In practice, departures from perfect tracking show up on the output meter in decibels relative to 0dB.

Since, at low frequencies, open-loop gain of any unity-gain-stable, fet-input op-amp is around 120dB, of which 60-80dB are needed, and since simple multimeters having a dB scale are only accurate at low frequency, the input should be 50-100Hz.

Tracking errors found can be related to attenuation by connecting another decibel meter between R_{1a} wiper and ground.

Wilfried Adam
Kappeln/Schiel
Germany

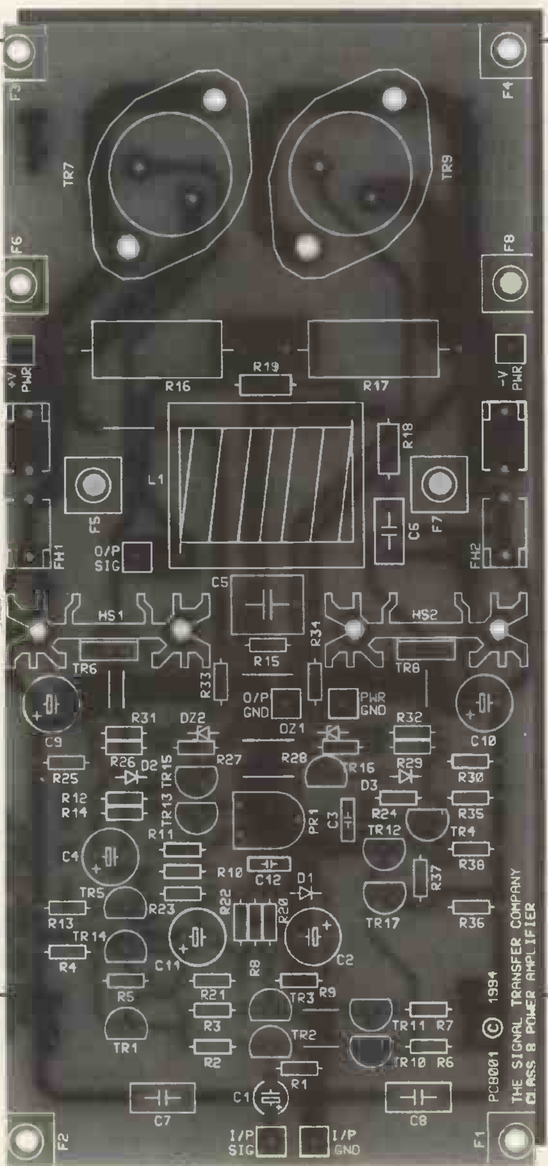
PCBs for Douglas Self's power amplifier series

Circuit boards for Douglas Self's high-performance power amplifier are now available via *EW+WW*.

Detailed on page 139 of the February 1994 issue, Douglas Self's state-of-the-art power amplifier is the culmination of ideas from one of the most detailed studies of power amplifier design ever published in a monthly magazine. Capable of delivering up to 100W into 8Ω, the amplifier features a distortion of 0.0015% at 50W and follows a new design methodology.

Designed by Douglas himself, the fibreglass boards have silk-screened component IDs and solder masking to minimise the possibility of shorts. Sold in pairs, the boards are supplied with additional detailed constructional notes.

Each board pair costs £45, which includes VAT and postage, UK and overseas. Credit card orders can be placed 24 hours on 0181-652 3614. Alternatively, send a postal order or cheque made payable to Reed Business Publishing to EW+WW, Quadrant House, The Quadrant, Sutton, Surrey SM2 5AS.



LETTERS

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

Powerful argument

Your report "50W psus cleaned up" (*Update*, December), contained some factually incorrect comments regarding power factor.

It is true that switched mode power supplies (smps) feed current harmonics into the mains supply. But those due to chopping action are at high frequencies and are more of a concern for conducted and radiated emission problems.

The power factor problem is due to the rectification of the ac mains to produce the dc voltage which feeds the dc-to-dc converter of the power supply.

A conventional full bridge rectifier-bulk capacitor circuit is normally used. The non-sinusoidal input current is because the rectifier diodes conduct around the voltage peaks of the mains voltage waveform when the instantaneous mains voltage exceeds that of the bulk dc voltage across the electrolytic capacitor. The waveform resembles a series of half sinusoidal current pulses coincident with the peaks of the mains voltage waveform, with a fundamental frequency equal to the mains frequency. Efficiency is also mentioned. Again, a lack of smps experience is evident. Power supply efficiency is measured as a ratio of output power to *real* input power (a typical figure might be 70% efficiency as quoted).

This method of calculation excludes the effect of power factor. If we use VA apparent power as the input power then the efficiency figure reduces significantly towards a figure of 40%.

Adding a power-factor correction circuit reduces the VA figure, improving power conversion efficiency. However, with the industry standard method of efficiency calculation, the efficiency will be worse. This is because the real power will be slightly higher due to losses in the power-factor correcting circuit. Where a switched mode pfc circuit is used, extra filtering will be required to cope with the extra noise, leading to further real power loss.

Introduction of power factor correction will ease the burden on power distribution networks but will create many headaches for the psu designer.

Andrew Wilkes
Berkshire

Feed-forward feedback

I should like to congratulate Giovanni Stochino on his fascinating article on error-correcting power amplifiers (*Audio design leaps forward*, *EW+WW*, October, pp.818-827). The concept of using a relatively tiny transformer for the output summation of two amplifiers by virtue of flux cancellation of the main output, and using the negative-feedback factor of the auxiliary amplifier to minimise core non-linearity is highly ingenious.

But the practical circuitry does end up being at least twice as complex as in a conventional amplifier – given output-injected error-correction this is probably inevitable. It also seems to demand a meticulous seven-step calibration procedure that makes setting the quiescent current of a class-B amplifier seem very quick and simple.

Regrettably Stochino does not give thd versus frequency curves so comparisons with other approaches are not easy to make. But from his figures for 20kHz, the distortion performance could be equalled by a conventional amplifier without much difficulty – though I readily concede that guaranteeing less than 0.003% at 20kHz (Stochino's high-bias condition) might require some form of two-pole compensation, with consequent careful checking of hf stability under all conditions.

One point puzzles me: thd for the uncorrected main amplifier is given as 0.3-1% at 20kHz, rather high for a circuit using 18 semiconductors. If the main amplifier could be made more linear, then the total output linearity might improve even more than proportionally, as the auxiliary amp would have less work to do, correcting lower initial distortion.

I imagine the fet output stage takes some of the blame for this, fets being much less linear in class B than bipolar devices.

Use of fets in the input stage also seems rather strange, given their poor V_{gs} matching. No doubt Stochino took these design decisions for very good reasons, and I hope he might expand on them, as in this application the design requirements may be subtly different from those of stand-alone class-B amplifiers such as I have discussed in the past.

I hope this is not taken as destructive criticism, as it is not intended as such and the article was most thought-provoking.

Last year I made some tentative investigations of my own into error-correction – being attracted like a moth to a flame by the prospect of abolishing distortion completely – and concluded that input-injected correction was probably the way to go.

I would be interested to know if Signor Stochino has any thoughts on this approach.

Douglas Self
London

Friendly interface user

Following John Davies' recent article on a simple I^{2}C interface for the pc (*I^{2}C via the pc*, December, pp. 994-996), I have laid out an equally simple pcb.

It is small enough to suit any D25 connector shell and we have fitted it into a modular adaptor (eg *RS 447-645*) which allows different cables to interface to different projects. If using an adaptor, the supplied crimp D25 connector needs to be discarded and replaced with a solder bucket type. The pcb then solders onto the buckets connections. A cheap four-wire RJ11 cable can then be adapted to connect to the I^{2}C devices.

For interested readers, I am happy to make the boards available at a nominal charge of a £1 coin plus a sae.

Simon Maddox
Maddox Broadcast Ltd
Unit 1
Stanley Centre
Kelvin Way, Crawley
West Sussex
RH10 2SE.

Polarised reflection

George Pickworth's articles on the electrolytic detector (*Detection before the diode*, December, pp. 1003-1006 and January, pp. 28-30) were most interesting. It is always a pleasure to read of our forbears' achievements, and to reflect on how much they could do with the knowledge they had and such materials as were available.

But I do not believe it necessary to invoke polarisation, by hydrogen layers or any other physical barrier, to explain the working of the electrolytic detector. When a soluble metal, such as copper, sits in an aqueous electrolyte comprising a salt of that metal – copper sulphate – the electrode will take up a definite 'equilibrium' potential with respect to the electrolyte. Forcing the electrode positive with respect to the equilibrium potential will cause current to pass from the electrode to the electrolyte, and copper metal into solution as copper ions. This is the 'unplating' reaction.

Forcing the electrode negative with respect to its equilibrium potential, will cause current to pass from the electrolyte to the electrode and copper ions to deposit on the electrode in the 'plating reaction'. These processes are linear, and so a cell comprising two copper plates in copper sulphate solution behaves as a resistor.

Operation of a platinum electrode in aqueous electrolyte is quite different. As the electrode is pushed positive from its (rather unstable) equilibrium potential, this noble metal is unable to pass into solution and so there can be no current. Nothing happens until the electrode is so positive that it is suddenly able to oxidise water, which it does by converting water into oxygen gas and hydrogen ions. The oxygen dissolves in the electrolyte and diffuses away, and the electrolyte is left a little more acid. If we now make the electrode slightly less positive, the current stops. The foregoing reaction can not run backwards because the oxygen has been lost, and it is in this sense that the electrode is irreversible.

Making the platinum electrode progressively more negative than its equilibrium potential means a point will eventually be reached where metal cations, eg from the lead counter electrode, will try to plate out on it. This must be avoided as it implies contamination of the platinum surface.

To make our electrolytic detector we need a large reversible counter electrode (such as a lead plate in lead sulphate) to effect a nice, stable, ohmic connection from the outside world to the electrolyte. The other connection requires a noble metal, such as platinum, to give an irreversible electrode, which should be small to keep the shunt capacitance down.

Applying a positive bias that just falls short of taking it into the oxygen-evolving state allows positive-going rf signal swings to cause the electrolytic diode to conduct. Negative signals take the small electrode down into a potential region where there are simply no chemical reactions available to carry any current. Unless the rf signal becomes so large that, on negative swings, the small electrode becomes plated with base metal originating in the counter electrode. The potential at which this happens may be regarded as the piv for the electrolytic diode. For a $Pb/PbSO_4/Pt$ diode the piv will be about 1.4V.

Working the diode with opposite polarity, on the verge of hydrogen evolution, is probably not a good idea. Hydrogen is not as soluble in water as oxygen and hydrogen polarisation becomes a possibility. More seriously, the platinum point will now be a cathode and vulnerable to plating-over with lead – even in the absence of an input signal.

P E K Donaldson
Sevenoaks

Detector detective

The two articles by George Pickworth have revealed some interesting aspects of the operation of electrolytic detectors. Unfortunately, both in the captions and in the text the term *barretter* was used as if it was a synonym for the electrolytic detector.

Fleming attributes the term barretter to Fessenden who used it to describe a thermal detector of radio waves.

A short loop of platinum wire a few millimetres long, and as small as 1.5×10^{-3}mm diameter, was exposed by dipping Wollaston wire in nitric acid to dissolve the silver. This fine filament was sealed in a glass bulb like an incandescent lamp, and the bulb was later evacuated and enclosed in a silver bulb to shield it from external radiation.

Ends of the loop were connected to a cell and headphones. Direct current passing through the circuit caused heating of the very fine wire and an increase in resistance. With suitable adjustment, a steady state was reached. Radio frequency energy coupled to the wire caused further heating and a sudden decrease in the steady current through the wire. This was detected as a sound in the headphones.

It was claimed that radio waves had been detected by this sort of device at distances up to 40km and, according to Fleming, both this and the electrolytic detector were suitable for telephony as well as telegraphy.

This device relied on the positive temperature coefficient of metallic conductors, rather than rectification, and the only thing it had in common with the electrolytic detector was that both employed a very fine platinum wire.

Fessenden seems to have used the term *liquid barretter* to describe devices which used a high resistance liquid enclosed in a fine tube or in which the two ends of the fine wire dipped into liquid[1, p213]. He may have done this because he advocated a thermal theory for the mode of action of the electrolytic detector[1, p217].

The term barretter was later used to describe a device which provided a constant current to a string of series-connected valve filaments. In the days before adoption of 240V ac as a standard for mains supplies this allowed sets to be connected to ac or dc mains and a range of voltages.

The term is appropriate because the device relied for its action on the positive temperature coefficient of a metallic filament. Steady rise in resistance with increasing current exerted a stabilising effect over a restricted range of applied voltages. Tables of barretters, or ballast tubes, and equivalents, are given by Norris[2]. Use of small filament bulbs to stabilise the output of audio oscillators is well known. Fleming may also have contributed the section on barretters of both types to the *Wireless Encyclopedia.* None of this detracts from George Pickworth's ingenuity in carrying out his investigations, but I think it would be preferable to stick to the term electrolytic detector for the class of devices he describes.

Les May

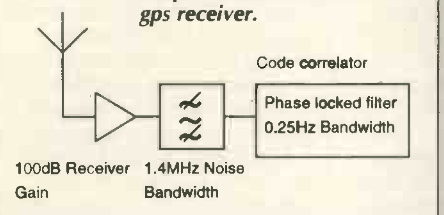
Further reading

1. Fleming J A. 'An elementary manual of Radiotelegraphy and Radiotelephony'. Longmans Green and Co. 3rd edition, 1919.
2. Norris R C (Ed). 'The practical radio reference book'. Odhams Press. undated but probably late 1940s.
3. Harmsworth's *Wireless Encyclopedia for the Amateur and Experimenter*. part 3.

False position on gps

'Jamming could undermine gps' (*Research Notes*, January) gave a false impression of the susceptibilities of gps receivers that might be employed in life-dependent instrument landing system (ils) and future microwave landing systems (mls). Assertions contained in the piece were based on published evaluations of 'two commercial receivers' and were not therefore indicative of the interference handling capabilities of receivers that might be used for aviation applications. The implication that the published findings place in doubt the use of gps for these roles was made without consideration of the characteristics of suitable receivers or what theoretical performance is available.

At Navstar Systems we manufacture a c/a code gps receiver (the *XR-5* series) which is involved in trials for commercial aircraft landing roles in the United States. Taking the system-guaranteed signal strength (for the l1 c/a code transmission) of –160dBW, above a 5° elevation angle (STANAG 4294) the carrier to noise ratios at 290K can be calculated to be:



Simple model of c/a code gps receiver.

No Jamming
Noiseless receiver:
c/no = –160 – (kTB)dBW = 50dB
Receiver with 4.5dB noise figure:
c/no = –160 – ((kTB)dBW + 4.5)
= 45.5dB
Critical c/no for reliable tracking
= 33dB

Tone Jamming
Tone spreading advantage (or correlation advantage) available to a spread spectrum receiver with 2.048MHz spreading bandwidth and 0.25Hz correlation bandwidth:
= 2.048MHz/0.25Hz = 69dB
Noiseless receiver:
Maximum received tone power
= (kTB)dBW + 69 + (50–33)
= –142 + 69 + 17
= –56.6dBW (–26.5dBm)
Receiver with 4.5dB noise power:
Maximum received tone power
= (kTB)dBW + 69 + (45.5–33)
=–142+69+ 12.5
=–61dBW (–31dBm)

These theoretically attainable figures compare well with the –103dBm loss of all satellites (las) susceptibility figure that is quoted in your article especially as they are calculated from the basis of the lowest (5° elevation angle) received powers from gps rather than more typical powers associated with high elevation satellites.

I doubt that the *XR-5* receiver is alone amongst units capable of realising very close to this theoretical limit of the c/a code system. In fact my company has demonstrated the loss of all satellites (las) threshold to occur at cw power levels between –33dBm and –23dBm. Such measured critical interference power levels may seem surprisingly high even with respect to the theoretically calculated levels shown above. The difference can be attributed to a combination of factors concerning the degree to which the theoretical calculation presents a worse case than the las criterion, and also because measured results benefit from higher than guaranteed power transmission from satellites.

Obviously these susceptibility figures transform the case for using gps for commercial aviation purposes to one well worth considering. This conclusion is especially valid when the use of gps ground 'pseudolite' transmitters are used in addition to, and in conjunction with, the chanellised Russian *Glonass* satellite navigation system (and others).

The suggestion that the use of such satellite navigation sensors is unsuitable for ils and mls replacement systems is not a responsible or well informed one.

Oliver Leisten
RF Engineering Group Leader
Navstar Systems
Northampton

Wire connection

I wish to thank J S Linfoot for his information regarding Woollaston wire (*Letters*, *EW+WW*, January, 1995). It not only explained how extremely fine platinum wire was produced, but gave a greater insight into the evolution of radio detectors.

From literature research, Fessenden discovered the barretter's principle while using nitric acid to dissolve the silver coating from platinum wire. I had been puzzled as to why the wire was coated with silver and why Fessenden should remove it; now I know.

Fessenden originally required extremely fine platinum wire for his 'hot wire' detector. This, as far as I can tell, was based on a design by Tesla whereby the wire formed one arm of a Wheatstone bridge. Radio frequency current heated the wire, thus changing its resistance and upsetting the bridge's balance.

George Pickworth
Kettering

Auf Wiedersehen sat

For this reader at least, it is becoming easier to understand the increased cynicism about cooperation between member states of the European Community. For some years, growing numbers of music lovers in the UK have enjoyed German digital radio via satellite – more especially so since the artistic and technical decline of BBC's *Radio Three*.

Access to all the major annual music festivals in Europe via the European Broadcasting Union has become a regular and much anticipated treat, most events normally ignored by the BBC.

Sales of the necessary tuners were actively promoted by a leading German company, Technisa,t after opening a branch in the UK. Now, despite active lobbying by various British MEPs the German Telecomms authority unilaterally closed down the satellite on New Year's Day, making thousands of pounds worth of equipment sold here useless overnight.

If any reader deceived in this way would like to contact me in writing via *EW+WW*'s editorial offices (enclose an sae) I will attempt to co-ordinate action in the European court. Or aggrieved ex-listeners can write direct to Gerd Tezner, Managing Director, Deutsche Bundespost Telekom, Godesberger Allee 87-92, 53105 Bonn 2, PO Box 2000 Germany.

Reg Williamson
Kidsgrove
Staffs

Shared experience

You don't often see shareware covered in either the trade press or newsstand magazines. Indeed, what little appears labelled as 'shareware' is more likely to be a mere demo program. I suspect that shareware authors' non-existent advertising budget explains this dearth of coverage.

Nonetheless, shareware is an important development and promises to become even more so as economical, high-speed digital networks proliferate. An unfortunate measure of its success is that certain unscrupulous marketers are constantly attempting to fob off demo programs and the like as shareware.

Is demo software an advertisement or is it shareware? Demo programs range from wretched slide shows to very sophisticated and well-supported 'educational' versions of programs (programs generally usable for small jobs but lacking the capacity or some features of the for-sale program). But in the middle of the spectrum, where the vast majority of demo programs lie, are various substandard programs masquerading as shareware. Personally, I think that most demo programs are a form of advertising, not shareware, and should be treated as such.

In the old days, back when floppy disks were expensive and I did not have my present, extremely well-paid position, I used to collect demo programs, erase the disks, and use them myself. Now that floppy disks are inexpensive, I think demo programs are useful only to very smart people who don't have anything important to occupy their time.

True shareware is a complete software package – including documentation – that anyone can obtain for no more than a modest media fee. And anyone can give the package to anyone else without fear in falling into software piracy.

If you try out the shareware and like it, you are honour-bound to send in a modest registration fee. Shareware is thus the only commercial exchange that I can think of that is based solely on faith in one's fellow human beings.

Shareware allows a small software firm to market its software very economically. It eliminates the usually perfectly ignorant middlemen, allowing you to deal directly with the software's author.

Using current software-marketing practices, a software vendor must charge you a minimum of $300 for a box of air if you buy his software over the counter. Most of the time, the author, or small team of authors, who actually wrote the software is not available if you contact the over-the-counter vendor looking for help or providing feedback.

Now that multitasking, message-passing operating systems such as Microsoft's *Windows* and Apple's *System 7* are the norm for pcs, software writers no longer need to concoct massive, do-all applications. Instead, they can write small, specialised programs for narrowly targeted markets. Such programs can use standard applications for the user interface, data acquisition, mathematical analyses, data-base functions, spreadsheet functions, graphing, and other i/o functions.

I expect small shareware shops to be able to compete quite readily with large software firms such as Microsoft.

High-speed digital networks should also open up consulting. Virtually all seers, prognosticators, pundits, and entrails readers see high-speed digital networks as a way to get at canned information stored on disk somewhere.

These people are as dead wrong as their 'information' is lifeless. Creative people such as engineers need knowledgeable minds on line. Having canned information on line is sufficient for handbook jockeys, but not for creative people.

The closer you are to the leading edge of the state of the art, the less likely you are to find even as much as a category for the information you need – let alone any substantive information filed away in that category.

Alas, the supply of talented, altruistic people who will help designers for free is sharply limited. So the opportunity also exists for on-line consultants to prosper. ■

Charles H Small
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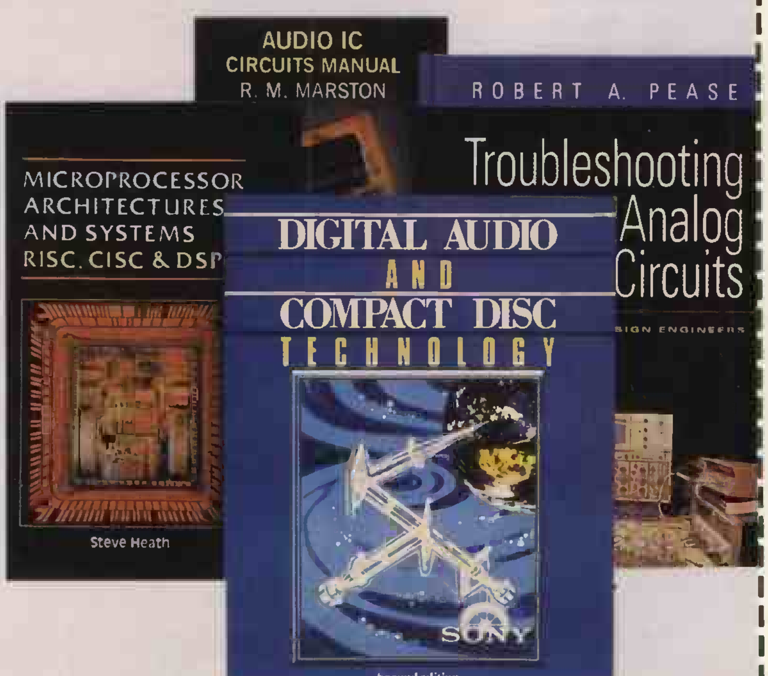

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UNUSUAL
oscillators and filters

Round up of simple but unusual filter and oscillator circuits for both af and rf. Created by Jim Williams of Linear Technology, these designs first appeared in a series called 'The Jim Williams Papers'.

Filters and oscillators share a common feature. They both deal with signals in the frequency domain. You can define a filter's function as rejecting the frequencies you do not want – a band-reject filter – or including only the frequencies you do want – a bandpass filter.

If you reorient your thinking, though, you will realise that all filters reject unwanted frequencies. When you view filters in this way, you see that any filter's function is the inverse of an oscillator's; oscillators synthesise individual frequencies or ranges of frequencies. Although there are more kinds of filters and oscillators than any magazine article of reasonable length can hope to touch on, herein are a few types of circuits that can meet a range of needs.

Figure 1a shows a highly selective bandpass filter using a resonant ceramic element and a single amplifier. Except at its resonant frequency, in this case, 400kHz, the ceramic element looks like a high impedance. For off-resonance inputs, IC_1 produces no output; it acts as a follower whose input is grounded.

At resonance, the ceramic element has a low impedance, and IC_1 behaves as an inverter with gain. The 100Ω resistor isolates summing point of IC_1 from the ceramic element's capacitance. This capacitance is quite substantial and limits the circuit's out-of-band rejection. **Figure 1b**, curve A, shows this effect. This plot shows very steep rejection, with output of IC_1 down almost 20dB at 300kHz and 40dB at 425kHz. The device's stray parasitic capacitance causes the gentle rise in the output at higher frequencies and also sets the –20dB floor at 300kHz.

Figure 2 shows how to use a nulling technique to partially correct problems caused by the ceramic elements parasitic capacitance. This circuit is similar to the previous one, except that a portion of the input goes to IC_1's positive input. The *RC* network at that input has an impedance close to the ceramic resonator's off-null impedance. Therefore, out-of-band components produce similar signals at IC_1 inputs and because of the device's com-

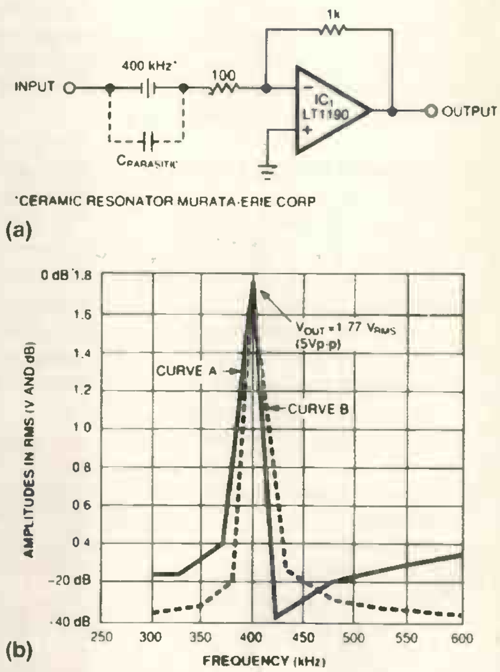


Fig. 1. One amplifier and a ceramic resonator create a bandpass filter (a). The solid curve of (b) shows the filters' response. Note the dip to –40dB on the high side of resonance. The dip is the result of the resonator's parasitic capacitance.

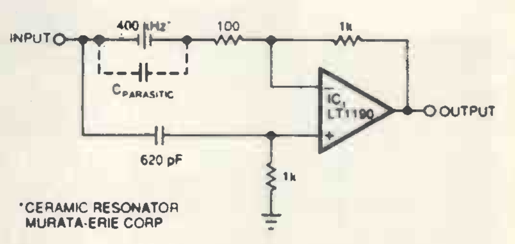
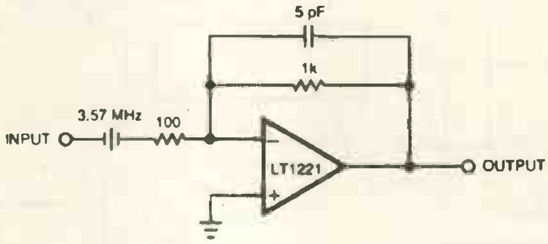


Fig. 2. A slight modification of the circuit in Figure 1a allows you to cancel out the effects of the resonator's parasitic capacitance. The dashed curve of Fig. 1b shows the effects on the filter response. Below resonance, the modified circuit attenuates by an extra 20dB. Above approximately 525kHz, the improvement is even more dramatic.

Walt Jung and James Wong are with Analog Devices Inc. This article first appeared in EDN.

Fig. 3. Replacing the ceramic resonator of Fig. 1a with a 3.57MHz crystal is the most significant change that leads to this crystal filter (a). You can see the crystal filter's response in b.



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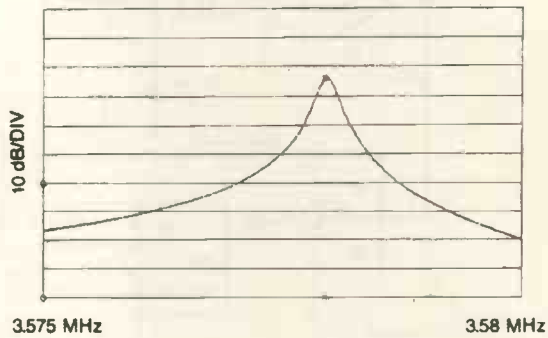

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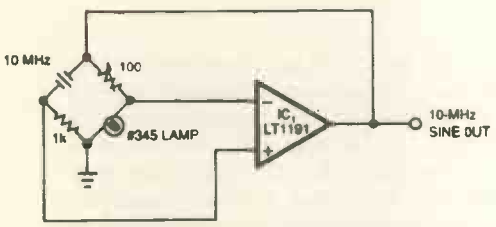


Fig. 4. An incandescent lamp's current-dependent resistance stabilises the oscillation amplitude of this 10MHz crystal oscillator

mon-mode rejection, produce little output. At resonance, the added *RC* network appears as a much higher impedance than the ceramic element, and the filter response is similar to that of the circuit in **Fig. 1a**, curve B. Figure **1b** shows that this circuit has much better out-of-band rejection than the earlier circuit. The high-frequency roll-off is smooth and at 475kHz, over 20dB deeper than that of the circuit in Fig. 1a. At 375kHz and below, on the low-frequency side of resonance, the circuits behave similarly.

By using quartz crystals, you can make filters whose high-frequency selectivity is even higher than that of filters based on ceramic resonators. **Figure 3a** replaces Fig. 1a's ceramic element with a 37MHz quartz crystal. Figure **3b** shows almost 30dB of attenuation only a few kilohertz on either side of resonance. The differential nulling technique used with the ceramic elements is less effective with quartz crystals. Crystals have significantly lower parasitic capacitance, making the cancellation less effective.

Oscillators use crystals and resonators

The circuit in Fig. 4 places a crystal within the amplifier's feedback path, creating an oscillator. With the crystal removed, the circuit is a familiar non-inverting amplifier with a grounded input. The impedance ratio of the elements associated with IC_1's negative input sets the gain. Inserting the crystal closes a positive feedback path at the crystal's resonant frequency and oscillations commence.

In any oscillator, you must control the gain as well as the phase shift at the frequency of interest. If the gain is too low, oscillation will not occur. Conversely, too much gain produces saturation limiting. In this circuit, gain control comes from the positive temperature coefficient of the lamp at IC_1's negative input. When you first apply power, the lamp's resistance is low, the gain is high, and the oscillation amplitude increases. As the amplitude builds, the lamp current increases and causes heating, which raises the lamp resistance. The increased resistance reduces the amplifier gain and the circuit finds a stable operating point. This circuit's sine-wave output has all of the stability advantages associated with quartz crystals. Although shown with a 10MHz crystal, the circuit works well with a variety of crystal types from 100kHz to 20MHz. Using a lamp to control the amplifier gain is a classic technique, first described by Meacham in 1938. Electronic gain control, though more complex, offers more precise amplitude control.

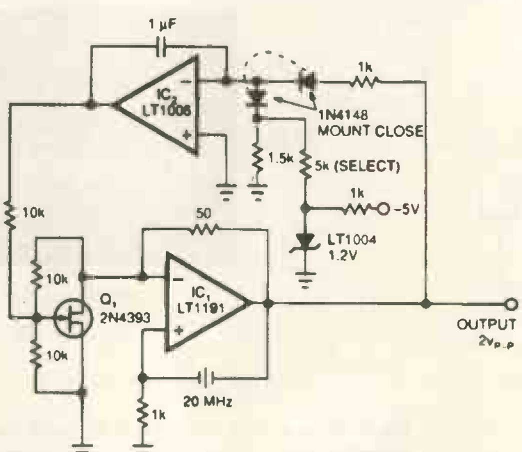
The quartz stabilised oscillator of **Fig. 5a** replaces the lamp with an electronic amplitude-stabilisation loop. IC_2 compares the IC_1 oscillator's positive output peaks with a dc reference. The diode in the dc reference path compensates for the rectifier diode's temperature dependence. Op-amp IC_2 biases Q_1, controlling the fet's channel resistance and influencing the loop gain.

Amplitude of the oscillator's output is a reflection of the loop gain. Loop closure around IC_1 stabilises the amplitude of the oscillator's output; the 1µF capacitor compensates the gain-control loop.

The dc reference network provides optimum temperature compensation for the rectifier diode, which sees IC_1's 2V pk-pk, 20MHz output waveform. IC_1's small output swing minimises the distortion attributable to channel-resistance modulation in Q_1. To use this circuit, adjust the 50Ω trimmer until 2V peak-to-peak oscillations appear at IC_1 output.

Figure **5b** is a spectrum analysis of the oscillator's output. The fundamental is at 20MHz; the second harmonic, at 40MHz, is 47dB down. The third harmonic, 50dB down,

Fig. 5. An electronic gain-control circuit that uses the voltage-controlled on-resistance of a fet stabilises the output amplitude of this 20MHz crystal oscillator (a). In (b), you see that the output's harmonics are at least 47dB below the fundamental.



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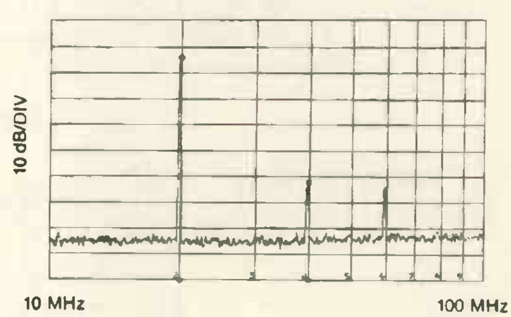

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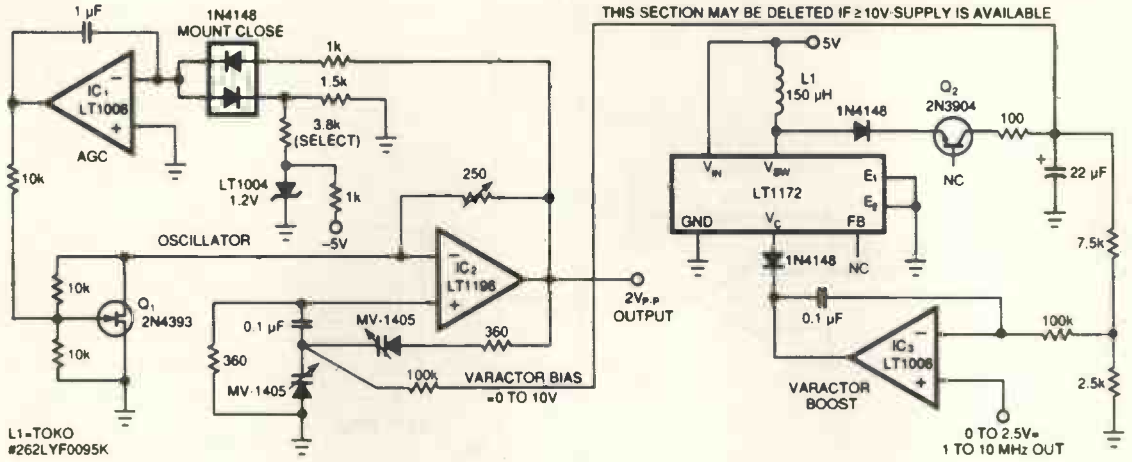

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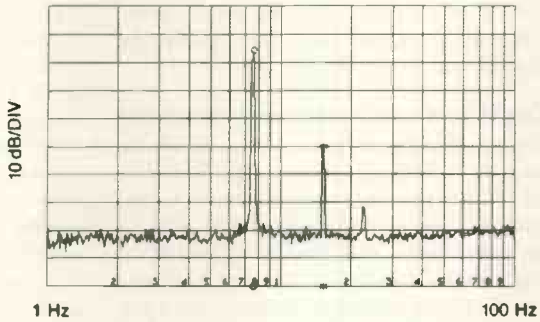


(b)

Fig. 6. A pair of varactor diodes lets you tune this Wien-bridge oscillator (a) from 1MHz to 10MHz by applying a 0 to 10V signal. Adding the components in the right half of the schematic lets you operate the circuit from a 5V supply and permits controlling the frequency with a 0 to 2.5V signal. The spectrum analysis in (b) shows that the sinusoidal output is clean.

occurs at 60MHz. Resolution bandwidth for the spectrum analysis is 1kHz.

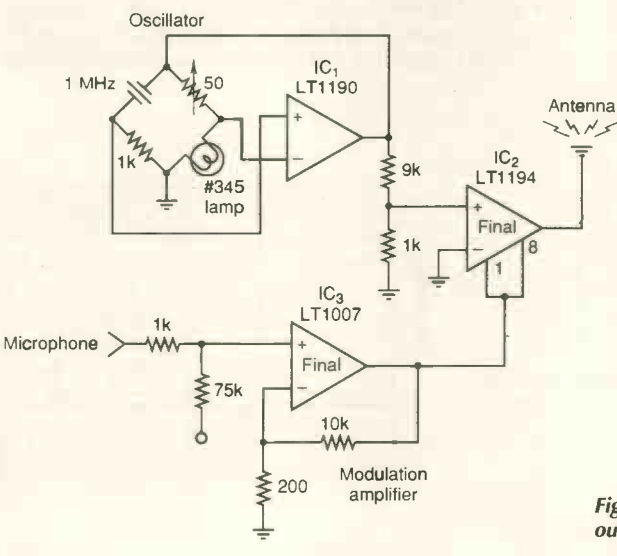
The circuit in **Fig. 6a** replaces the quartz crystal with a Wien network at IC_2's non-inverting input. IC_1 controls Q_1 to stabilise the amplitude of IC_2's oscillations. Operation is identical to that of the circuit in the previous figure.

Although the Wien network is not nearly as stable as a quartz crystal, it has the advantage of a variable frequency output. Normally, you vary the frequency by varying either R or C or both. The use of manually adjustable elements, such as dual potentiometers and two section variable capacitors is common. The circuit in Fig. 6a uses fixed, 360Ω Wien-network resistors and varactor diodes as capacitors. Voltage-variable capacitance of the varactor diodes allows dc tuning of the oscillator. Applying 0 to 10V dc to the varactors shifts the oscillation frequency from 1 to 10MHz. The 0.1µF capacitor blocks the dc bias from IC_2's positive input but lets the Wien network function normally. Op-amp IC_2's 2V peak to peak output minimises the varactors' junction effects and thereby limits distortion.

This 5V powered circuit requires a voltage step-up to develop adequate varactor drive. IC_3 and the *LT1172* switching regulator form a simple voltage step up regulator. Op-amp IC_3 controls the *LT1172* to produce whatever output voltage is required to close a loop at the negative input of IC_3. The 22µF output capacitor stores high-voltage inductive-flyback pulses from L_1 after they have been rectified by the diode-and-zener-connected Q_2.

The 7.5kΩ/2.5kΩ divider closes the loop by providing a sample of the output value to IC_3's negative input The 0.1µF capacitor stabilises this feedback action. Transistor Q_2's zener drop allows the circuit to produce controlled outputs at voltages as small as zero. This arrangement permits a 0 to 2.5V input at IC_3 to produce a corresponding 0 to 10V varactor bias. Figure **6b**, a spectral plot of the circuit running at 7.6MHz, shows second harmonic down 35dB and the third harmonic down almost 60dB. Resolution bandwidth is 3kHz.

Figure 7a shows the schematic of an am radio station – complete from microphone to antenna. Op-amp IC_1, set up as a quartz stabilised oscillator generates the carrier. Op-amp IC_1's output feeds IC_2, which functions as a modulated rf power output stage. The bias applied to offset pins 1 and 8 restricts IC_2's



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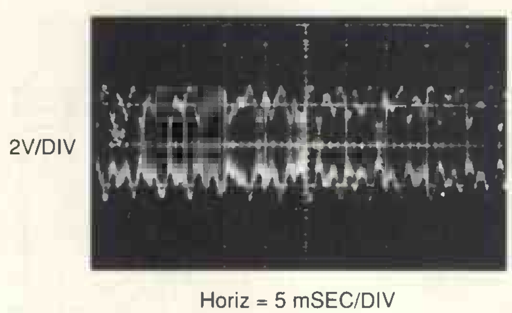


(b)

Fig. 7. This circuit is a complete am radio transmitter. Modulated output, at (b), is Chuck Berry playing 'Johney B Goode'.

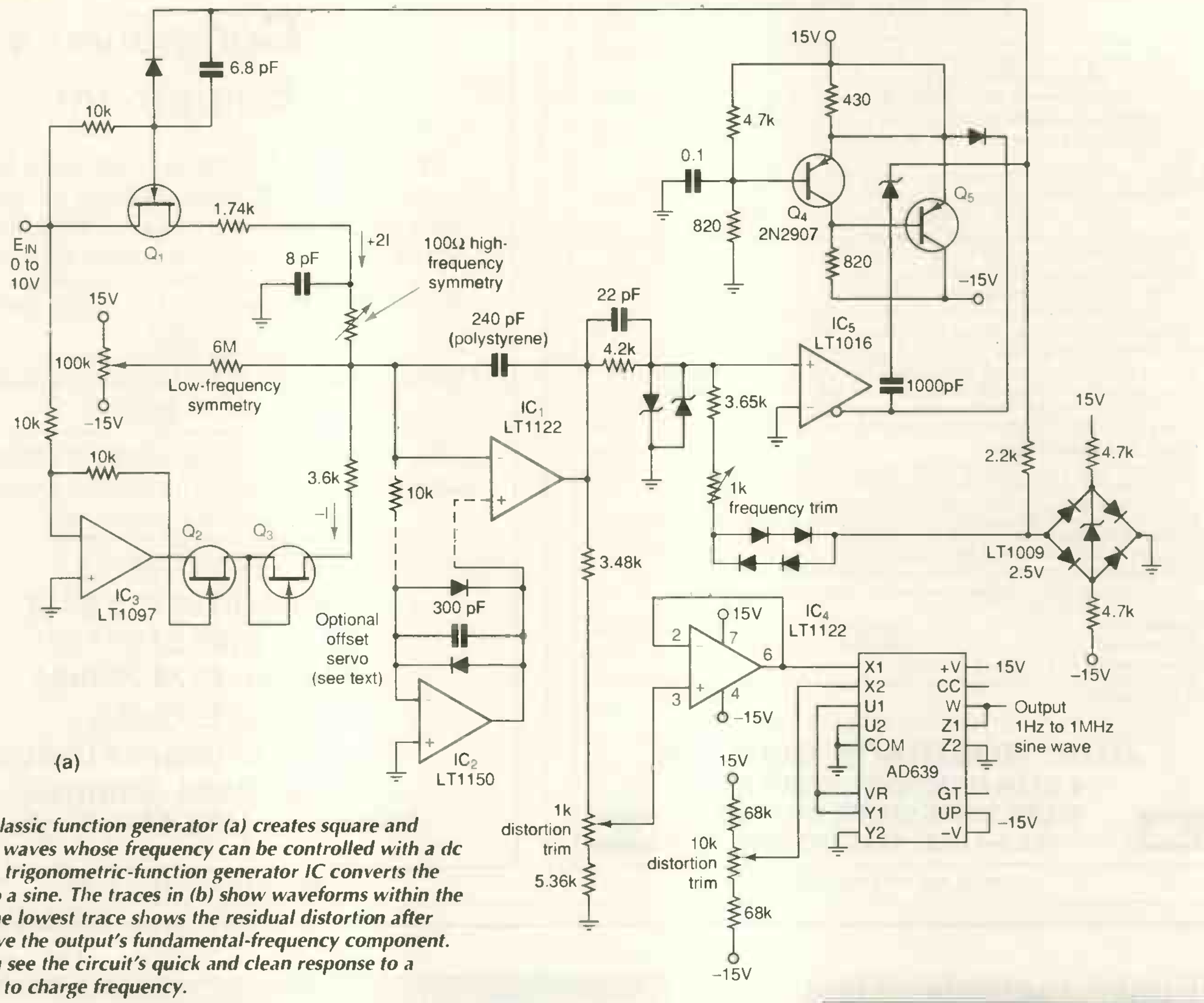


Fig. 8. A classic function generator (a) creates square and triangular waves whose frequency can be controlled with a dc voltage. A trigonometric-function generator IC converts the triangle to a sine. The traces in (b) show waveforms within the circuit. The lowest trace shows the residual distortion after you remove the output's fundamental-frequency component. In (c), you see the circuit's quick and clean response to a command to charge frequency.

input-signal range. There are more details in the *LT1194* data sheet. Op-amp IC_3, a microphone amplifier, supplies bias to the offset pins, resulting in an amplitude-modulated rf carrier at IC_2's output. The dc voltage summed with the microphone output biases IC_3's output to the appropriate level for good quality modulation characteristics.

Calibrating this circuit involves trimming the 100Ω potentiometer in the oscillator for a stable 1V pk-pk 1MHz output from IC_1. Figure. 7b shows a typical am carrier output at the antenna.

Start with a triangle; end up with a sine

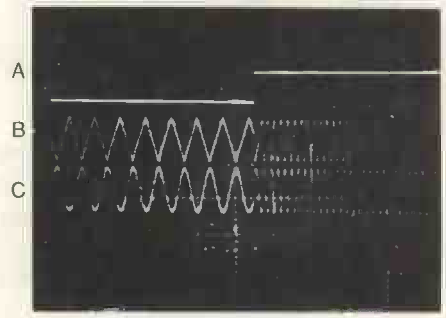
To this point, the oscillators presented have limited tuning-frequency range. Although the circuit in **Fig. 8a** is not a true oscillator, it produces a synthesised sine-wave output over a wide dynamic range. Many applications such as audio, shaker-table driving, and automatic test equipment require voltage-controlled oscillators with sinewave outputs. This circuit meets this need, spanning a range of 1Hz to 1MHz, equal to six decades or 120dB, for a 0-10V input. The circuit maintains 0.25% frequency linearity and 0.40% distortion.

To understand the circuit, assume Q_5 is on and it's collector (Fig. **8c**, trace A.) is at −15V, cutting off Q_1. Op-amp IC_3 inverts the positive input voltage and biases the summing node of integrator IC_1 through the 3.2kΩ resistor and the self-biased fet's. It pulls a current, $+I$ from the summing point. Precision op-amp IC_2 provides dc stabilisation of IC_1.

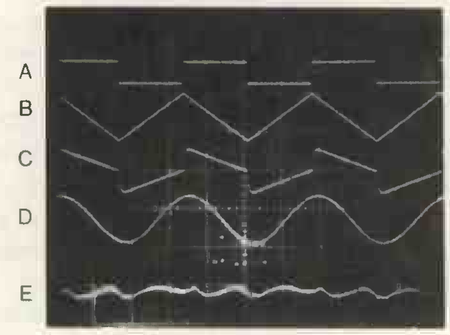
Output of IC_1 in trace B ramps positive until IC_5 input, trace C, crosses zero and causes IC_5's inverting output to go negative. The Q_4/Q_5 level shifter then turns off and Q_5's collector goes to +15V, allowing Q_1 to turn on.

Values of the resistors in Q_1's path result in a current, $+2I$, exactly twice the absolute magnitude of the current, $+I$, that flows out of the summing node. As a result, the net current into the junction becomes $+I$, and IC_1 integrates negatively at the same rate it did during its positive-going excursion.

When IC_1 integrates far enough in the negative direction, IC_5's non inverting input crosses zero and the circuit's two outputs change state. The state changes, switches the Q_4/Q_5 level shifter's state, causing Q_1 to go off and the entire cycle to repeat. The result is a triangular waveform at IC_1's output. The frequency of this triangle depends on the circuit's input voltage and varies from 1Hz to 1MHz with a 0 to 10V input The *LT1009* diode bridge and the series-parallel diodes provide a stable bipolar reference that always opposes



HORIZ = 10 µSEC/DIV
A = 5V/DIV B = 5V/DIV C = 5V/DIV



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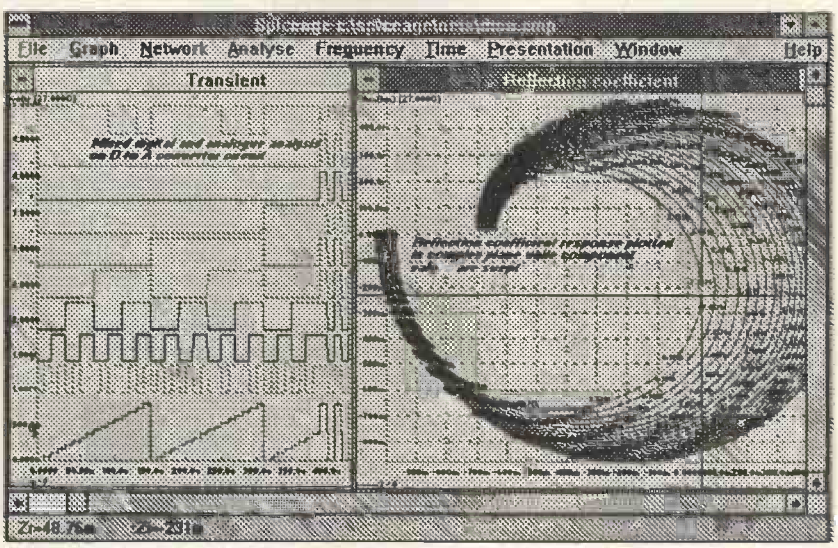

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ACTIVE

A-to-d and d-to-a converters

14-bit, 20MHz a-to-d. Analogic's *ADC3120* sampling a-to-d converter gives 14-bit resolution at conversion rates up to 20MHz; spurious-free dynamic range is 90dB. It will digitise a 10MHz signal at 20MHz sampling rate with no missing codes and exhibits a 75dB signal-to-noise ratio with inputs in the 0-10MHz range. Analogue input range is ±1.28V and the output is offset binary or twos-complement form. Total harmonic distortion is –85dB. Thame Components Ltd. Tel., 01844 261188; fax, 01844 261681.

14-bit a-to-d. *ADS-916* is a 14-bit, 500kHz a-to-d converter, pin-compatible with the earlier *ADS-926* and the 1MHz *ADS-927*. No missing codes are guaranteed to the 14-bit level over the military temperature range and signal-to-noise ratio is 77dB with thd of –86dB. It operates from either ±15V or ±5V and dissipates 1.6W. Conversion starts at the rising edge of a pulse and there is no pipeline delay. Datel (UK) Ltd. Tel., 01256 880444; fax, 01256 880706.

Digital signal processors

FFT processor. *PDSP16515A* from GEC Plessey is a stand-alone Fast Fourier Transform processor with increased accuracy of 18bits, compared to its predecessor's 16bits. It carries out forward or inverse ffts on complex or real data sets of up to 1024 points in 98μs; an internal ram holding up to 1024 complex data points removes the memory transfer bottleneck found in other solutions, and multiple devices can be paralleled to increase sampling rate up to the 40MHz system clock speed. The device will simultaneously accept new data, transform stored data and output previous results. Either Hamming or Blackman-Harris window operators are applied internally, operator values being calculated internally with no external rom. GEC Plessey Semiconductors Ltd. Tel., 01793 518510; fax, 01793 518582.

Discrete active devices

Avalanche mosfet. Zetex avalanche mosfet, *ZVN4206GV*, needs no diodes or vdrs to protect it against load dumps and transients. It drives relays at up to 600mA continuous and avalanche breakdown dissipates any energy stored in the relay's magnetic field to a maximum of 5mj. There is no need for interface components between logic and the mosfet driver, which has a high input impedance, V_{DS} of 60V and pulsed I_{DS} 8A. Turn on and off times are 8ns and 12ns. Zetex plc. Tel., 0161-627 5105; fax, 0161-627 5467.

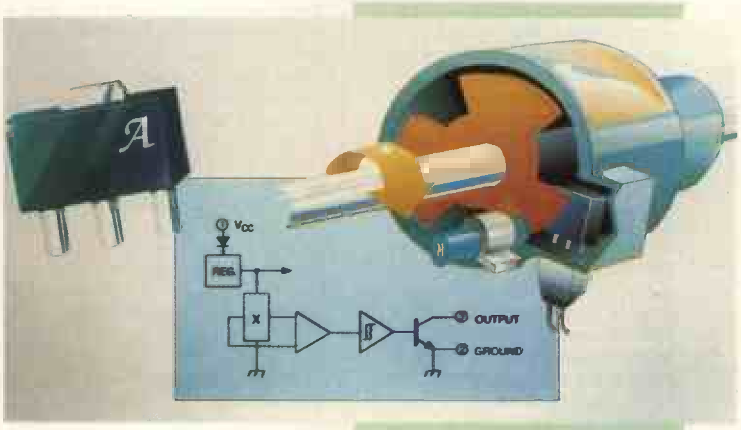
'Hard' power mosfets. Fourteen new *Hexfet* power mosfets from IR are qualified to JAN-S radiation hardness standards up to 1000Krad. N-channel devices are available in 100V, 200V and 500V types at 5.5-34A; p-channel fets come in –100V and –200V versions at currents in the 4-15A range. International Rectifier. Tel., 01883 713215; fax, 01883 714234.

Memory chips

Pentium caches. IDT offers three families of low-profile, plug-in cache modules to provide the full range of cache options for 3.3V Pentium systems from Intel, VLSI and OPTi. *IDT7MPV61xx/62xx* modules contain everything for a complete cache system on a 160-lead card-edge pcb. They operate at up to 66MHz and options include 256Kbyte and 512Kbyte sizes with synchronous burst or asynchronous working. Integrated Device Technology. Tel., 01372 363734; fax, 01372 378851.

Microprocessors and controllers

Temperature measurement. OKI microcontrollers *MSM64164-001/002* are 4-bit types for temperature measurement and energy consumption meters. They need no code development, since the relevant code already is held in rom. Both drive an eight-digit lc display with 20 flags, the time display including hour, minute, am/pm and 12/24h selection with alarm. Temperature measuring range is –40°C to 70°C and 60° to 200°C with 1% resolution and ±1°C accuracy. Both ranges are convertible to Fahrenheit. A memory function for min/max is included and readings are transmitted via a serial port. Gothic Crellon Ltd. Tel., 01734 788878



Motor control. NEC's *μPD78366* 16-bit microcontroller is expressly designed for pulse-width control of three-phase motors, its speed allowing repetition frequencies above the audio range. Since, at 20kHz rep. rate, only 16% of its cpu power is in use, capacity is available for tasks such as interfacing with the axis-control computer or controlling other sub systems. NEC Electronics (UK) Ltd. Tel., 01908 691133; fax, 01908 670290.

Mixed-signal ICs

'Smallest' PCMCIA power switch. Claimed to be the smallest PCMCIA power interface switch to comply with the new voltage switching standard, the Siliconix *Si9712DY* is a single chip replacing up to nine discrete components and drivers. It switches 0V, 3.3V or 5V to the PCMCIA slot and 3V, 5V or 12V to the card with on resistance of 111mΩ for 5V and 85mΩ for 3.3V on the V_{cc} supply voltage. Ramp time on power-up is 2ms, avoiding spikes on the system supply. Internal logic allows interfacing between active-high and active-low control signals for wide ranging compatibility. Siliconix/Temic Marketing. Tel., 01344 485757; fax, 01344 427371.

Brushless motor controller. Micro Linear's *ML4425* is a single-chip controller for brushless dc motors, containing all the circuitry needed to start the motor and maintain speed to within 5%, forward or reverse, without the assistance of either microprocessor or Hall-effect sensors. A patented position sensor integrates back emf to eliminate noise. Ambar Components Ltd. Tel., 01844 261144; fax, 01844 261789.

Hall-effect switch ICs. Meant for use in automotive applications, the *Type 3121/2/3 and 3141/2/3/4* Hall-effect switches operate at up to 150°C continuously with high stability, all types having a unipolar characteristic. Each device has a voltage regulator to allow operation from 4.5V to 24V, a reverse-battery protection diode, quadratic Hall voltage generator, temperature compensation, amplifier, Schmitt trigger and open-collector, 25mA output. Allegro MicroSystems Inc. Tel., 01932 253355; fax, 01932 246622.

Optical devices

Flush front-panel leds. *Dialight's* new pcb-mounted leds have 4mm flat tops to fit flush with front panel, but still have a viewing angle of 135° and luminous intensity of 7mcd (red and green) and 10mcd (yellow). The leds are single or dual types, the dual variety allowing mixed colours. Forward voltages are 2.2V for red and green and 2.3V for the yellow type. Dialight. Tel., 01638 665161; fax, 01638 660718.

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PASSIVE

Connectors and cabling

Universal zif socket. From Aries, the *Universal SOIC/TSOP II* programming/test/burn-in socket handles over 50 different package sizes, including all SOIC and TSOP type II zero insertion-force types on a 1.27mm pitch, including gull-wing and J-lead devices with up to 44 leads. It is adjustable to take out-of-spec packages and can be adapted for custom and hybrid devices. Aries Electronics (Europe). Tel., 01908 260007; fax, 01908 260008.

Extender extender. To protect and extend the top connector of its *AT-EXT* active pc extender board, Amplicon Liveline has introduced a connector extender that raises the test connector 1in above the computer's top cover, which then has a slot to take it and brackets to hold it firmly. Boards for test can then simply be inserted in the *AT-EXT* extender, which is in place permanently. The PC-AT powers up sequentially, is fully buffered and has over-current protection. Amplicon Liveline Ltd. Tel., 0800 525 335 (free); fax, 01273 570215.

Displays

21in colour plasma display. *FPF21C8060UA* is claimed by Fujitsu to be the world's first 21in colour plasma display. It has 260,000 colours selected by 900,000 64-grey-scale sub-pixels on a 0.22mm pitch. White peak luminance is 180cd/m^2, contrast ratio 60:1 and viewing angle 140°. Thickness of the display is 32mm and it uses 100W. Fujitsu Microelectronics Ltd. Tel., 01628 76100; fax, 01628 781484.

Filters

Network interface filters. Bandpass network interface filters in the *NIF* series by Newport complete the company's collection of components for the Echelon local operating networks. Two models are available: *NIF78* copes with a data rate of 78kb/s at a centre frequency of 90kHz and a 3dB bandwidth of 25-250kHz; and *NIF125* centres on 1.7MHz for a data rate of 1.25Mb/s at 0.2-9.6MHz 3dB bandwidth. Both types have protection diodes to reduce transients on the twisted pair cable. Newport Components Ltd. Tel., 01908 615232; fax, 01908 617545.

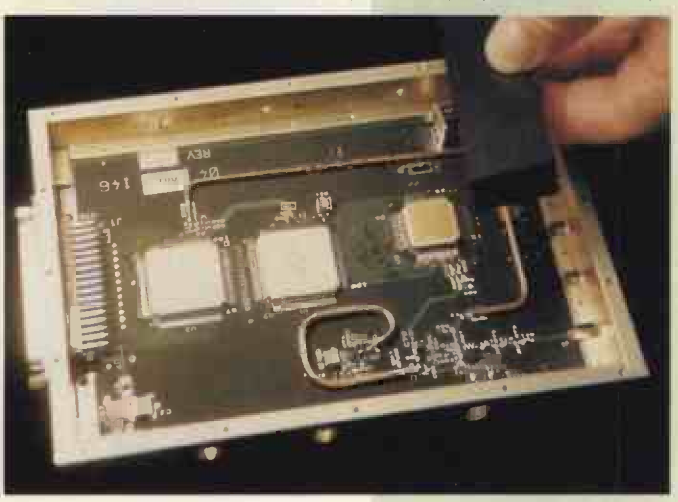
Hardware

Screened cases. Vero's *EBX* range of 274mm wide rfi-screened desktop instrument cases are finished in black epoxy powder and are available in three heights of 50, 75 and 100mm and in 200 and 300mm depth. Rfi screening is held, although there is ventilation and access to the interior. Standard rf attenuation is 40dB, but optional beryllium copper fingers increase that to 55dB. Cases have guides for 6U double Eurocards and an optional mounting kit is available for other sizes. Vero Electronics Ltd. Tel., 01703 266300; fax, 01703 265126.

European/US fuseholder. A fuseholder for pcb mounting from Schurter takes either 5 by 20mm or 6.3 by 32mm types, the design of the *OGD* ensuring that only one fuse can be inserted at any time. Kinked terminals allow single-step installation and the units are insulated for use on crowded boards. Holders are rated for 10A fuses and are approved internationally. Radiatron Components Ltd. Tel., 01784 439393; fax, 01784 477333.

Radio communications products

Clean DDS. Sciteq's *ADS-631/632* are broadband direct digital synthesisers with a clock rate of 500MHz to provide <0.12Hz resolution over a 230MHz bandwidth under 32-bit emitter-coupled logic control. Depending on the output filter, worst-case spurs can be better than a guaranteed –65dB for a >100MHz bandwidth. *ADS-631* is the basic model and needs an external clock, while *ADS-632* has a 500MHz clock generator, which is lockable to an external 10MHz reference or an internal 1ppm 10MHz reference. Tuning takes place in less than 0.1µs. Lyons Instruments Ltd. Tel., 01992 768888; fax, 01992 788000.



Ribbon cable assemblies. New 3M cable harness assemblies use pleated-foil-covered ribbon cable and miniature delta ribbon connectors with either two-piece metal or overmoulded metal-junction shells. The metal-shell assemblies have a squeeze-to-release latch for reliable, repetitive operation and the moulded type thumbscrew fixings with a choice of metric and unified threads. Both shell types accept 50Ω or 75Ω controlled-impedance PFC cable, handling data rates up to 200Mb/s, and the assemblies afford shielding of over 90dB in the range 30MHz-1GHz. Pin counts of 20-68 are available in both types of assembly. Low dielectric constant of the thermoplastic elastomer cable insulation reduces propagation delay and attenuation. 3M United Kingdom plc. Tel., 01344 858000; fax, 01344 858758.

Instrumentation

Digital storage oscilloscopes. Tektronix *TDS300* series digital storage oscilloscopes use the Tek digital real-time technique to produce a display that looks and responds in the same manner as an analogue type, supporting 200Msample/s for the *TDS310*, 500Msample/s for the *TDS320* and 1Gsample/s for the *TDS350*. The instruments are able to acquire any signal up to the full bandwidth in real time, including transients as fast as 2.5ns and single-sweep events. Thurlby Thandar Instruments Ltd. Tel., 01480 412451; fax, 01480 450409.

Probe adaptors. To ease the task of connecting oscilloscopes and logic analysers to microprocessor chips with up to 304 pins, H-P offers two probe adaptors capable of probing 0.5mm and 0.65mm pitch plastic and ceramic quad flat pack sm devices. In both types, a base is attached to the pcb around the chip with either adhesive or four screws and the adaptor slid over the chip to contact the base. Flexible and rigid adaptors are available to fit the probe adaptors and H-P instruments, up to four being used to probe all pins at once. The adaptors have a bandwidth of 600MHz and inter-contact capacitance of 2pF. Hewlett-Packard Ltd. Tel., 01344 362277; fax, 01344 362269.

Lan cable tester. Wavetek's *LANTEK PRO* cable testers carry out a complete performance verification in 40 seconds, including crosstalk at both ends. A one-button Autotest facility performs a series of tests on line mapping, signal attenuation, dc loop resistance, mutual capacitance and cable length and calculating attenuation:crosstalk ratio over the 100kHz-100MHz frequency range, displaying pass/fail reports on a graphical lcd screen. Battery life is 10-12h/charge. Wavetek Ltd. Tel., 01384 442394; fax, 01384 44025.

EMI tracer. *ScanEM* is a self-contained, hand-held near-field emi tracer by CTI that detects the presence of an electromagnetic field and gives audible and visible indication of its strength. Tor Applied Technologies Ltd. Tel., 01455 844114; fax, 01455 844116.

Power meter. Yokogawa's *2534* is a digital power meter having a maximum error of 0.25% of full scale at 50/60Hz. Its use of digital sampling allowing it to measure the non-sinusoidal waveforms found in switched-mode power supplies and inverters, and also direct current and power. Three displays give readings of voltage, current and power, In-built circuitry integrating the measurements to show energy consumption, active power, reactive power and phase angle. Current transformers are available and the instrument will also display frequency in the 4Hz-22kHz range. Martron Instruments Ltd. Tel., 01494 459200; fax, 01494 535002.

Accelerometer calibration. Instead of dropping balls for shock calibration of accelerometers, the Endevco *2925 POP Comparison Shock Calibrator* fires them pneumatically from a barrel at an anvil on which the accelerometers are mounted. The

Analogue signature analysis. Polar Instruments offers the *T2500*, an entry-level analogue signature analyser to trace faults at component level on pcbs. It displays good and bad signatures in true dual-trace manner and carries out its functions without the need to power the board and without any documentation, on a component-by-component basis. The user simply applies an ac signal, limited in both current and voltage, via a probe to components and views the resulting signature, which is characteristic of the component's behaviour, on screen; if a known good board is available, in dual-channel mode. Comparison is either visual or automatic, with pass or fail messages. An upgrade is available to scan 40 ICs in the same manner. Whingate Test Services. Tel., 01202 605239; fax, 01202 691118.

instrument produces accelerations in the 10-10,000g range with combinations of anvils, masses and pressure settings. Wide pulse durations from the test ease analysis and reduce the need for filtering. Operation is manual or by computer control. Endevco UK Ltd. Tel., 01763 261311; fax, 01763 261120.

Waveform recorder. The *Hioki 8804* B5 book-sized portable waveform recorder is a development of the company's earlier instruments, this one having a 5in backlit lc display to allow selection and printout of the required section on 60mm thermal paper to a resolution of 8 dots/mm. Inputs are two analogue channels with a 0-100kHz bandwidth and eight logic channels. Sampling rate is 400ksample/s with a 64K memory capacity. Also included is a true rms recording function. Telonic Instruments Ltd. Tel., 01734 786911; fax, 01734 792338.

Radiometers. Ultra Violet Products offers a range of *Multi-Sense* radiometers to measure ultraviolet, infrared and visible light and magnetic fields. Multi-Sense 100 is a basic instrument for intensity and dosage calculations and Multi-Sense 200 uses a Psion II hand-held computer to perform data storage and manipulation. Both instruments use the company's *SEN100* and *200* plug-in sensors, the latter being fitted with interference filters for research and production use. A feature of the ultra-violet measurement is that it is direct and consequently more accurate than the usual method of measuring from fluorescence. Ultra Violet Products Ltd. Tel., 01223 420022; fax, 01223 420561.

Optical reflective sensors. *VLS* visible-light sensors by Compact Instruments are of the reflective type, offering an operating distance of 1m and an operating angle of ±45° to the reflective tape surface. There is also the *DP6 Dataprobe* for non-contact speed and rotation sensing, a 12V device. Voltages of 5V and 7-40V dc energise the *VLS5/7* versions. Use of visible light eases alignment and a led indicates signal reception. Mountings are available. Compact Instruments Ltd. Tel., 01204 532544/5; fax, 01204 22285.

Colour notebook 'scope. Hitachi Denshi has the second in its series of 'notebook' oscilloscopes, the *VC-5410* being a 20MHz, dual-channel, 15Msample/s instrument that will store up to 10 waveforms of 2Kword each, with backup. Display is a thin-film transistor lc type with fluorescent back light, providing 160 by 220 dots in three colours and giving a 180° viewing angle. Signal processing includes exponential average and weighting coefficient 2 to 2^{16}. Cost is under £1500. Hitachi Denshi (UK) Ltd. Tel., 0181 202 4311; fax, 0181 202 2451.

Literature

IGBTs. Toshiba has a short brochure describing the company's family of insulated-gate bipolar transistors, which cover the 600V-1700V range. Toshiba Electronics (UK) Ltd. Tel., 01276 694600; fax, 01276 691583.

Data on CD-rom. *Hitachi's* documentation is going onto cd-rom, that available so far being an optoelectronics guide and a memory data book. A parametric search facility enables a user to find a device by entering requirements, whereupon relevant data sheets are put up on screen. Hitachi expects more frequent updates than has been possible with data books. Hitachi Europe Ltd. Tel., 01628 585000; fax, 01628 585200.

Navigation systems

GPS receiver. For fleet management, navigation or search and rescue, the *VP Oncore* global positioning receiver provides six parallel GPS channels and has an accuracy of less than 25m or 1-5m with the differential facility, with dynamic capability of over 515m/s at altitudes up to 60,000ft. Time to first fix is 20s typical, and features include one pulse/s timing, a ttl serial interface, the American National Marine Electronics Association interface, Loran emulation and 100-waypoint navigation. Power consumption from 5V DC is 1.1W. Macro Group. Tel., 01628 604383; fax, 01628 666873/668071.

Load-cell controller. Indicators/controllers for load cells by Control Transducers interface directly between load cells and processing pc. *Model L1* is a 3.5-digit red led digital indicator for the normal Wheatstone bridge, three-wire amplified sensor and current-loop transmitter, providing readout in engineering units. The instrument is in a standard $^1/_8$ DIN case. Control Transducers. Tel., 01234 217704; fax, 01234 217083.

Power supplies

Universal 20W supply. Gresham's *GEU 200* series of 20W encapsulated dc power supplies accepts inputs of 84-264V ac without ranging or links to make usable worldwide. It is designed to take mechanical shock, vibration and pollutants and has an i/o isolation of 3750V rms. Outputs are from 5V at 4A to 15V at 1.5A and there are dual ±5V, ±12V and ±15V versions. Protection is provided. Gresham Power Electronics Ltd. Tel., 01722 413060; fax, 01722 413034.

Twin Pentium PSU. Intended to power dual-Pentium systems, Semtech's *MP7308-VR* is a 93% efficient, 35W dc-to-dc converter operating over the 4.5-5.5V dc input range and housed in a board-mounted case measuring 2.6 by 0.6 by 0.5in. It is voltage and current protected and standby current is 30mA. Semtech Ltd. Tel., 01592 773520; fax, 01592 774781.

2W dc-to-dc converters. XP's *NMH* series 2W dc-to-dc converters are available in outputs of ±5V, ±9V, ±12V and ±15V, running from supplies of 3-48V dc. Total power can be taken from one pin to form a single-output device. Packaging is sip, dip or surface-mounting and i/o isolation is 1000V dc. XP plc. Tel., 01734 845515; fax, 01734 843423.

Lead-acid charger. *TRUEcharge 20i* by Statpower is microprocessor-controlled to provide three sequential charging phases: bulk, to give the first 80% of capacity; absorption, to condition the battery; and a float charge to give maximum output. It automatically compensates for loads imposed by equipment during charging and, every 21 days, partially discharges the batteries and replenishes them for conditioning. An lcd shows charging state and conditions such as overheating, low power supply and shorts. Models are offered to take flooded or gel lead-acid batteries from 100Ah to 400Ah in 12V or 24V versions. Merlin Equipment. Tel., 01491 824333; fax, 01491 824466.

File server UPS. Vero offers the *Share-UPS*, a free-standing or rack-mounted UPS controller for up to eight file servers; two connected together handle up to 15. On mains failure, the UPS sends messages to each server by hard-wired serial connection, instructing each to start a graceful shutdown and maintaining the power supply until the server has shut down. The UPS then goes into battery conservation mode and initiates re-boot of all servers when mains is restored. There is a modem link for remote diagnostics and re-boot of locked-up systems. Vero Electronics Ltd. Tel., 01703 266300; fax, 01703 265126.

UPS for InterNet. *BBD450VA ByteBak* is an uninterruptible power supply for InterNet-ready file servers and Pentium Pcs. At an alarm condition, it dials up to four telephone numbers, reports status and takes calls back from engineers to provide remote diagnosis. Features include phased start-up to power a master socket before slaves; selected load shedding to turn slaves off first; automatic self-test, including site wiring; and a front-panel lcd for input, output and battery data. Galatrek International Ltd. Tel., 01492 640311; fax, 01492 641828.

60W dc-to-dc converters. *PKG* converters from Ericsson for use in convection-cooled systems need no extra cooling hardware. They measure 74.7mm by 61.3mm by 15mm, provide 86% efficiency for the 5V, 60W type (82% for 3.3V, 45W) and 1.5kV i/o isolation. Ericsson Components AB. Tel. (Sweden), 0046-872 17045; fax, 0046-872 17001.

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2kW modular power. Accepting a universal ac input with power-factor correction, XP's *VS* series of psus consist of a standard PFC input module delivering a 385V dc bus to up to four output modules: each can be a single-output type or a multiple-output model providing three outputs, at a total output power of 1500W or 2000W. Benefits of this system include the avoidance of special transformers, one reason for the rapid despatch - three days to configure a supply to order. Size is also reduced to 5 by 5 by 11in for the 1500W unit and 5 by 8in for the 2kW package. All outputs are adjustable and current-shared and there is full protection and indication. EMI filtering to international standards is provided. XP plc. Tel., 01734 845515; fax, 01734 843423.

Switches and relays

Modular switch system. EAO *Series 61* is a universal system of push-button switches, indicators and keylocks designed to switch up to 250V at 6A. Various bezel designs, lens colours and switching elements can be combined, snap-action and slow-make switches being available with solder, plug-in, pcb or screw contacts. Each element can have up to three contacts. Panel cutout is 16mm in diameter, but the units can also be flush-mounted in a standard 22.5mm cutout. EAO-Highland Electronics Ltd. Tel., 01444 236000; fax, 01444 236641.

Custom switch shafts. Grayhill rotary switches now have a range of variations in bush size and shaft length for thicker panels or staggered mounting. Shafts have various flats for special knobs or can be round for collet fittings. EAO-Highland Electronics Ltd. Tel., 01444 236000; fax, 01444 236641.

Absolute encoders. *Absolute Modular Encoders* from Control Transducers are very low-cost devices to mount directly onto a shaft to provide position feedback or control. They are single-turn, non-contact, optical devices giving a 360° range and, being absolute, need no reset or homing cycle. Model *AA* provides a 0-4096mV analogue output, while model *AD* gives a serial encoder interface output to RS-232, both to 12-bit accuracy and with a single supply. Both models can be supplied with incremental encoders and as modules for mounting to existing shafts. A full range of accessories is on offer. Encoders are able to handle up to ±0.01in axial play, and wiring is via a telephone-type plug. Control Transducers. Tel., 01234 217704; fax, 01234 217083.

Rotary-coded dip switches. Low-profile, rotary dip switches by EECO in the *Micro-Dip 4500/4700* through-hole (4500) and surface-mount series stand only 2.5mm from the board surface, minimising shielding of other components and allowing heat to reach them in the assembly process. They are available with 10 and 16 positions in a variety of codes, top adjusted and with an extended shaft. Transico Inc., EECO Division. Tel., 01954 781818; fax 01954 789305.

Keylock switches. *EAO-Swisstac* keylock switches come in a choice of configurations, including versions with two or three positions with various switching arrangements and angles, normally open or closed and behind-panel depth of 45mm or 70mm. Bezel sizes are 18mm and 24mm round or square and 18mm by 24mm. All are sealed to IP65 and are fitted with terminals for pcbs or standard. EAO-Highland Electronics Ltd. Tel., 01444 236000; fax, 01444 236641.

No-nuisance trip switch. Mechanical Products *MP 2200* front-panel rocker switch will not trip at up to 35 times rated current unless the overload lasts for more than half a cycle of 50 or 60Hz to provide improved inrush capacity and protection against nuisance tripping. A double contact break gives better short-circuit protection. Current ratings from 5A to 25A are offered, as is a trip-free mode of operation. Radiatron Components Ltd. Tel., 01784 439393; fax, 01784 477333.

1kV relay. Although only 15 by 7.4 by 8.2mm, *Matsushita's TX* relay possesses a 1kV rms withstand voltage between open contacts and 2.5kV rms between coil and contact. There are two changeover contacts, capable of switching from dry to 60W, 2A, 240V; coil voltages are from 1.5V dc to 48V dc and pickup/dropout times are 2/1ms. Monostable or latching versions are available. Matsushita Automation Controls Ltd. Tel., 01908 231555; fax, 01908 231599.

Transducers and sensors

Linear Hall sensors. *A3506/7* sensors by Allegro are sensitive, linear Hall-effect devices producing an output proportional to the incident magnetic field. Both ICs contain a quadratic Hall element, temperature compensation, signal amplification and a ratiometric, rail-to-rail output amplifier to interface directly with a-to-d converters or microprocessors. Sensitivity is 2.5mV/Gs, at operating temperatures up to 150°C. Allegro MicroSystems Inc. Tel., 01932 253355; fax, 01932 246622.

Accelerometers. Surface-mounted piezo-resistive accelerometers by IC Sensors are only 3.4mm square and weigh 300mg. *Model 3031* is a machined silicon mass suspended by multiple beams from a silicon frame, the active elements being piezoresistors in the beams. Silicon caps provide over-range stops and high shock resistance and durability. Ranges are available from ±2g to ±500g and, depending on the range, frequency response is from 0-400Hz to 0-2500Hz. Sensitivity varies from 2.9mV/g to 0.1mV/g, depending on range. Eurosensor. Tel., 0171 405 6060; fax, 0171 405 2040.

VMEbus PCMCIA carrier. Arcom has a two-slot PCMCIA carrier board for Type I, II or III cards, with data loggers and instrumentation in mind. It is backed by PhoenixCARD Manager Plus software running on a range of PC-compatible 486 and 386 processors under dos or Windows. The *VPCMCIA* 6U board slots take any type of PCMCIA card and other features include a 128K eprom socket for local configuration data storage, power distribution for 5V and 3.3V cards and esd protection. Arcom Control Systems Ltd. Tel., 01223 411200; fax, 01223 410457.

High-temperature probe. Fluke's *80PK-7* temperature surface probe has a range of –320°F to 1120°F (–196 to 600°C), using a ribbon sensor that plugs into any temperature meter with a type K connector. It is made of 303 stainless steel and high-impact PVC, being equipped with a four-foot cable. Fluke UK Ltd. Tel., 01923 240511; fax, 01923 225067.

Custom speakers. *Vitavox* has a new design and manufacture service for custom-made loudspeakers for low or high-volume production. The non-standard, low-profile designs can be made with outputs in the 1-10W range at 3-100Ω impedances, with standard or tropical paper or plastic cones in circular or elliptical form. Magnets are ceramic or Ticonal and frequency response is typically 100Hz-12kHz. Vitavox Division, Secomak Ltd. Tel., 0181 952 5566; fax, 0181 952 6983.

Vision systems

Video comms camera. Sony's *CCM-PC5P* camera is meant for desktop video conferencing and incorporates a lens and capacitor microphone, the whole being able to sit on top of a computer monitor. The $^1/_3$in ccd camera resolves 330 television lines at a minimum sensitivity of 20lux at f2.8. Power needed is 7-13V dc; a privacy shutter is directly connected to the power supply switch. Model *CCM-PC5P* is for use with PAL systems; *CCM-PC5* with NTSC. Sony Computer Peripherals & Components. Tel., 01932 816000; fax, 01932 817001

COMPUTER

Computer board-level products

Pentium board. Concurrent Technologies's *CP SBC/PEx* Multibus II cpu uses a 100MHz Pentium processor and provides several high-performance interfaces. The board is a single-slot upgrade for 486 iRMX and UNIX users. Processor and memory subsystem has its own 64-bit bus and is connected to a SCSI interface by the PCI, used as an on-board local bus, being connected to FastI/O bus and Multibus II interface via a bridge. There are also Ethernet, iSBX and serial/parallel i/o interfaces. Concurrent Technologies Tel. 01206 752626; fax 01206 751116.

Computer systems

Embedded PC. AMC has introduced the *2001*, an embedded pc on a half-sized cpu card, with expansion through a PC-104 interface or through the ISA bus and using processors varying from the 386SX to the 486SLC2. Features include a flat-panel display interface, SCSI II, PC-104 interface, optional RS-485 and optional bootable 3Mb flash drive. It can be used alone or in a passive backplane system. Advanced Modular Computers Ltd. Tel., 01753 580660; fax, 01753 580653.

Data communications

PCMCIA serial i/o. In PCMCIA format, a serial i/o card for use in sub-notebooks or, indeed, any computer equipped with a Type II slot, is introduced by Smart. It provides a high-speed serial port and is compatible with dos and Windows-based software and connects to standard serial peripherals. A 16550 uart and 8350 usart provide either asynchronous or synchronous communication; the asynchronous port can be configured as COM 1, 2, 3 or 4 and the synchronous port is programmable – both without jumpers. Power-down and sleep modes are available to conserve battery life. Smart Modular Technologies. Tel., 01604 497735; fax, 01604 497739

Low-cost modem. AT&T's *Controllerless Modem Chip Set* uses a pc host to perform AT commands, data compression and error correction to eliminate a separate controller chip and associated ram and rom and allow the construction of low-cost V.32bis modems. It supports data rates and fax speeds up to 19,200b/s, the AT command set, V.42 and MNP*4 error correction and V.42bis and MNP5 data compression. The chip set is designed to run under Windows. AT&T Microelectronics. Tel., 01734 324299; fax, 01734 328148.

Development and evaluation

Motorola micro emulators. From Hewlett-Packard, a series of in-circuit emulators for Motorola 68331, 68332, 68F333 or 68336 microcontrollers. Supporting all standard features, they also have up to 8Mb of dual-ported emulation memory for modification and display of C debug functions, improved memory-map resolution of 256-byte boundaries, background or foreground monitors, emulation of 8-bit and 16-bit target memories and support for 20MHz or faster chips. Other features include the support of clock speeds to 33MHz with no wait states out of target memory and 25MHz with no wait states out of emulation memory for debugging at full processor speed, and automatic creation of initialisation code. Hewlett-Packard Ltd. Tel., 01344 366666; fax, 01344 362269.

Software

HTBasic for Windows. Workstation Source has a version of *HTBasic*, Hewlett Packard's scientific and engineering programming language,to run under Windows Chicago, NT, 3.1 or 3.11. It is completely in 32-bit code and supports high-level matrix, complex and scalar maths, an HP-type screen, plotter and printer graphics, SCPI compatibility and HP-GL plotter output. IEEE-488 cards are supported, as are data acquisition and RS-232 instrument control statements. It is file compatible with dos, HP700 series and Sun SPARCstation. Workstation Source Ltd. Tel., 01734 759292; fax, 01734 757522.

68K Modula-2. 68000-series target systems can now benefit from the programming language *Modula-2*, which is recommended for use in applications where safety is an important feature. It is highly structured and its compiler does not allow erroneous constructs to pass, unlike some other languages; it is also cheaper and easier to use than Ada. The package is Windows-based and the system integrates debuggers, simulators and emulators. Pentica Systems Ltd. Tel., 01734 792101; fax, 01734 774081. ■

...continued from page 216

just v_1. But until line 160 has been calculated, the program cannot know the value of v_2.

The way round this difficulty is to use pairs of periods Δt, during the first of which v_1 is held constant giving a new value i_2 for i, and then during the second this new value of i is held constant, giving a new value v_2 for v. The only modification required to the program is that line 60 becomes $v_2=v_1-i_2t$. With this correction, peak values of voltage and current remain constant, as would be expected for a tuned circuit with no damping, even if the time step is made much longer than the 50ms defined in line 127.

Amplitude remains constant even if Δt is made as large as, say 800ms, a significant fraction of the 6.28 second period of the sinewave across the tuned circuit. This is shown in **Fig. 3**, where you can see that there are $2\pi/0.8$ or nearly eight segments per cycle; that the theory of finite differences really works never ceases to amaze.

Using the present value of voltage to predict the new value of current, and then using the new value of current to predict the new value of voltage is a standard way of generating quadrature sinewaves such as those shown in the figures. It can be done to any desired degree of accuracy in dsp[1], and it can equally well be done analogwise in hardware. In this case, the result is likely to be a stepwise approximation to a sinewave, as in an article which appeared in this journal some years ago[2], rather than the linear interpolation representation shown in Figs. 2 and 3.

Figure 4, reproduced from reference 2, shows the sinewave generated at 500Hz, and Lissajous figures showing the very close approximation to quadrature achieved between the sine and cosine outputs both at 5Hz and 5kHz. The only essential difference between the circuit used in reference 2 and the program of Listing 1 is that instead of using a transient as the excitation, the loop gain was made slightly greater than unity so that the amplitude built up to the point where some diode gain reducing networks came into play.

Returning to our starting point, the program in Listing 1 is easily modified to include a finite loss resistance r in series with the inductor. Following t=0, as the current increases, the volt drop across the resistor must be deducted from the capacitor voltage to find the voltage impressed across the inductor. The reduced *volt.second* product impressed on the inductor means the peak current will be reduced. This applies twice each cycle as the current flows first one way and then the other, the waveform dying away exponentially. And instead of the delta function generator of Fig. 1a, one could connect a 1V step function generator between the lower plate of the capacitor and ground in Fig. 1b. Inductor voltage and current waveforms would be unchanged. But after the waveform had died away, there would remain an amount of energy $CV^2/2$=1 Joule, in this case, stored in the capacitor. ■

Voltage-current relationship

In an ac circuit, voltage v and current i are related as follows:

for an inductor, L henries,

$$v = L\frac{di}{dt}$$

$$\underbrace{\int_{t_1}^{t_2} v\,dt}_{\text{volt.second product}} = L\underbrace{\Delta i}_{\text{change in current over interval } t_1 \rightarrow t_2}$$

for a capacitor, C, farads,

$$i = C\frac{dv}{dt}$$

$$\underbrace{\int_{t_1}^{t_2} i\,dt}_{\text{charge}} = C\Delta v = \Delta Q \text{ Coulombs}$$

References

1. Accurate sinewave oscillator, N. Darwood, *Wireless World*, June 1981
2. Accurate sinewaves, D H Follett, *Wireless World*, Nov. 1981

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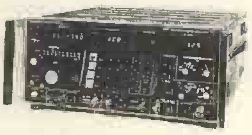

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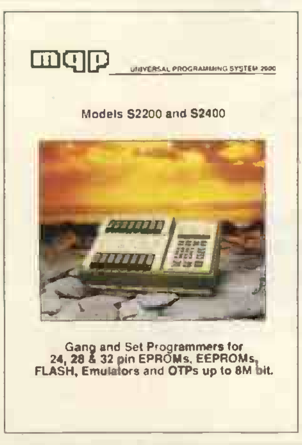

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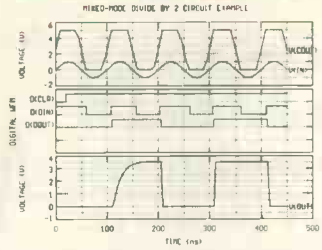

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PROGRAM 8 CHIPS IN THE TIME IT TAKES FOR ONE!

At £645 costing around half the price of slower gang programmers, the Speedmaster 8000 gang programmer uses a simple 2 button operation in stand-alone mode. PC operation gives comprehensive file handling and editing functions. Capable of gang and set programming it supports 32 pin EPROMs to 8M with no adaptors required. Programming cycle times of only 23 seconds for 8 27C010's mean your throughput can now be faster than ever before.

ROM/RAM EMULATOR PLUG IN CARDS

Using these expansion cards your programmer can run as if there's an EPROM or RAM plugged into the target socket. Available as 8 bit wide 128k x 8 as standard, upgradable to 512k x 8, and 16 bit capable of emulating 40 pin EPROMs. They can emulate both 5V and 3.3V devices.

PACKAGE ADAPTORS

A full range of package adaptors is available for non DIL devices and parts with more than 40 pins. Prices from £65.

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FREE SOFTWARE UPGRADES! - KEEP UP TO DATE WITH NEW DEVICES

Before you choose your programmer, check out the cost of ownership. While other manufacturers charge for every update or require expensive libraries and modules, ICE Technology programmers support the whole range of devices at no extra charge*. And keeping up to date is FREE for life at no charge on our BBS service.
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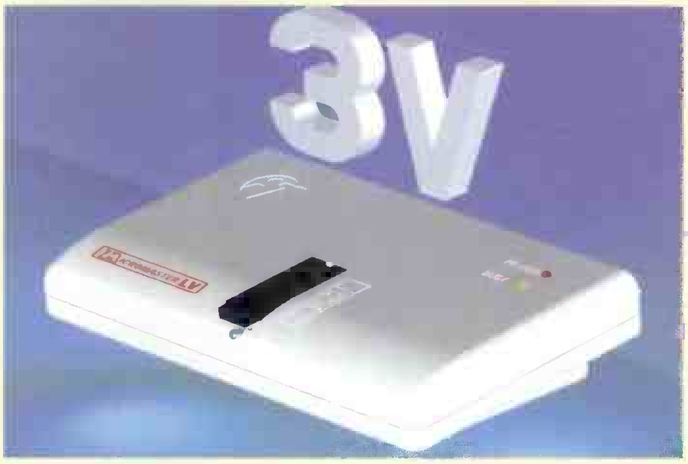

The latest universal programmers from ICE Technology, the Micromaster LV and Speedmaster LV, now support programming *and* verification of 3.3V devices, now you can test devices at their actual operating voltage.

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- Widest ever device support including: EPROMs, EEPROMs, Flash, SPROMs, BPROMS, PALs, MACH, MAX, MAPL, PEELs, EPLDs Microcontrollers, etc.
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CIRCLE NO. 101 ON REPLY CARD

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8 bit Emulator card

Expansion card containing 8 bit wide ROM RAM emulator, includes cable and software. 128K x 8.

£125

16 bit Emulator card

Expansion card containing 16 bit wide ROM RAM emulator, includes cable and software. 128K x 16.

£195

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