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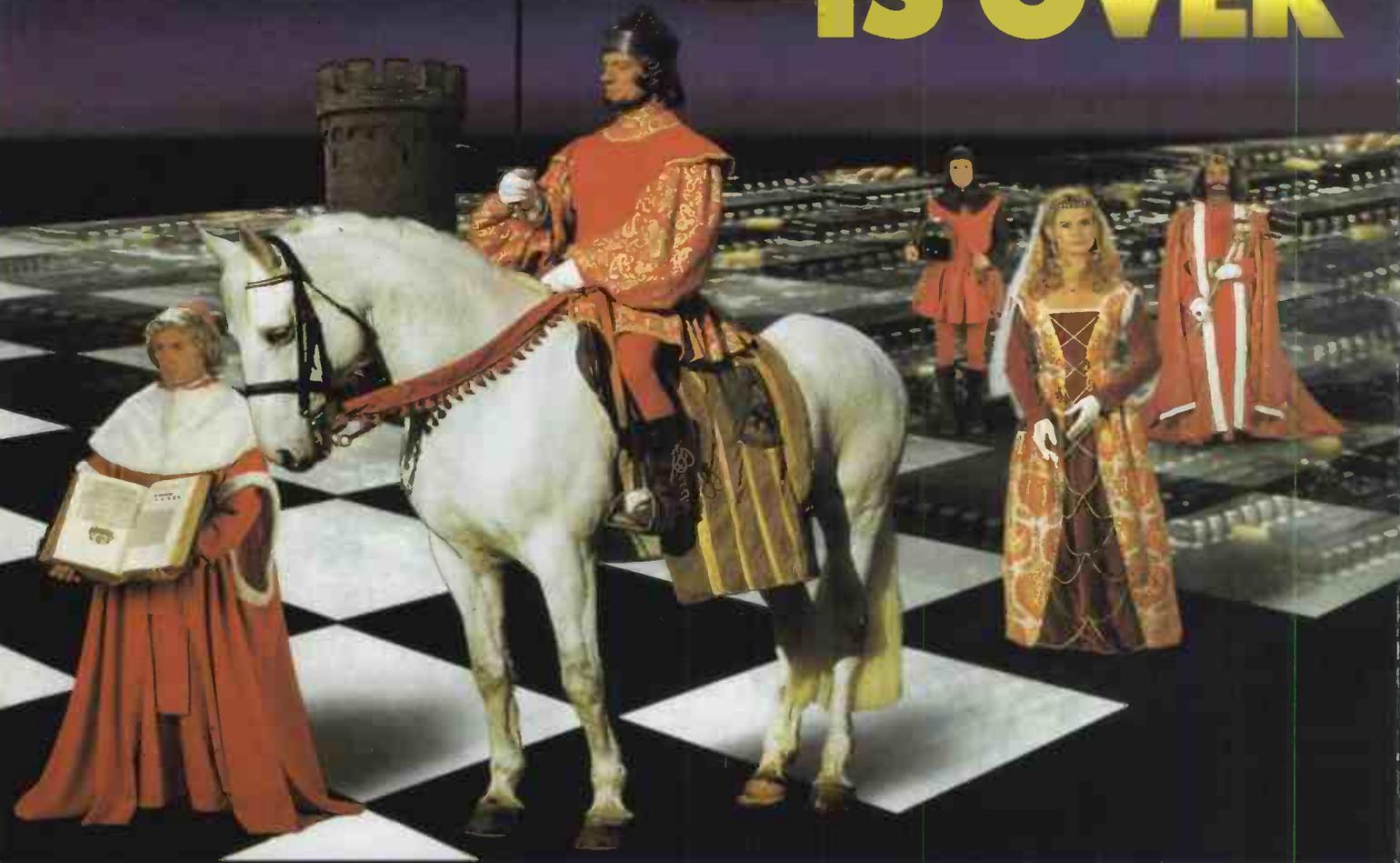


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Cover – Hashim Akib



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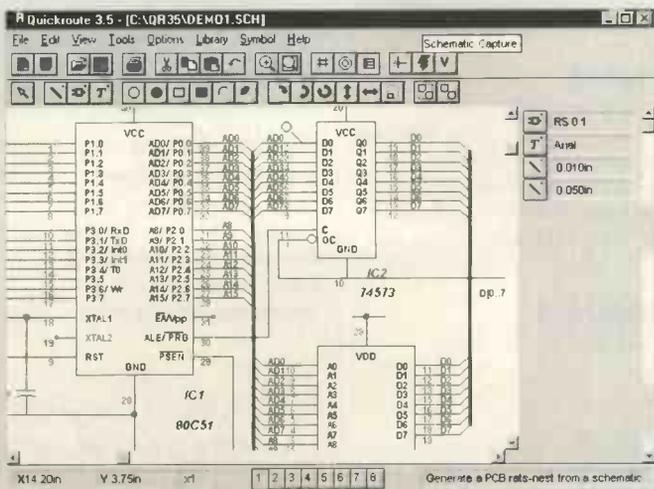
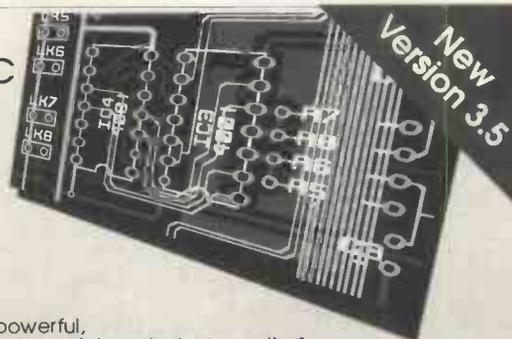
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A computerised confidence trick

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Although I'm not particularly ancient I can remember quite clearly what one kilobit of program ram and a large helping of ingenuity could produce from the original Sinclair ZX81 computer. It provided an object lesson in efficient programming. Working with such limited resources required absolute focus on the intended application for the program.

I also remember equally clearly that, as a former components editor on our sister publication *Electronics Weekly*, I wrote with some awe about the power of the MIPS R3000 risc architecture. This cache equipped microprocessor had been designed to run multitasking Unix at the heart of a high end graphics workstation.

Much has happened over the last 10 to 15 years, six of which I spent as editor of this magazine. But I am not sure that the progress I see around me is always in a forward direction. Having marvelled at the latest generation of games hardware and its astonishing polygon animation, it should have been no surprise to learn that the particular one that I was looking at used as its main processor the same workstation architecture in its R4000 incarnation. There is no problem here. The games machine companies always said that they would eventually lead the way in the use of raw processing power and, being totally focussed on the trivial, we pundits never really believed them or paid much attention.

It is amazing what a single-minded application feeding into a mass market can achieve. Comparing the performance of Sony's Playstation with that early ZX81, it is easy to see what the extra megaflops and megabits have achieved.

Not so with the Intel/Microsoft

personal computer market. The combine regularly doubles the power of its offerings but without a commensurate doubling of utility. Microsoft continues to turn out sloppily programmed, convoluted and memory intensive software. Intel matches this with increasingly powerful general purpose microprocessors to make the inefficient software architecture run at halfway decent speed.

It is not even as though we need the dubious sophistication of the duopoly's latest offerings. The majority of business applications should require relatively modest memory and processor requirements; spreadsheets, wordprocessors, presentation graphics and networking involve relatively little data computation and manipulation. It shouldn't take 16Mbits of ram, 133MHz processors and a Gbyte of hard disk to run the things that you or I want to do – unless we are working in professional graphics.

Yet the Intel/Microsoft axis has pumped so much marketing money into persuading us all that computer gear more than a couple of years old is fit for nothing but scrap that it almost amounts to a confidence trick. It causes people and businesses to spend needlessly on upgrades while at the same time creating chronic inefficiency by forever trapping users on a perpetual learning curve.

Even the computer manufacturers are not immune from the effects of the duopoly's cavalier marketing. One household name PC maker found that it was unable to pay for the massive shipment of microprocessors and other chips which it had bought from Intel.

Being loaded with your money and mine, Intel simply turned the debt into a loan. The reason that the bill went unpaid in the first place was that the PC maker had gambled

that the hot PCs for the season would be 75MHz Pentiums and geared up production accordingly. In the event, duopoly marketing ensured that the punters were demanding 100MHz and 133MHz machines. The loan allowed the PC maker to unload the 75MHz machines at bargain prices, causing all sorts of people to lose out in the process.

And then there is the Internet. While much of the purpose which is supposed to drive this technological hula-hoop verges on insanity, uses will emerge which make sense – and virtually none will require intensive local processing. Thus



Playstation screen photo.

Games machines have advanced phenomenally – but have pcs made the same progress?

may be born the PC Killer, an Internet access appliance which allows users to do all the genuinely useful things of which the net is supposed to be capable, but without having to invest thousands of pounds in the latest Intel/Microsoft confidence crisis. A simple computer, fitted out with just enough memory and processing power to make sense of the Internet, may induce a change of thinking.

It is about time that we ignored the advertising and started to reconsider what we actually do with the machines which we purchase.

Frank Ogden

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Government rejects microelectronics advice

The recommendations of the Government's high technology advisory panel highlighting the need for a microelectronics R&D centre have fallen on deaf ears.

The first 63 projects bidding for funding under the Technology Foresight initiative were named last week and the proposal to set up a university-based microelectronics centre providing R&D and training support was not among them.

Although the strong emphasis on multimedia and broadcast technologies in the IT projects is welcome, there is some surprise at the omission of the microelectronics bid.

According to Neil Downie at semiconductor manufacturing

equipment and materials trade association JEMI (UK) there is a need for technology development and training support and companies were prepared to fund it. "Inward investors don't do R&D here and we need to encourage them," said Downie.

One of the Government's IT advisors on the Foresight programme Ian Barron, founder of Inmos, said there was a need for a microelectronics R&D house to support the high levels of inward investment by semiconductor manufacturers. "We need to maintain a source of expertise for inward investors to call on and use," said Barron.

The shortlisted projects were selected from over 500 involving

collaborations between industry and university research groups. The average size of the projects is £4m, ranging from £500,000 funding for the smallest up to £15m for the largest. The winning groups will be named at the end of May.

There were just six projects named under the heading of Information Technology, Electronics and Communications (ITEC), they included the setting up of four UK centres of excellence in mobile communications, microsystems at the Rutherford Laboratory, broadcast and data mining at the University of Ulster. There were also projects looking at multimedia at Lancaster University and display technology at Thom EMI's CRL.

Road toll trials: "No problem"

Department of Transport (DoT) has denied its motorway tolling trials, due to start in July, are in jeopardy following the decision of a second consortium to withdraw last week. The Tollway group, led by US firm Amtech and including Serco and WS Atkins, has pulled out just one week after Siemens announced its withdrawal citing lack of government commitment to a commercial system.

But the DoT says it expects a preliminary shake-out. "We are basically in discussion with many companies which want to participate in the trials and these discussions will go on for another six weeks. We'll know which companies will definitely participate in the trials in six weeks time," said a DoT spokeswoman.

However, the departures of

Siemens and Tollway could be followed by others if the UK trials do not begin on time. A number of the six remaining consortia prefer involvement in trials for a South Korean system, scheduled to begin this Spring.

But Tony Kellett, the technical director of Peek, the company leading the Tollstar consortium, believes competition will remain intense.

Small format memory-card imminent

A miniature memory card a quarter the size of the PC Card (PCMCIA) format will appear in the summer as part of an initiative by electronics firms in Europe, the US and Japan.



Small format memory cards, due for summer appearance, will be a quarter the size of their PCMCIA predecessors.

The standard format, dubbed the Miniature Card, is expected to be small enough for use in cameras and audio equipment as well as notebook PCs. Details of the format, including interface specifications, will be released during February, with first products using the card expected in July.

Bill Howe, general manager for memory components at Intel, a leading member of the Miniature Card Implementers Forum, said: "We believe flash memory is the ideal storage medium for hand-held consumer devices and for bringing that data back to the PC for further use. We have several design wins for Miniature Card in hand and will ship product in the second quarter of this year".

Internal memory of up to 64Mbytes of either flash, DRAM or ROM will enable the cards to store image, text and voice data. Data transfer to the PC will probably be via a standard PCMCIA adaptor. The

ruggedised package with a pinless connector, at 38 x 33 x 3.5mm, is a quarter the size of PCMCIA cards.

The small format memory cards are expected to have applications in digital cameras, audio recorders, mobile phones and other consumer equipment.

Companies supplying the core semiconductor technology include Advanced Micro Devices, Fujitsu, Intel and Sharp. End user system support includes Compaq Computer Corporation, Hewlett Packard, Konica Corporation, Nokia Mobile Phones and Olympus Optical Company.

The Miniature Cards will compete against Compact Flash from SanDisk and the as yet unannounced Solid State Floppy Disk Card from Toshiba. Compact Flash however only offers flash memory and uses connector pins less suited for consumer systems.

Richard Ball, *Electronics Weekly*

Circuit board designers unprepared for EMC

PCB designers have a lack of knowledge concerning the new EMC Directive, according to a survey by Zuken-Redac, the pcb and multi-chip module CAD/CAM software company.

The survey of 135 UK printed circuit board designers showed that 94 per cent were unprepared for the EMC Directive even though 84 per cent were fully aware of the new regulations.

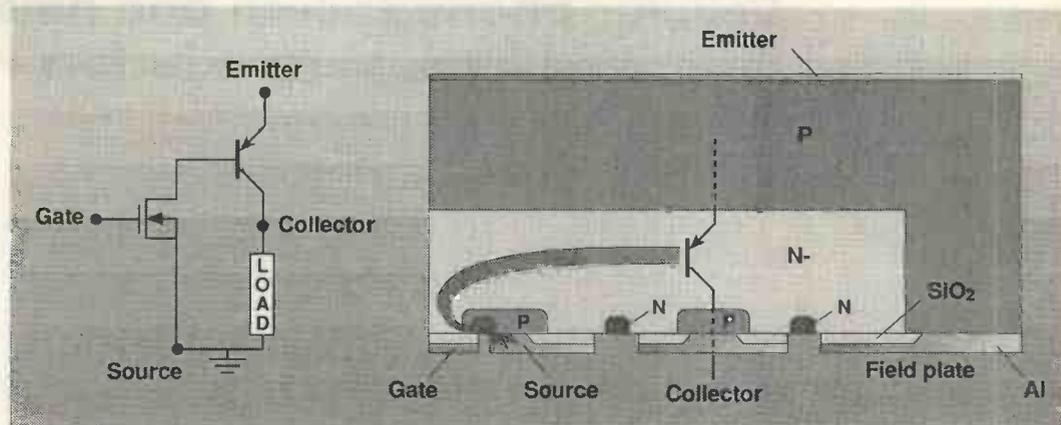
Zuken-Redac said most companies had not taken the opportunity to improve their designs with respect to emc in advance of January 1996, the starting date for complying with the EMC Directive.

Another worrying point is that 80 per cent of all pcb designs are carried out on pcs. However, there are few pc-based pcb CAD design tools with integrated emc design rules. Suppliers such as Zuken-Redac, VeriBest and Mentor Graphics are only now introducing the required tools.

Even though pcb designers do not necessarily have to make their products comply, they may be supplying to original equipment manufacturers, designing without emc in mind during the early stages can lead to rework and time-to-market delays later. In most cases this will cost a significant amount of time and hence money until emc design at the pcb level is more fully appreciated. Layout designers with no emc knowledge have no way of knowing whether their design is helping or hindering the final system in terms of emc.

Currently available design tools for emc can be roughly classified into predictive and analysis types. Tools using predictive and analysis techniques, such as Ultimate's EMC Expert or Zuken-Redac's EMC Adviser, now released for the pc, compare the design with a set of standard emc rules giving a qualitative rather than quantitative result, at the expense of some accuracy.

Modelling the emc characteristics of individual components on the pcb, such as carried out with Mentor's Quad Quiet, has the potential to give the most accurate results, but this depends on the accurate modelling of each device. The disadvantage here is the heavy processing overhead and only the fastest pcs can handle the work load. It may be that accurate emc analysis cannot be carried out on pc based systems.



New combined bipolar/fet switching device

Zetex has produced a novel device for high side switching from low side control signals.

The ZHD100 'treeswitch' was originally designed for a specific automotive application but is now being offered for general sale.

The equivalent circuit is a p-n-p transistor with its base connected to the drain of an n-channel mosfet. This is very similar to an igbt, where the collector and source are also connected. Dave Casey, co-developer of the device said: "There is a parasitic igbt in the structure which we have suppressed."

The advantage of separating the collector and source is that the low saturation voltage of the transistor can be fully exploited. In saturation, the igbt becomes a diode with a voltage drop of about a volt. The source of the treeswitch can be taken more negative than the collector which allows the transistor to fully saturate down to a few millivolts.

The p-n-p transistor is made by diffusing n, then p structures into a p-type substrate.

The fet is diffused as a p-well, similar to the emitter, with an n-type island in it. The gate metallisation for the fet overlaps the n-island and the

main base n-type diffusion. It is insulated from the substrate by an oxide layer.

Pulling the gate positive with respect to the source attracts electrons into the p-type material below it which forms an n-channel between the base and source. This allows base current to flow and turns the transistor on.

The transistor has Zetex's matrix architecture which results in a low saturation voltage. The fet sits in the middle of it and is small in comparison, limiting the amount of current it can pass. This makes the fet approximate a constant current device, removing the need for a drain-base current limiting resistor.

The ZHD100 is rated to switch half an amp at 80V. At high voltages the device dissipation is predominantly from the 10mA base current. If the full output current of the device is not required this can be reduced by adding a series resistor between the source and ground.

There are other high side switches on the market. Casey said: "The matrix architecture results in a very small chip compared with its competitors. The small chip leads to a low device cost."

Similar to an igbt, the new high-side tree switch has separate source and collector terminals. Small relative to the bjt, the fet approximates a constant-current device, removing the need for a drain-base resistor.

Pentium P7 slips back

Sources close to Intel report that the forthcoming P7 microprocessor, the successor to the Pentium Pro, will be delayed by as much as year and will not involve Hewlett-Packard to the same extent as was previously believed.

The P7, now code named the Merced, is likely to be launched in 1998, not 1997 as has been indicated earlier by Intel. Also Hewlett-Packard's involvement in Merced now seems to be to define the 64-bit instruction set and software interfaces. Originally, Hewlett-Packard and Intel said they would both be

involved in the design of the chip.

The Merced will feature a new 64-bit instruction set but it will also run x86 instructions in native mode rather than relying on emulation technologies. Hewlett-Packard is believed to be working on its own microprocessor design to support the same microcode.

The P7 delay could be a deliberate Intel ploy to give it more time to establish Pentium Pro in the marketplace. There is less pressure on Intel to bring out new microprocessor architectures since its competition is struggling.

Cellular phone speech and noise are improved by superconductors

Superconducting technology is being used in the US to improve the noise performance and voice quality of cellular phones.

Ameritech Cellular Services reported excellent results from the use of a cellular system radio-frequency filter incorporating superconducting devices from Illinois Superconductor. Ameritech said that use of the filter at the cell site, improved voice signals by more

than a third and was 10,000 times more effective at eliminating interference and signal noise.

The SpectrumMaster filter also allowed more cellular phone users to access the system and increased the receive path range. There were also fewer dropped calls and an increase in the number of usable channels. Ameritech said that the filter is especially effective at cell sites that have problems from interference,

either from buildings or from other radio signals. If other cellular phone system providers adopt the filter, it could open a major new market for superconducting materials and help drive the development of other superconductor based technologies.

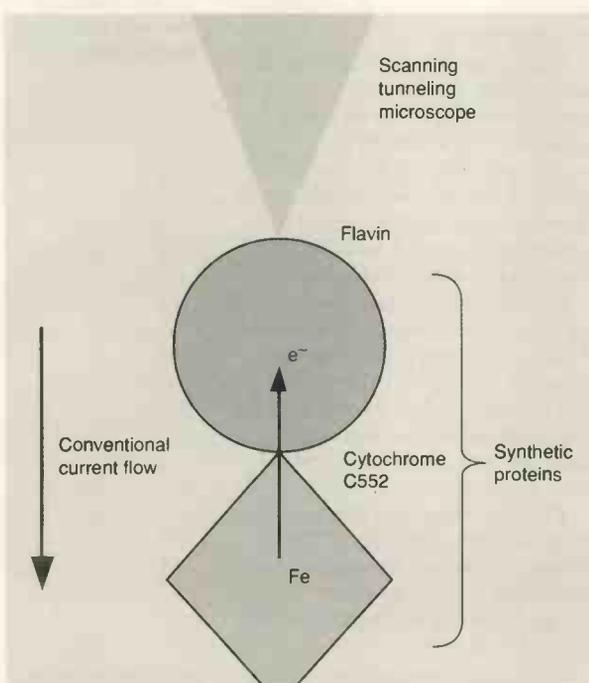
High temperature ceramic superconductors continue to operate above the boiling point of nitrogen (77K).

Poor demand for video games hits 3DO

Video games console designer 3DO, said that its third quarter results will be worse than expected due to poor demand for video games systems during the recent Christmas season. 3DO said that it

will be forced to establish reserves against large inventories and a five million dollar charge due to changes in its business model which includes a focus on Internet related products. The troubled company

continues to struggle to establish its video game hardware in an increasingly competitive market that includes Nintendo, Sega and newcomer Sony with its PlayStation.



Scientists in Japan have fabricated a diode only 2.5nm across using a synthetic protein molecule. The molecule has been made by Mitsubishi Electric and is a combination of the natural protein cytochrome C552 and the vitamin flavin. Under an electric field an iron atom in the cytochrome can be persuaded to release an electron to the flavin. Conversely, a similar electric field in the other direction cannot push an electron to the cytochrome. Forward voltage drop is reported to be 900mV at 70pA and 10,000 of these devices would fit onto a diode made with leading edge 0.25µm semiconductor technology. Like most of the current wave of molecular electronics, it is still in development and requires several hundred kilos of specialist equipment to make it work. To operate the molecule a scanning tunnelling microscope is used to select one of a number of the molecules that has been laid down on a metal substrate. The substrate is then the cathode and the microscope the anode.

Speech secure enough for banking?

Cambridge-based speech recognition technology firm Vocalis, is to develop a speaker verification system for secure banking over telephone lines as part of the EU-funded CAVE (CALLER VERIFICATION in banking and telecommunications) project.

The project's primary objective is to prove that speaker verification can be applied to financial transaction services, providing an initial level of protection.

"It is a pilot scheme acting as a demonstrator for the feasibility and

applicability of this technology. Banks have been very cagey and we have to prove that this technology is feasible," said Richard Winsky, senior research scientist at Vocalis.

CAVE is expected to last two years, after which the technology and the security provided by it will be upgraded. First field trials are expected within the year.

The actual deployment of speaker verification technology will await the outcome of the R&D work in CAVE. The resulting system is to complement other security mechanisms.

Modem makers boosted by Internet connections

Upward spiralling numbers of Internet connections in the US, with consumers linking their pcs to online services, is good news for modem suppliers.

More than 18.6 million modems were shipped last year, generating revenues of about \$5.8bn, exceeding earlier expectations, said US market research firm International Data Corporation (IDC). The market grew 82 percent compared with 1994 shipments of 10.2 million.

The increased shipments, however, came at the expense of falling prices for almost all modem speed categories except in sales of top speed V.34 (28.8kbit/s) equipped modems

which held their sales value. The sales boom was led by U.S. Robotics, GVC and Hayes Microcomputer Products. IDC foresees continuing boom times for the modem industry for at least the next few years as the popularity of the Internet and online services continues to attract millions of new users and as modem owners upgrade to faster modems.

● US Robotics, the Illinois-based modem maker is reported to be planning a move into the market for consumer telephone hand sets, to offset falling prices which is squeezing margins in its core modem business. ■

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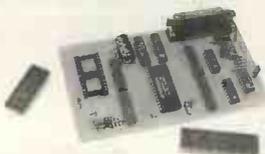


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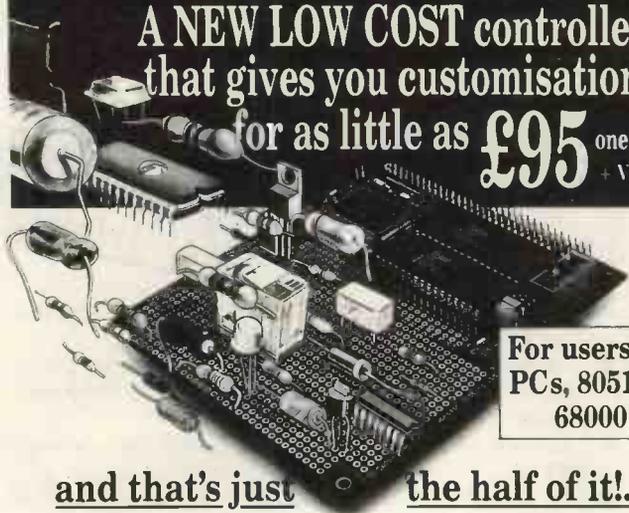
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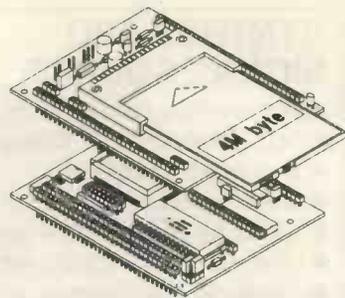
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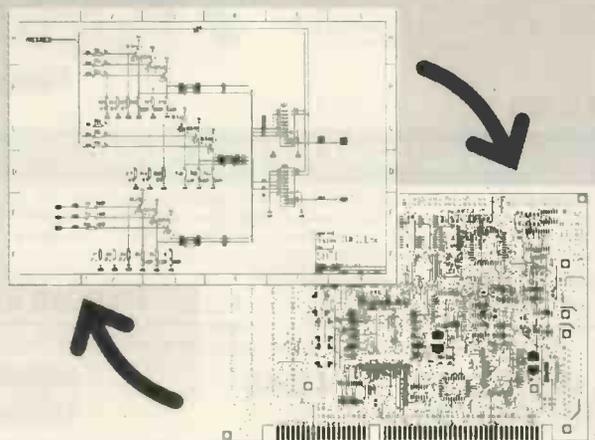
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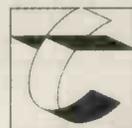
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Single electron switches presage quantum computing

Many obstacles have to be overcome before we can make practical use of single-electron devices, and fundamental research is feverishly being carried out to this end around the world. But the ultimate goal, of developing devices whose principles of operation are quantum mechanical – progressing electronics far beyond the scalability and performance limits of conventional circuits – means the field is hot one for science.

The work of one team in the US gives a good indication of where we are in development of real single electron devices – and also gives a glimpse of what could be achieved.

M G Ancona and colleagues at the Naval Research Laboratory in the US noted that comparatively little attention has so far been paid to the development of circuit architectures capable of exploiting single-electron device performance at circuit level. So he has devised a set of principles for single-electron digital circuits (sedc) and then used these rules to design some circuits ('Design of computationally useful single-electron digital circuits', *J App Phys*, Vol 79, No 1, pp. 526-538).

For computational use, circuits require non-linear rather than linear operation, which means mixing control and data circuits. Since the data now are single electrons, some portion of the control must also operate on this level. To achieve this Ancona has devised a single electron switch (ses).

Physical basis for the the ses is direct electron-electron repulsion. The switch is composed of two circuits, a switching circuit and control circuit, made up of a number of islands. The circuits interact through a capacitor, and the principle is that repulsion makes it energetically unfavourable for islands 3 and 6 (see figure) to be occupied.

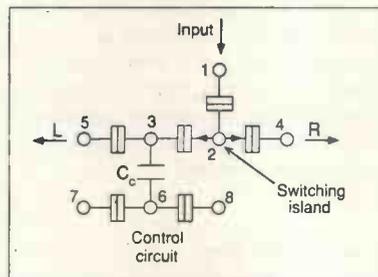
When the 'switching island', 2, is occupied by an electron, the gate biases must cause this electron to transition, in a fully Coulomb-

blockaded manner, to island 4 if island 6 is occupied, and to island 3 if island 6 is unoccupied. In this way, the electron in the switching circuit would be under the control of the electron in the control circuit.

Ancona says he has discovered a number of capacitance and bias values that allow this switch to operate, though have not discovered the optimum as yet.

Ancona also shows how the switches could be linked together to make And/Or and Xor circuits, and 1 bit memories.

M G Ancona is at the Naval Research Laboratory, Washington, DC 20375, email ancona@estdsun4.nrl.navy.mil



Electron-level switching. This SES comprises two circuit interacting through coupling capacitor C. Basis of operation is electron repulsion which inhibits the simultaneous occupancy of islands 3 and 6.

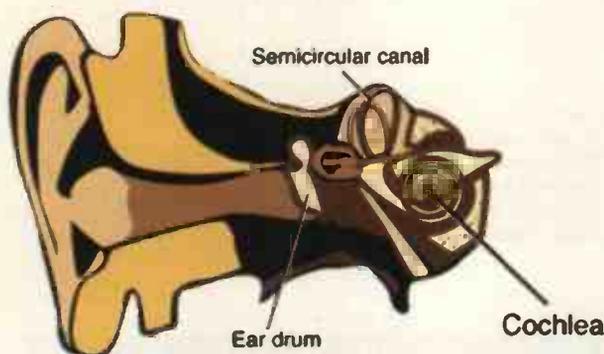


Sun set: We will not be around to see it, but what will our Sun look like at its death? Images produced by Nasa's Hubble Space Telescope of planetary nebula NGC 7027 show remarkable new details of the process by which a star dies. Features captured by Hubble include faint, blue, concentric shells surrounding the nebula; an extensive network of red dust clouds throughout the bright inner region; and the hot central white dwarf, visible as a white dot at the centre.

The nebula is a record of a star's final death throes. Initially the ejection of the star's outer layers, when it was at its red giant stage of evolution, occurred at a low rate and was spherical. The photo reveals that the initial ejections occurred episodically to produce the concentric shells.

Hearing in two different ways

George Zweig at Los Alamos National Laboratory has spent a lot of time looking into ears. As a result he has shown that how we hear loud sounds is quite different from how we hear soft ones. Now his theories of the mechanics of hearing are opening new directions in acoustic research that could lead to better hearing aids, improvements in the technology of cochlear implants and further development of speech-recognition machines.



Cross-section of the human ear shows the inner ear and the cochlea, where sensory hair cells respond to vibrations and send impulses to the brain. Travelling waves, amplified by the hair cells, can be reflected back and forth in the cochlea, causing the ear to ring or whistle. (Graphic by Edwin Vigil)

For loud sounds, the textbook understanding of hearing is essentially correct: sound waves enter the ear canal and vibrate the eardrum, whose oscillations are transmitted by the tiny bones of the middle ear to the inner ear, creating waves in the fluid-filled tubes of the cochlea. Sensory hair cells in the cochlea respond to motion of the fluid, generating electrical impulses that are interpreted by the brain as sensations of tone. Low-frequency tones excite hair cells further from the middle ear than do tones of higher frequency. Deafness is often associated with the destruction of hair cells.

For very quiet sounds, however, the ear is not just a passive receiver of sense impressions. Instead, the ear responds to and amplifies faint waves generated by soft noises.

Zweig, a physicist in Los Alamos' Theoretical Division, has recognised a symmetry governing behaviour of these waves in the cochlea and has developed an equation that describes what happens to these subtle travelling waves within the ear.

"Understanding how the ear functions has important implications for signal processing, for how you go about extracting information from many kinds of signals, not just speech and not just signals in the frequency range of hearing," says Zweig.

Back in the 70s his work led to the

discovery of the continuous wavelet transform, a way of displaying and extracting time and frequency information in a signal. Now, continuous wavelet transforms are used by other researchers in mathematics and engineering, with implications for a broad range of endeavours from music production to seismic testing to submarine surveillance.

Zweig hopes a clear picture of how the ear works also can help build better speech recognition systems. "When you look at how the ear responds to speech, what you see in it are the acoustic signatures or the resonant modes of the mouth, including how the vocal folds are moving in the throat. This is the kind of information you would need if you wanted to identify electronically whether one thing was being said rather than another, or who said it," says Zweig.

Zweig can also use his wave equation to explain common ringing in the ear. He shows that wave energy not transferred to the middle ear is reflected again and amplified again, combining with the original wave. This backward and forward wave amplification and reflection can set up a standing wave or resonance in the cochlea. If the process runs away with itself, the ear begins to whistle spontaneously.

Scientists rethink superconductivity

Could a new type of electron behaviour be responsible for 'high-temperature' superconductivity?

A team at IBM has conducted an experiment to suggest that this could be the case, opening the door to the production of new high temperature superconductors with predictable properties. Eventually, scientists hope they will be able to find superconductors that don't need to be cooled at all.

Superconductivity is believed to result from the ability of electrons to overcome their mutual repulsion and pair up in ways that enable them to pass unimpeded through the host material.

In recent years, two different types of electron behaviour have received significant support from theoretical physicists as possible keys to the mechanism of high-temperature superconductivity: 's-wave' electron pairing where charge-induced vibrations in the

material hold the pairs together, and 'd-wave' where the electron magnetic spins are critical.

Conventional low-temperature superconductivity is caused by a well-understood form of s-wave pairing. But some scientists have been sceptical that this should be the starting point for theoretical explanations of the higher-temperature phenomena.

Unfortunately, past experiments to distinguish between s-wave and d-wave have given mixed results. Now Chang Tsuei, John Kirtley and co-workers at IBM's TJ Watson Research Center and at State University of New York campuses in Buffalo and Stony Brook have come up with the most convincing evidence yet for d-wave electron pairing. The result is expected to be a shift in research that could spawn superconductors designed to exploit d-wave pairing, and perhaps nearing the goal of room temperature operation.

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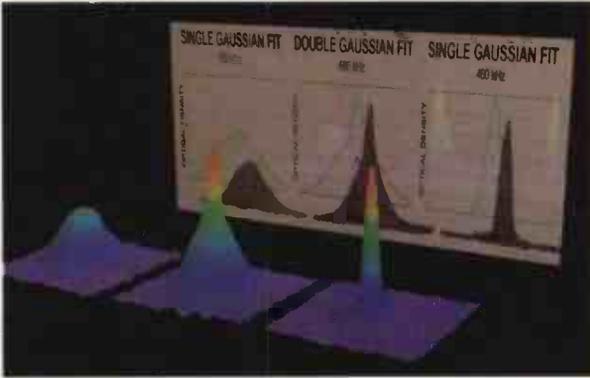
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Bose-Einstein postulate is condensed

A US group looks to have made a breakthrough in the decades-long effort to understand the rare state of matter known as the Bose-Einstein condensate (BEC). BEC, a mysterious quantum-mechanical state in which matter is coherent and has 'laser-like' properties, was first postulated more than 70 years ago by



Bose-Einstein condensation is a phenomenon in which a macroscopic number of atoms end up in the ground state of the trap. This means that we can produce a macroscopic system that exhibits quantum behaviour. This coherent state of matter is to 'normal' matter what laser light is to 'normal' light. The three images show the optical density of a dilute sodium gas as a function of position after the gas has been allowed to freely expand for several milliseconds. The images were taken at three different temperatures: one just above the transition temperature, one just below the transition temperature, and one well below the transition temperature. These 'time of flight' images show us the velocity distribution of our atoms. Above the transition temperature, as expected, the velocity distribution is a spherical gaussian. But as we cross the transition line, we see a sudden change. The distribution becomes bimodal, with two separate contributions from the ground state and from all other states. In the third picture, the temperature has been lowered enough so that most of the atoms are in the ground state, and the distribution looks gaussian again.

Indian physicist Satyendra Nath Bose and Albert Einstein. It has been observed in liquid helium and superconductors, and more recently in semiconductors.

Now work of professor Wolfgang Ketterle and his colleagues in the Research Laboratory of Electronics (RLE) and the Department of Physics at Massachusetts MIT is expected to allow scientists to move beyond demonstrating BEC to using it to study little-understood quantum mechanical effects.

BEC is achieved by chilling a gas of atoms to such a low temperature that the normal motion of the atoms is suddenly halted and they lose individual identity and display uniform behaviour as required by the rules of quantum mechanics. Because of this, scientists believe they may be able to study quantum effects on the rather large scale of several

hundredths of a millimetre (atoms are 10,000 times smaller).

The MIT advance has been to obtain Bose-condensed atoms with a very much higher production rate than previously, and reaching this level in 9s, compared with the several minutes required by devices used in the other experiments.

Keys to the achievement were a special arrangement of laser beams to collect and cool the atoms and a unique magnetic trap that kept the atoms from escaping.

The result: a Bose condensate with about 500,000 sodium atoms

BEC is a ubiquitous phenomenon which plays significant roles in condensed matter, atomic, nuclear and elementary particle physics, as well as in astrophysics.

According to the MIT team: "The study of BEC in weakly interacting systems holds the promise of revealing new macroscopic quantum phenomena that can be understood from first principles, and may also advance our understanding of superconductivity and superfluidity in more complex systems."

For more information please contact professor Ketterle at MIT – wolfgang@amo.mit.edu

War machine that won't stop

You might think "It doesn't get tired, it doesn't get hungry, it doesn't get sleepy – and it's expendable" sounds like a quote from Research Notes' favourite film The Terminator. In fact you wouldn't be far wrong. Because it is actually a project manager's chilling description of a prototype robot soldier that is currently being tested in the US.

'Sarge', being developed at Sandia National Laboratories, could well become standard battlefield equipment, serving as a 'force multiplier,' to increase soldier/Marine effectiveness and survivability. Its prime aim is to engage in remote surveillance, as evidenced by its full name – surveillance and reconnaissance ground equipment.

The final, complete tugv (teleoperated unmanned ground vehicle or 'tug-vee' to you, me and Norman Schwarzkopf) will be produced by the hundreds and put into the armed forces inventory. Individual or multiple robots will be assigned to infantry units and battalions.

Unlike the walking, talking, metal humanoid Hollywood robots with lasers for eyes, Sarge is a much simpler machine, and the latest in a long line of prototype

battlefield robots.

Its base platform is a commercial recreational 'four-wheeler,' a Yamaha Breeze, with the addition of a roll cage and four video cameras – two for surveillance and two for driving – on a pan/tilt platform.

Everything – steering, throttle, cameras – can be remotely operated from a suitcase-size operational control unit miles away.

Sarge's predecessor, Dixie, had to be teleoperated via a 1200-baud radio link.

Coupled with the slow speed of its processor, that caused a 75ms delay between user command and machine response.

Operators had to drive ahead, or plan for what was coming up to compensate.

With Sarge, the lag time has been much reduced. Its command/response delay is approximately 20ms, thanks to its much faster modern processors and communications equipment.

The developers hope that Sarge will make it unnecessary for a contingent of soldiers to have to go out on reconnaissance during combat, to determine the enemy's position and assess the situation.

Instead, the robot could be sent ahead, and images captured by its video cameras would

be relayed back. If there is an enemy ambush, the number of casualties would be less than one.

"Obviously, using a robot for surveillance is different from using a person," says project manager Bryan Pletta. "It's not going to be as good at some things as a person would be, with eyes and ears and a brain".

Gaining acceptance of the use of robotics among infantry soldiers may be a challenge too.

"Right now, using robotics is a pretty radical departure from the way they currently do things," Pletta says.

The next critical part of the project is the manufacture of eight to ten Sarge units to be given to infantry battalions, getting them involved in development up-front.

"The program will actually give them to infantry battalions and say, 'This is yours, keep it. Take it home, learn how to use it. Try and figure out what you could do with it if you had one,'" Take it home??

Take it home??

Bryan Pletta can be contacted at Sandia National Laboratories, Albuquerque, USA.

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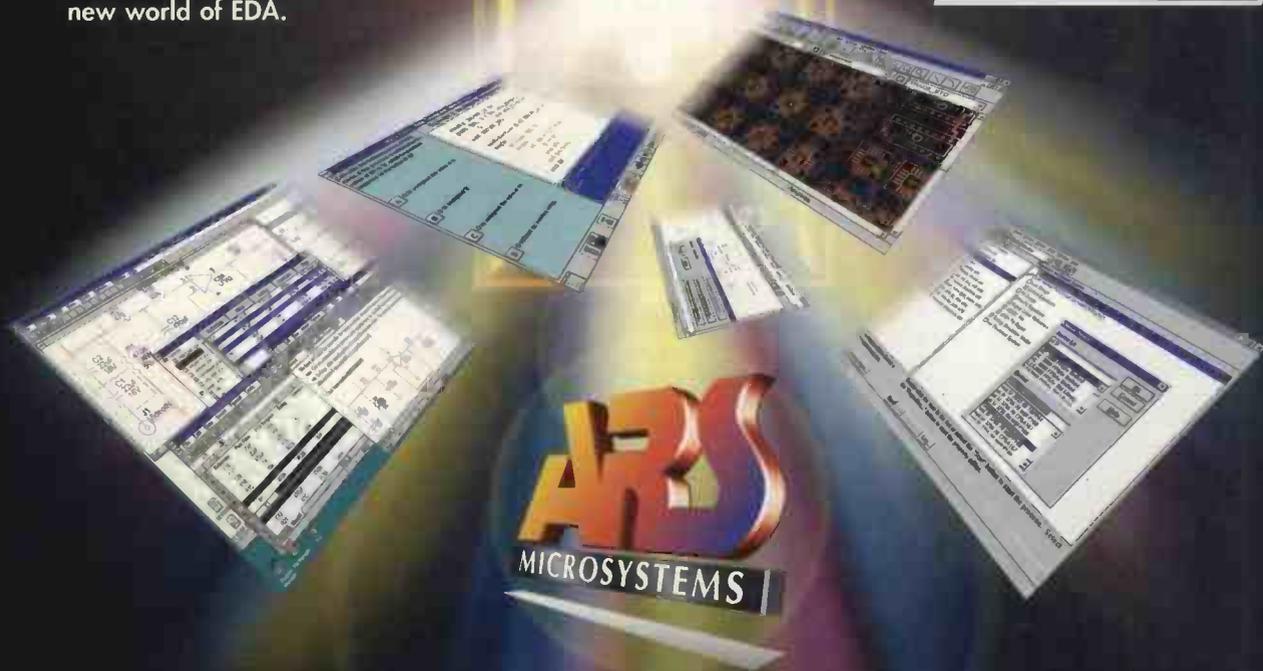
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Designing valve preamps

Morgan Jones details how to get the best from valves in hi-fi preamplifiers, but his article should prove equally useful to designers opting for transistors.



Preamplifiers receive a variety of signals at different levels, process them, and pass them to the power amplifier. They should do this without adding noticeable noise or distortion and they should be convenient to use.

Figure 1 is a block diagram of a typical pre-amplifier. Working back from the output, there is a line stage providing a limited amount of gain. This stage may be designed to drive long cables. It is preceded by the volume control and input selector; associated with this will probably be some form of switching to provide facilities for tape machines, which may, or may not, be buffered.

One of the sources to the selector switch will be an RIAA equalised disc stage, although many modern pre-amplifiers neglect this stage because it is so difficult to design. The excuse for this is that the long-playing record is obsolete. This may be so, but there are still many treasured collections of LPs that need to be played, so a proper pre-amplifier should include a disc stage.

The pre-amplifier only has to provide a very limited output voltage to the power amplifier; even the most insensitive power amplifiers will not require more than 4 or 5V rms to drive them into clipping. This means that the line stage may be optimised for linearity, rather than headroom, but it will have to drive the capacitance of the output cable without loss of high frequencies. It may need to be able to drive power amplifiers of lower input impedance, such as transistor amplifiers.

Since the stage will be preceded by the volume control, the power amplifier will amplify all of its self-generated input noise, so we will need to ensure that this noise is minimised, the stage also requires a low output impedance, coupled with low gain which ought to be quantified.

Low capacitance screened cable has a capacitance of around 100pF per metre. To avoid inducing hum from the power amplifier mains transformers into the pre-amplifier, it will probably be necessary to separate them by one metre. By the time the routing of the cable between the line stage and the input connector of the power amplifier has been allowed for, about 1.5m of cable will have been used. This is equivalent to 150pF. A valve amplifier will typically have input capacitance around 20pF,

so it should be possible to drive 170pF.

In combination with the shunt capacitance of the cable, the source impedance forms a low-pass filter whose -3dB cut-off we can calculate from;

$$f_{-3\text{dB}} = \frac{1}{2\pi CR}$$

It would be useful, however, to have the high frequency roll-off within the audio band to be far less than 3dB . As a result, it is necessary to find out what $f_{-3\text{dB}}$ corresponds to a given amount of loss at a given frequency. This can be found from,

$$f_{-3\text{dB}} = \frac{f_{(\text{dB limit})}}{\sqrt{\frac{1}{10^{\frac{\text{dB}}{10}}} - 1}}$$

As an example of using this equation, we find that for 0.1dB roll-off at 20kHz , we require $f_{-3\text{dB}} = 131\text{kHz}$. It should be noted that this formula is only valid for a single high-frequency CR or LR network. We can now determine that for 0.1dB loss at 20kHz , driving 170pF of capacitance, we need an output impedance of $7\text{k}\Omega$, and preferably less.

For a single low-frequency CR or LR network,

$$f_{-3\text{dB}} = f_{(\text{dB limit})} \sqrt{\frac{1}{10^{\frac{\text{dB}}{10}}} - 1}$$

Using this formula, 0.1dB roll-off at 20Hz corresponds to an $f_{-3\text{dB}}$ of 3Hz .

Once stages are cascaded, both high and low-frequency cut-offs begin to move towards the mid-band. For 'n' stages, each with identical low-frequency cut-off, the cut-off frequency of each individual stage is related to the cut-off of the composite amplifier by,

$$f_{-3\text{dB}(\text{individual})} = f_{-3\text{dB}(\text{composite})} \sqrt{2^n - 1}$$

Applying this formula to a three-stage ($n=3$) capacitor coupled amplifier, you will now find that the 3Hz cut-off for the entire amplifier requires each stage to have a 1.5Hz cut-off. The traditional value of $0.1\mu\text{F}$ coupling capacitor into $1\text{M}\Omega$ grid-leak gives a cut-off of 1.6Hz .

Traditional power amplifiers had input impedances of $1\text{M}\Omega$ or more; this is a useful impedance, because it allows a low value of coupling capacitor from the pre-amplifier. A value of 47nF almost meets our 20Hz 0.1dB criterion, but 100nF is better. Note that many modern valve amplifiers have an input impedance of $100\text{k}\Omega$,

requiring a 1µF coupling capacitor.

A power amplifier input stage using a triode with a sensitivity of 2V rms has excellent noise performance. But the higher gain of a pentode not only results in increased sensitivity – 125mV rms is common – but the intrinsically noisier pentode further worsens the already compromised signal to noise ratio.

Because of the previous low-frequency cut-off and noise considerations, valve pre-amplifiers should be designed to drive 2V into 1MΩ – even if it means modifying the power amplifier to achieve this match.

Typically, a sensitivity of around 250mV rms is needed at the input of the line stage. This results in an A_v of 8 for the line stage, but it may be useful to have 3dB to 6dB more than this, to allow for unusually low recording levels. As a result, maximum allowable gain A_v is 16, so an A_v of 12 would be fine.

This stage will be preceded by the volume control, which is discussed later. For the moment, it will suffice to simply state that it will probably be a 100kΩ potentiometer, whose maximum output resistance will be 25kΩ.

The question of potentiometer output resistance is crucial, because it forms a low-pass filter in conjunction with the input capacitance of the line stage. Using the earlier argument of 0.1dB high-frequency loss at 20kHz (–3dB at 131kHz), you can see that the maximum allowable input capacitance of the line stage is around 50pF.

If the input sensitivity of the stage is around 170mV, i.e. 2V/12, and a signal to noise ratio of 100dB or more is needed, then the self-generated noise of the stage referred to the input would be 170mV–100dB, which is 1.7µV. This is certainly achievable with triodes.

Together with the previous arguments, this results in a table of requirements as follows.

A_v	12
Z_{out}	7kΩ max.
C_{in}	50pF max.
V_{noise}	1.70µV
max.	
Output coupling	100nF

A good design is a simple design, so you should check to see whether the common cathode triode such as the ECC82, a low µ triode will suffice Fig. 2.

A_v	15.5
Z_{out}	7.7kΩ
C_{in}	30pF

Gain is certainly satisfactory, as is the input capacitance due to Miller effect and strays. The output impedance however is a little over the required value. Although the output impedance is higher than the ideal, input capacitance is considerably less than the allowable maximum.

Reduced roll-off at this point compensates for the slightly



Fig. 1. A typical pre-amplifier selects input signals at different levels, processes them, and passes them to the power amplifier.

increased high-frequency roll-off at the output. Provided that great care is taken with the capacitive loading of the output cable, or the stage is converted to a µ-follower, this would be a satisfactory solution.

Sadly, the ECC82's octal predecessor, the 6SN7, would have an input capacitance of around 70pF, because C_{ag} is 3.9pF, and would therefore only be suitable if a 50kΩ volume control could be tolerated.

Controlling volume

The volume control is an essential part of a pre-amplifier and should be treated with the same care as any other part of the design.

Human ears have a logarithmic response to sound pressure level, so if you want a volume control that has a uniform perceived response to adjustment throughout its range, you need to use a logarithmic potentiometer. This is the root cause of all our problems.

It is not a problem to make a linear potentiometer; all you need do is to deposit a strip of carbon of uniform width and thickness onto an insulator, put terminals at each end, and arrange for a contact to scrape its way round.

In an attempt to produce a logarithmic law, the coating thickness is made variable; in deference to audio sensibilities, a pressed metal screening can is fitted, and two potentiometers are ganged together on one shaft. Making the coating thickness continuously variable would be expensive, so the logarithmic law is approximated by a series of straight lines, Fig. 3.

It is surprising how good a fit to the ideal

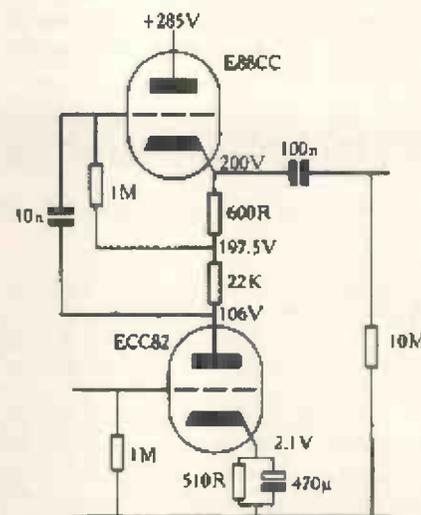


Fig. 2. Common cathode ECC82 triode stage suitable for preamplifier output use.

logarithmic curve can be made using only four different resistance tracks, but it will come as no surprise to you that this still results in steps in the response as the knob is rotated.

You would also expect the mechanically linked potentiometers to produce identical levels of attenuation all the way from 0dB to 60dB. Some of them are remarkably good, but the carbon track potentiometer's natural habitat is buried in the undergrowth of a television.

If quality is paramount, and a control that is not continuously variable is acceptable, a

Listing: QBasic routine for calculating potentiometer tap resistors.

```
CLS
A = 0
B = 0
N = 0
PRINT "This program calculates individual values of resistors between"
PRINT "taps of a potential divider string."
PRINT "How many switch positions can you use";
INPUT S
PRINT "What step size (dB)";
INPUT D
PRINT "What value of load will be across the output of the potentiometer";
INPUT L
PRINT "What value of potentiometer is required";
INPUT R
DO UNTIL N = S - 1
Y = ((R - L / 10 ^ (-A / 20)) + SQR((L / 10 ^ (-A / 20) - R) ^ 2 + 4 * R * L)) / 2
C = R - Y - B
PRINT A; "dB "; C; "ohms"
B = B + C
A = A + D
N = N + 1
LOOP
PRINT A; "dB "; R - B; "ohms."
```

switched attenuator can be used. Such an attenuator has conventional resistors connected to a switch in order to control volume. Adherence to the logarithmic law can now be perfect, as can channel balance. Commercially made switched attenuators are available with resistors fabricated directly onto the ceramic substrate of the switch wafer. Their performance is excellent.

The practical disadvantage of the switched attenuator is that you can only have as many different volume levels as switch positions. Although rotary switches are available with 30 positions – as opposed to the more usual 12 – this still limits us to 26 or 27 positions once an end stop is fitted.

For a normal volume control, it is often desirable to have a mute position, followed by

a –60dB position, then uniform steps all the way up to 0dB. Already 6 of the 30 positions have been used, so 60dB divided by 24 steps gives 2.5dB per step. This is too coarse, and commercial attenuators coarsen the lower levels to allow finer control at the upper levels, but this still only brings the basic step size down to 2dB.

If you don't mind wiring individual resistors

Requirements for RIAA equalisation

This panel defines the requirements for a high-performance RIAA equalisation circuit.

- Low noise and no hum. It has to be admitted that valves are not as quiet as the latest generation of low-noise IC op-amps, but they can be made quiet if you use dc heater supplies. Pentodes are too noisy, and care is needed when using triodes.

- Constant input impedance. This might seem obvious, but many designs have failed to appreciate this requirement. Cartridge manufacturers design for a specific loading of resistance and capacitance. They use this to equalise mechanical deficiencies of the generator system. This is particularly noticeable for moving magnet designs such as the older Shure and Ortofon models.

- Accurate RIAA. It is unbelievable how many designs have incorrect RIAA equalisation – ancient and modern. This is either down to a failure to use the correct equations, or to appreciate the loading conditions.

- Low sensitivity to component variation. Valves age, and as they do so, their anode resistance rises. When a valve is replaced, neither the new value of r_a , or C_{ag} , may be the same as the old valve. Neither of these effects should noticeably affect the accuracy of RIAA equalisation.

- Good overload capability. But what capability is necessary? Using a Tektronix TDS420 digital storage oscilloscope, the dynamic range of LPs was investigated in conjunction with a high quality record playing system. The TDS420 was first used in 'envelope' mode to find maximum output of the cartridge, and monitored an entire day of listening to music. The largest musical peaks were found while playing a Mobile Fidelity pressing of

Beethoven's 9th Symphony. Before equalisation these peaks rose to +16dB above the nominal 5cm/sec level, but clicks due to dust or scratches rose to about twice this level at +22dB, Fig. 6.

Individual clicks were then captured, and it was found that the vinyl/cartridge tip mass resonance was being excited, producing an oscillation at 56kHz for this particular moving coil cartridge, Fig. 7. If these ultrasonic signals were to overload the pre-amplifier, they would generate intermodulation products that would come back down into the audio band, and make the clicks much more noticeable.

- You should now allow for variable cartridge sensitivity of about 6dB; if you need more than this, the disc stage should be reconfigured. A good design should not operate permanently at its limits, so a further 6dB margin is desirable. This gives a total of 28dB in the audio band, rising to 34dB or more at ultrasonic frequencies. Very few pre-amplifiers of any age achieve this requirement simultaneously with low noise.

- Worn/old discs generate more ultrasonic energy than a new disc. This may be due to dirt ground into the groove walls, or because they were played by a cartridge that mistracked, causing wall damage as the stylus flailed from side to side of the groove. Inadequate ultrasonic overload margin is the reason why some pre-amplifiers will make worn records sound unplayable, but a good pre-amplifier is able to extract the best from the disc.

- Low distortion. This is an obvious requirement, and is linked to overload capability.

- Low output impedance. Ideally, the stage should be able to drive cable capacitance, so that it can drive a tape machine, or be sited within the plinth of the turntable.

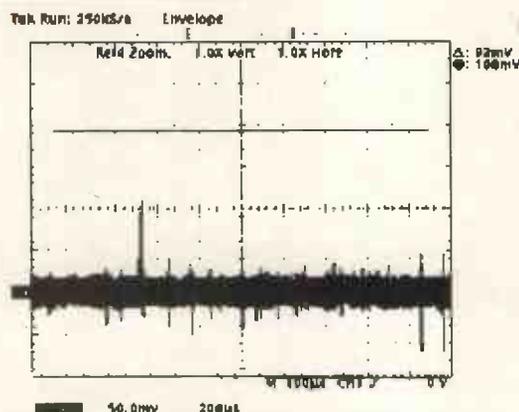


Fig. 6. Unequalised enveloped music output from cartridge. Peaks are dust/clicks. The rest is Beethoven's 9th.

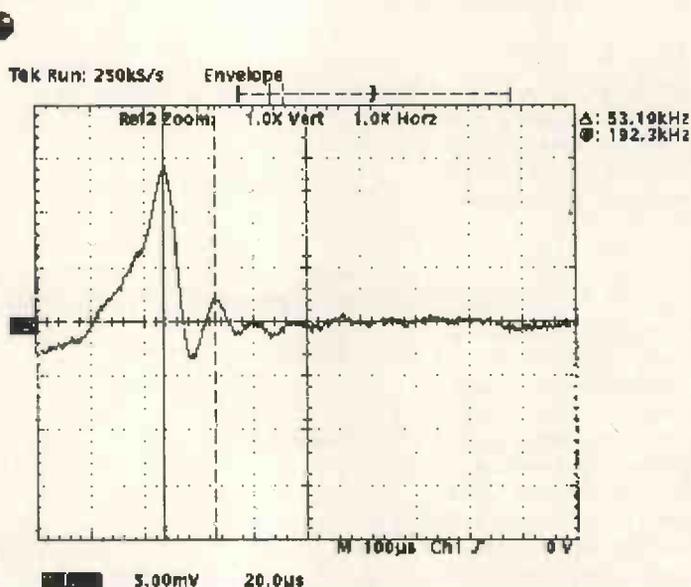


Fig. 7. Unequalised output from cartridge showing excitation of vinyl/tip-mass resonance.

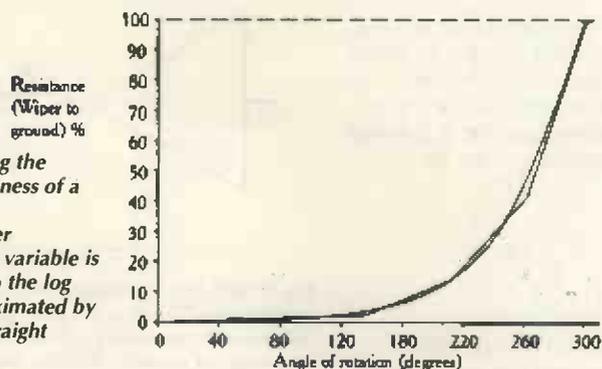


Fig. 3. Making the coating thickness of a logarithmic potentiometer continuously variable is expensive, so the log law is approximated by a series of straight lines.

onto switches yourself, you can do rather better than 2dB steps. It is necessary to design your pre-amplifier so well that all incoming signals arriving at the volume control are at precisely the same level. Additionally, the output of the pre-amplifier is perfectly matched to the power amplifier, and is only able to overload the amplifier on the last few steps of the volume control. This is feasible, and will allow you to use 1dB steps.

Usually, you will also need a separate mute switch in addition to the volume control. This type of arrangement is surprisingly convenient, and a matched, truly logarithmic volume control is a pleasure to use.

Assuming that you have a pair of 30-way, switches available to make a volume control, you need to calculate the values of resistors required. You could do this by hand, but a programmable calculator or computer makes life so much easier.

One form of switched attenuator is similar to the conventional potentiometer. It has a string of resistors from which you can take the appropriate tapping. The Qbasic Listing shown generates the resistor values for this attenuator; it is not a miracle of programming, but is quick and easy to use, and can easily be modified for different versions of Basic. It could also be made to run on a programmable calculator.

The program asks for the load resistance across the wiper. This is the grid-leak resistor of the following valve. It is tempting to try to use the potentiometer as the grid-leak, but this is poor practice, and can cause noise problems. It is also unnecessary, since the program accounts for its loading in designing the attenuator.

The final value given by this program is connected between the last useable switch contact and ground; it is often convenient to use one of the spare contacts on the switch as a ground terminal.

Ideally, an earthed metal screen should be fitted between the individual switch wafers in order to eliminate capacitive crosstalk between the stereo channels.

Selecting inputs

It is quite likely that you will have a number of alternative sources to the pre-amplifier, for example lp, cd, digital tv and radio. These will need to be selected to the volume control. You may also want to be able to record them onto tape, in which case you will need a tape loop, Fig. 4.

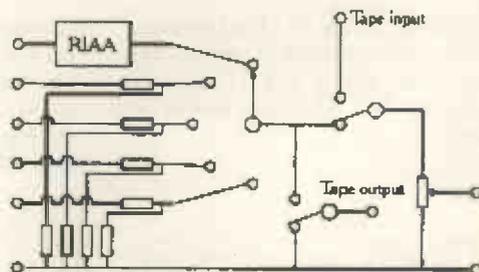


Fig. 4. Input selectors with switchable tape output. Attenuators and an RIAA equaliser preprocess the inputs to bring them to similar levels.

You will see that the arrangement is very simple. Incoming signals are routed through the input selector to the volume control via the tape monitor switch. This allows off-tape or source monitoring. The tape monitor switch should be of the on/off/on type in order to provide a centre mute position.

An unusual feature is the inclusion of a switch in the output to tape. All of the sources to the selector will be low impedance, and perfectly capable of driving the cable capacitance to a powered-up tape machine. As a result, a tape buffer is not required.

An unpowered tape machine, however, presents a non-linear load to the source. This load is in the form of unbiased transistor junctions, and may cause distortion. Most of the time, you will not be recording the source, so the tape output is switched off.

As the circuit stands, the rotary selector switch could suffer from crosstalk due to capacitance between adjacent contacts. On high-quality traditional pre-amplifiers, this problem was solved by having two switches. One selected the source, and the other deselected the short-circuit to ground on that source. Unfortunately, such wafer switches are no longer available, but a method that works almost as well is to use alternate contacts as inputs on a standard wafer switch, and connect the unused contacts to ground, which then guards the signal contacts.

A further advantage of the alternately grounded contacts is that if a tape loop is not required, the combined source/mute/tape switch may be dispensed with, since alternate positions of the selector switch provide the mute function, Fig. 5.

Criteria for RIAA equalisation

RIAA is the abbreviation for 'Recording Industry Association of America'. It is the worldwide standard for equalisation of 'microgroove' records, as opposed to the numerous standards for 78s. The European standard is known as IEC, and has an additional -3dB point at 20Hz (7950µs) on replay only in order to reduce rumble.

Most manufacturers of quality pre-amplifiers assume that their products will be complemented by equally good turntables, and that rumble will not be a problem, so they ignore the IEC recommendation. Their equalisation is therefore RIAA.

If power amplifiers were let down by their

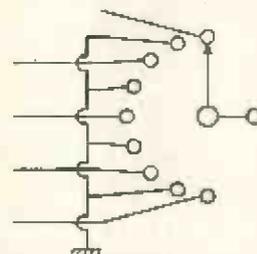


Fig. 5. Capacitive crosstalk on the input selector is reduced by guarding. A further advantage of this is that if a tape loop is not required, the combined source/mute/tape switch may be dispensed with.

phase splitters, then the Achilles heel of the pre-amplifier must surely be the RIAA disc stage. The stage has to satisfy so many contradictory requirements at the same time that its design and execution is fraught with problems.

The 'Golden Age' of valves produced power amplifier designs that became classics. There are no classic RIAA stages; they vary from mediocre to plain awful.

Requirements for an RIAA stage are shown in the panel. The next stage is to consider a suitable topology. Constant input impedance and low noise requirements eliminate shunt feedback. Low noise also rules out the pentode. You are therefore left with a combination of triode stages having active equalisation determined by series feedback, or with passive equalisation. Each of these contenders may be further broken down into performing the equalisation all in one go, or splitting it over a number of stages.

To tackle the problem of RIAA equalisation,

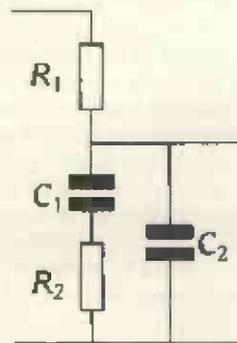


Fig. 8. Passive RIAA de-emphasis network. This is the only feasible network for a valve pre-amplifier.

it is necessary to define RIAA equalisation. The equalisation is specified in terms of time constants; 75µs, 318µs, and 3180µs. The RIAA equation plus some spot results are given below.

$$s=2\pi f$$

$$G_s = \frac{318 \times 10^6 \times s}{(1 + 3.18 \times 10^{-3} \times s)(1 + 75 \times 10^{-6} \times s)}$$

Frequency (Hz)	Gain (dB ref. 1kHz)
10	+18.802
20	+17.592
30	+16.558
40	+15.657
50	+14.862
70	+13.509
100	+11.899
200	+8.418
300	+6.289
400	+4.778
500	+3.612
700	+1.862
1000	0
2000	-3.789
3000	-6.204
4000	-8.034
5000	-9.522
7000	-11.876
10000	-14.502
20000	-19.915
30000	-23.218
40000	-25.605
50000	-27.474

From the table we see that considerable gain is needed at low frequencies, while high-frequency attenuation must continue indefinitely.

Because the high-frequency attenuation continues indefinitely, you can now exclude the series feedback 'all-in-one-go' topology. This is because the gain of this topology can only fall to unity. Although this failing can be exactly compensated after the feedback amplifier, it does mean that the response before compensation is rising. In turn, this means that ultrasonic overload capability within the amplifier is being compromised.

Because the 1kHz level is around 20dB below the maximum level at low frequencies, any 'all-in-one-go' passive network must have a minimum of 20dB of loss, and probably more. This is because the network will have the grid-leak resistor of the following valve in parallel with it, which will cause additional attenuation. You will find that it is extremely difficult to design a pre-amplifier of acceptable noise and overload capability using such a network, so this topology can also be excluded.

If you decide to use either of the two previous topologies, the relevant formulae are given in the definitive paper by Stanley P. Lipshitz, 'On RIAA equalisation Networks', *Journal of the Audio Engineering Society*, 1979 June, Vol. 27, No 6, pp. 458-481.

Of the four possible networks that Lipshitz gives, these reduce to two for passive equalisation. Of these two, only one has a capacitor

in parallel with the lower arm of the network. This feature is important because it allows stray and Miller capacitance to be taken into account. It is therefore the only feasible network for a valve pre-amplifier. Fig. 8. Relevant equations for this passive network are,

$$\begin{aligned} R_1 C_1 &= 2187\mu s \\ R_1 C_2 &= 750\mu s \\ R_2 C_1 &= 318\mu s \\ C_1 / C_2 &= 2.916 \end{aligned}$$

These numbers have not been rounded.

Remember that any grid-leak resistor in parallel with the lower arm of the network, or

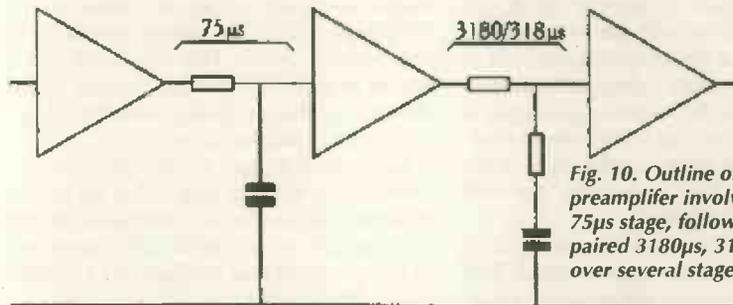


Fig. 10. Outline of an RIAA preamplifier involving a passive 75µs stage, followed by passive paired 3180µs, 318µs constants over several stages.

non-zero output impedance of the driving stage, changes the effective value of R_1 as seen by the network. Therefore, the values for the network must be calculated using the Thévenin impedance seen by that network. Likewise, any stray or Miller capacitance will need to be subtracted from the calculated value of C_2 .

For any 'all-in-one-go' topology other than the above network, it is essential to refer to the Lipshitz paper, and read it thoroughly before embarking on design.

You are now left with only two possibilities for equalisation – split active, and split passive. The first job is to define how to split the equalisation. Fortunately, there is only one rational way to split the equalisation, and that is to pair the 3180µs with the 318µs, but to perform the 75µs separately.

The 75µs time constant defines a low pass filter whose -3dB point is at around 2122Hz and rolls off at 6dB/octave thereafter. This is an ideal filter for use early in the pre-amplifier since it allows high-frequency overload capability after that stage to rise at 6dB/octave above cut-off. This is exactly what is needed.

It is usual to perform the 75µs time constant passively following the input stage. This has the advantage of ensuring that the impedance seen by the cartridge is constant with frequency, apart from input capacitance, for which a value is usually specified by the cartridge manufacturer, Fig. 9.

The reason for choosing a passive network is that a series feedback amplifier cannot achieve an A_v of less than unity, and a shunt feedback amplifier would have noise problems.

Additionally, although it was not noted earlier, a feedback amplifier attempting this response would find its output stage faced with a heavy capacitive load. This capacitive

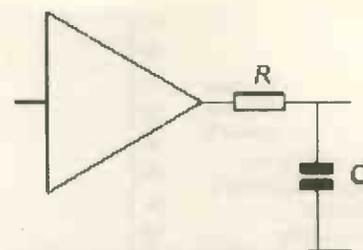


Fig. 9. Split RIAA de-emphasis. It is usual to perform the 75µs time constant passively following the input stage. This has the advantage of ensuring that the impedance seen by the cartridge is constant with frequency, apart from input capacitance.

load demands a large current at high frequencies, and would be equivalent to changing the ac loadline to a far lower value of load resistance. The result would be additional distortion before the feedback loop was closed.

Note that all of the previous observations are equally relevant to discrete semiconductor or IC based pre-amplifiers.

The 3180µs, 318µs pairing defines a shelf response with a level variation of exactly 20dB. Using IC op-amps it is equally convenient to perform this actively or passively, but with valves it is more convenient to use passive equalisation.

The preceding description allows you to define the optimum way of achieving RIAA equalisation in a valve pre-amplifier. Assume a passive 75µs stage, followed by passive paired 3180µs, 318µs over several stages of triodes. All you need now do is to define the topology and operating conditions of each stage, and calculate component values.

It is now possible to draw a block diagram of the pre-amplifier, Fig. 10. Note that the block diagram has completely ignored practicalities such as coupling, or decoupling, capacitors and grid-leak resistors. Nevertheless, it represents a simplicity of design to which we should aspire, i.e. dc coupling throughout. This ideal is achievable, but it is not ideal for the novice designer; you will need to be a little more cautious in your first attempt.

Morgan will discuss implementing the valve RIAA stage, psrr and balanced preamplifiers in a second article.

Further reading

- Lipshitz, S P, 'On RIAA equalisation Networks', *Journal of the Audio Engineering Society*, 1979 June, Vol. 27, No 6, pp. 458-481.
- Wright, A, 'The tube pre-amp cookbook' 1994.

TELFORD ELECTRONICS

HP EQUIPMENT

HP8161A Programmable Pulse Generator (opt. 001-020)
 HP5370B (HPBI) Universal Time Interval Counter
 HP8445B Automatic Preselector (opt. 002-003)
 HP8656A Signal Generator (opt. 001)
 HP8901A Modulation Analyzer
 HP3366B Selective Level Meter
 HP5335A Universal Counter (opt. 020-040)
 HP3336B Synthesizer Level Generator
 HP3570A Network Analyzer
 HP8015A Word Generator
 HP3571A Tracking Spectrum Analyzer
 HP3330A Automatic Synthesizer
 HP3334A Distortion Measurement Set
 HP5328A Universal Counter
 HP11722A Sensor Module
 HP8015A Pulse Generator
 HP8532A Frequency Meter
 HP8761A S.P.D.T. RF Switch
 HP435A (B) Power Meter c/w RF Head
 HP432A Power Meter c/w RF Head
 HP170A Logic Pattern Generator
 HP9501A Power Supply Programmer
 HP9307A VHF Switch
 HP9306A Relay Actuator
 HP461A Amplifier
 HP400FL A.C. Voltmeter
 HP3200B VHF Oscillator
 HP9502A Transmission Reflection Test Set 500kHz-1.3GHz
 HP3565D System (8 slot) c/w HP3565ZB x4, HP35653A, HP35651B
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 HP203A Variable Phase LF Generator
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 HP5305A 1100MHz Counter

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 TEK2465A 35MHz Oscilloscope
 TENP9042 Current Probe D.C.-50MHz
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 2830 Multiplier Tester
 2829 Digital Analyzer
 2828A Digital Simulator
 2831 Channel Access Switch
 2833 Digital Line Monitor
 MARCONI Microwave Education Test Bench 'X' Band - Brand New
 2015 Signal Generator 10-520MHz
 2019 Signal Generator 80MHz-1040MHz

OTHER TEST EQUIPMENT

ROHDE & SCHWARZ SMS Signal Generator 0.4-520MHz
 ROHDE & SCHWARZ SMUV Signal Generator 10MHz-130MHz
 WATSU SAS8130 Waveform Analyzer c/w SH-16 P.I.D.C.-3.5GHz
 WATSU DM2350 Digital Memory 10Bt 20ns
 WATSU DM6430 Digital Memory Scope
 NIKUSU TOS8850 W1 Auto Tester
 NARDA SMA Coaxial 90° Hybrid 2-4GHz
 NARDA SMA Miniature Stripline Coupler 2-4GHz
 NARDA Various SMA Attenuators DC-6GHz DC-12GHz
 SOLARTRON 7045 Digital Multi Meter 4 1/2 Digit
 FLUKE 8860A Digital Multi Meter
 FERROGRAPH Recorder Test Set RTS2
 TOOL CHESTS (8 drawer) Made by H Fine & Son (BRAND NEW)
 CLARE Flash Tester Model GCH01P
 RACAL 9081/9082 520MHz Synthesised Signal Generator
 GQULD 338 10Hz-1000Hz Low Distortion Oscillator
 ADVANCE HIE 15-500MHz LF Oscillator
 RADFORD LD01 Low Distortion Oscillator
 RADFORD LDMS22 Low Distortion Measuring Set
 PHILIPS PM5132 Function Generator 0.1Hz-50MHz
 PHILIPS PM5715 Pulse Generator 1Hz-50MHz
 ADRET Type 2230A
 ROHDE & SCHWARZ SUF2
 FARNELL TSV70 MK2 0-70V @ 5A/0-35V @ 10A
 VARIOUS LAMBDA & KEPCO PSU'S IN STOCK
 HAVEN Temperature Calibrator DTB-5 oil/water bath
 HAVEN Thermo Cal IS Thermocouple Simulator/Calibrator

TIME CD Millivolt POT Source Model 404N

LEADER LOC705 Scope Calibrator
 CASSELLA Y6C1
 PPM 411F Current Reference
 CROPHO Resistance Standards - Various In stock
 W TINSLEY Resistance Bridge Type 5761
 WALLACE & TIERMAN Precision Pneumatic Calibrator
 MICRODINE Telemetry Receiver 2.2-2.3GHz
 AVO DALL16 Digital Multi Meter
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 CORNIX 4B-4B Audio Switch Matrix Type 921B/1
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Fractional-N synthesisers

Cosmo Little discusses implementing the fractional-N synthesiser, and shows how adding a second accumulator to the basic synthesiser system described last month reduces low-frequency spurs.

In my first article, I discussed the basic fractional-N synthesiser with a single accumulator, and provided a simulation of its performance. In this article, I examine a modification to the method which uses a second accumulator to generate a different sequence of divisors. This important enhancement reduces the low frequency spurs seen in the previous simulation.

In addition, I discuss here some methods of implementing the fractional-N synthesisers in hardware and software.

Two-accumulator fractional-N loop

Invented by Racal, this modification to the fractional-N loop is based on the idea that it is possible to change the frequency spectrum of the error waveform. This avoids the large discrete spurs that appear at low frequencies when generating a small fractional increment. This could be possible by dividing by a greater range of numbers based on N .

Obviously, it is feasible to divide by any number, provided that the average over a full cycle of accumulator additions – a maximum of the accumulator modulus – equals the required fractional divider.

Numbers are selected on the following basis. The loop filter is an integrator at low frequencies and as a result integrates the error voltage which builds up as a ramp as described last month. If a large negative error voltage is generated at the right intervals, the integrator can be reset to zero. This negative error voltage could be produced by dividing by $N+1$, and then immediately by $N-1$ at the next reference cycle.

Provided that a division by $N+1$ always has a corresponding division by $N-1$, the mean loop divisor has not been altered. As the normal control mechanism requires periodic division by $N+1$, the combination of the two controls results in divisions by $N-1$, N , $N+1$, and $N+2$.

I have never found this argument – ref 2, p201 – convincing as the error voltage is modified by the loop forward transfer response, not the loop filter. Forward transfer response is a constant multiplication by N at low frequencies. However the above algorithm certainly works, as will be seen later.

The second control may be implemented by means of a second accumulator, which accumulates the contents of the normal accumulator. These will be referred to as accumulators A and B. As an accumulator is an integrator, overflow of accumulator B can be used to initiate a correction cycle of division by $N+1$ and $N-1$.

A simple combinational logic function can be designed to combine the overflow state of both accumulators and the previous overflow state of accumulator B which must be stored in a latch.

Operation of accumulator B is shown in the document **Program 1**, as is the effect on the time error waveform. This waveform still has about the same peak to peak amplitude as that for the single accumulator fractional-N loop, but the peaks are only reached on pulses. These pulses contain less energy at low frequencies. If you make a mental average through the area of pulses, you will find that the main low frequency waveform is a triangle wave of much lower amplitude.

A correction waveform may be constructed for the two-accumulator fractional-N loop, but the relationship to the contents of accumulator A is more complicated. The rule is that if B has overflowed in the previous reference cycle, then the value in accumulator A must be complemented – i.e. subtracted from the modulus.

A 5% accuracy of the correction voltage has been assumed in the document, and all other parameters are the same as for the right-hand *MathCad* document in the last issue of *EW*. This allows the Fourier transforms of the error

voltages to be directly compared. But note different Y axis) The improvement is dramatic. The frequency range in both graphs extends to $f_{ref}/2$, so a typical natural loop frequency would be 5kHz. Frequency components higher than this are progressively filtered out.

Automatic correction of the correction voltage by synchronously rectifying the error voltage is possible for the two-accumulator loop. This turns out not to depend upon accumulator A at all, but only upon the overflow state of accumulator B. The graph in the document shows that the function does work.

There is a possibility that other algorithms could result in even lower low frequency spurious content to the error voltage. You could divide by a larger set of integers, for example. Alternatively, there might be a better sequence of the set of four integers than that generated by the two accumulator implementation.

Implementation the fractional-N loop

The lack of widespread use of the fractional-N technique may partly be due to the lack of any commercial large-scale-integration digital chip for implementing the accumulators and correction circuits required. This contrasts with the large number of single-loop synthesiser chips available, most with provision for controlling dual modulus prescalers, and having microprocessor interfaces for setting up the various internal dividers.

Figure 1 gives a possible block diagram for a single chip implementation of the two accumulator loop. Everything is included except the d-to-a converters for generating the correction voltage. Because of the difficulty of implementing the inverse scaling of the correction voltage with change of vco frequency, a simple latch is provided for controlling a d-to-a converter. It is assumed that the controlling microprocessor will calculate the scaling factor.

For many applications binary accumulators

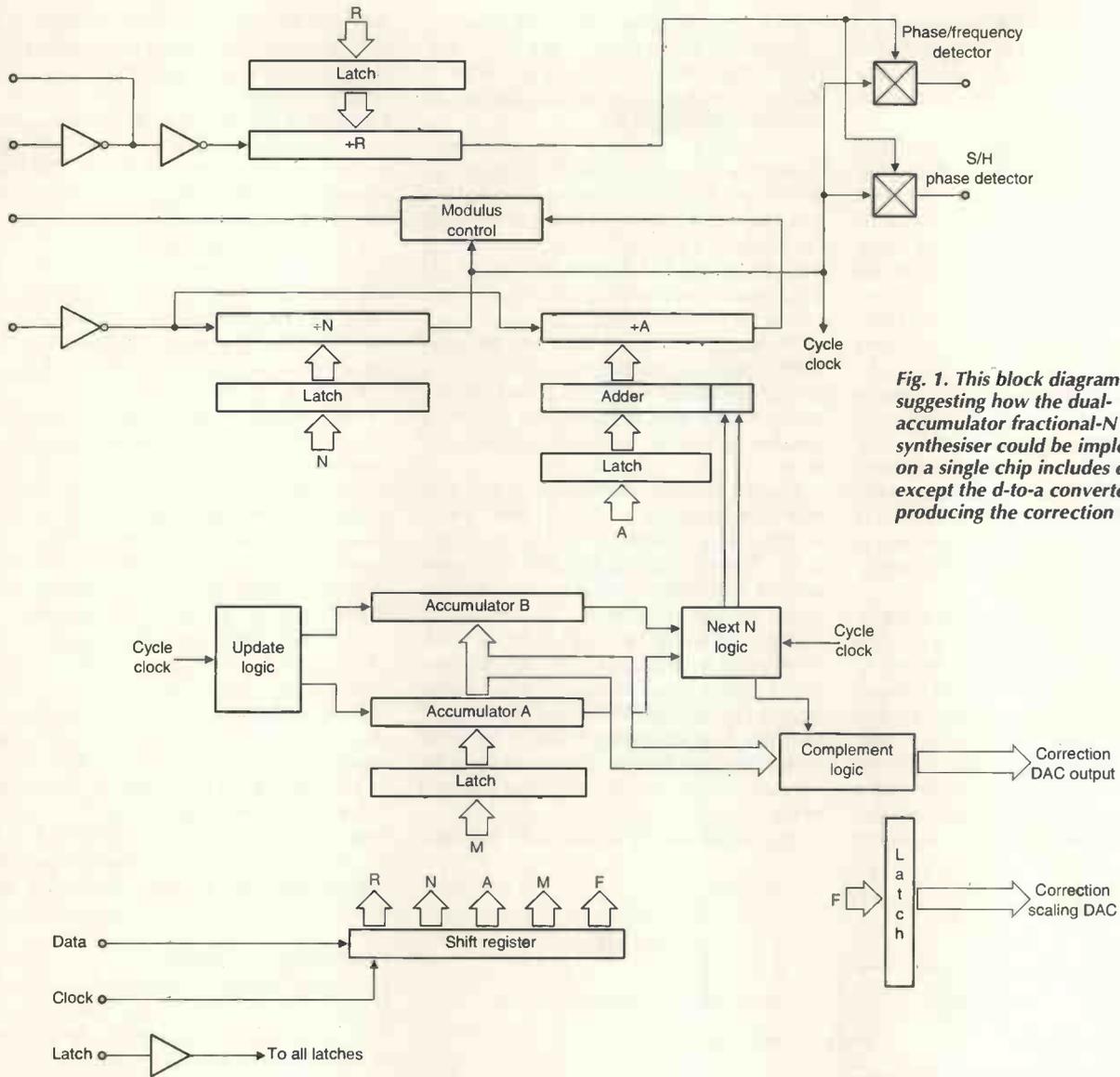


Fig. 1. This block diagram suggesting how the dual-accumulator fractional-N synthesiser could be implemented on a single chip includes everything except the d-to-a converters for producing the correction voltage.

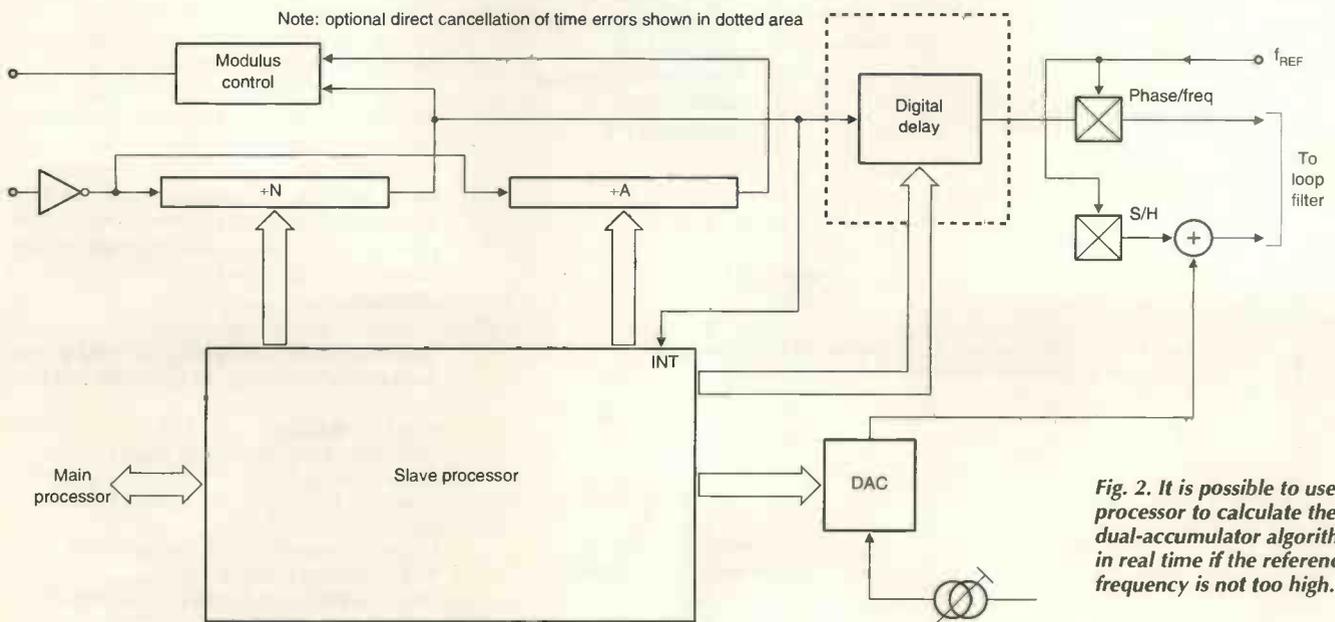


Fig. 2. It is possible to use a processor to calculate the dual-accumulator algorithm in real time if the reference frequency is not too high.

are adequate. This means that the output can only be an approximation to a decimal frequency. For some applications such as signal generators, binary-coded-decimal accumulators will be needed.

It is important that the phase detector used should have very good linearity at small phase offsets. Otherwise the error voltage from the phase detector will not match the correction voltage. Digital tristate phase/frequency detectors are very poor in this respect, and should only be used for initial acquisition. A sampling phase detector should be used for tracking. This approach is used in the GEC/Plessey range of chips.

Until a manufacturer decides to make a suitable lsi chip, low budget designers must look for other ways to implement the synthesiser. Single chip microprocessors are cheap, and are powerful general purpose logic machines. It is feasible to use a processor to calculate the two accumulator algorithm in real time provided that the reference frequency is not too high.

A possible block diagram is shown in Fig. 2. The processor must be synchronised to the overflow of the loop divider, either by an interrupt or by polling a port. The processor may take most of a reference cycle to calculate the next value of N and the appropriate error voltage does not appear at the phase detector output until the loop divider has finished a cycle. As a result, the correction voltage will need to be delayed internally to the processor

to keep it synchronised to the error voltage.

It is also assumed in Fig. 2 that the processor will scale the correction voltage output to the correction d-to-a converter, either by calculation of lookup table.

I constructed a prototype fractional- N synthesiser for an high-frequency receiver design about a decade ago. It was based on the two accumulator algorithm and used an 8041 slave processor to calculate the algorithm in real time. The reference frequency was 10kHz, and the processor used double precision arithmetic to give 16-bit accumulators.

The processor ran at 6MHz, and the algorithm just fitted into the 100µs available. The correction did not use d-to-a converters, but corrected the time errors before the phase detector by the use of binary increments of delay.

I used a miniature analogue delay line with four sections of 80, 40, 20, and 10ns. These sections were switched in and out of circuit using bipolar transistors. The prototype worked reasonably well, but it proved difficult to get the delays accurate. This method of directly correcting the time errors is worth considering as an alternative to the usual system of generating a correction voltage. A chip such as the Analogue Devices AD9501 which provides 8-bit resolution of delay with sub nanosecond resolution would seem ideal for the purpose. This correction system is shown in Fig. 2.

Microprocessors have advanced consider-

ably over the past decade, and it is likely that a 16-bit digital signal processor could calculate the algorithm in 10µs, making possible a 100kHz reference frequency. For even faster implementations, the processor could calculate the algorithm in advance, and store all the values of N and the correction in ram. The processor could then respond to an interrupt and output the next values in less than 1µs.

A low-cost implementation is possible by using a single chip synthesiser with dual modulus prescaler control, and adding logic to delay the change of modulus from $P+1$ to P by up to three input cycles. This will change the overall division ratio. This idea is outlined in Fig. 3.

Programming of the ' N ' and ' A ' counters must be altered from the usual arrangement to ensure that the four divisors are available without requiring the ' A ' counter to be zero, as otherwise the logic would not work. A little thought will show that provided the ' A ' counter has a higher modulus than the prescaler, P , then programming for any four adjacent integers will always be possible.

This implementation would require only 1x2 bit and 1x8 bit port on the processor, and one spare interrupt. If sufficient ram was available to store the divisors and corrections, then the main system processor might have sufficient capacity to run the synthesiser as well.

One advantage of the ram storage technique is that decimal accumulators can be as easily programmed as binary. As the algorithm is calculated once only for each frequency change, the extra time required to calculate for decimal accumulators would not matter.

Predicting vco spuri

One final MathsCAD document is offered, Program 2. This 'document' calculates the level of the vco sidebands due to the fractional- N mechanism. It communicates with either Programs 1 or the right-hand MathCad document in the last issue of EW using a data file, so the appropriate program should be run first. It is also necessary to design a sensible phase-locked loop using the phase-locked-loop MathCad document presented last month.

Loop time constants are copied across manually to Program 2. The document then calculates the voltage-controlled oscillator spurious sideband levels. The simplification is made that the voltage-controlled oscillator phase modulation index is small, so each component of the phase modulation of the voltage-controlled oscillator only results in one pair of sidebands.

Cosmo is head of the rf design, prototyping and consultancy company, RF Solutions, Penryn.

Further reading

- Digital PLL Frequency Synthesisers, Ulrich L. Rhode, Prentice Hall.
- Frequency Synthesis by Phase Lock, William F. Edan Wiley.
- Frequency Synthesisers, Theory and Design Vadim Manassewitsch, Wiley.
- Phase Locked Loops. Application to coherent receiver design, Alain Blanchard, Wiley.

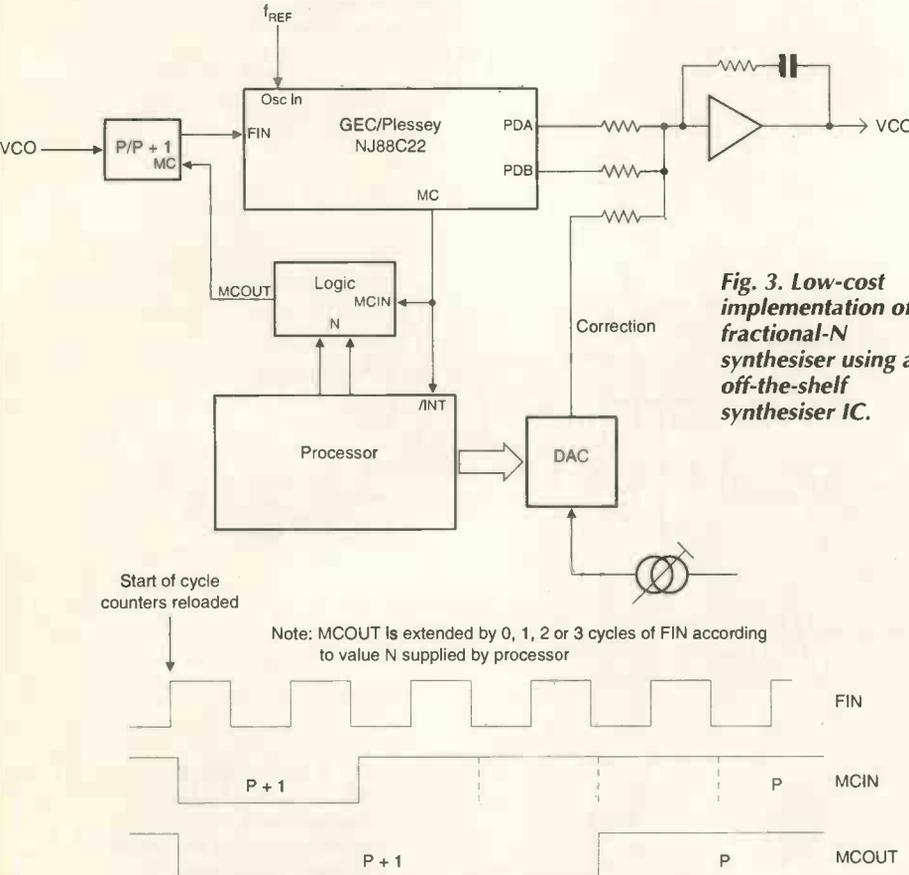


Fig. 3. Low-cost implementation of the fractional- N synthesiser using an off-the-shelf synthesiser IC.

Program 1. This MathCad document is for evaluating operation of the fractional-N synthesiser's second accumulator, designated B. It also shows how the time error waveform is affected.

Phase lock loop comparison frequency, $f_{ref} := 10^5$. Integer part of loop divider, $N := 10$.

Fractional part of loop divider, $M := 7$. Accumulator modulus, $Modulo := 2^8$.

Correction DAC modulus, $DAC := 2^8$.

$$f_{VCO} := f_{ref} \left(N + \frac{M}{Modulo} \right) \quad T_{ref} := \frac{1}{f_{ref}} \quad i := 0..(Modulo - 1) \quad A_0 := 0 \quad B_0 := 0 \quad N_{D_0} := N + 1 \quad T_{error_0} := 0$$

$$f_{VCO} = 1002734.375$$

The next two statements form the vectors of values in the two accumulators. $A_{i+1} := \text{mod}(A_i + M, Modulo) \quad B_{i+1} := \text{mod}(B_i + A_{i+1}, Modulo)$

Special vector function definition: returns 1 if next entry in vector is less than current entry. Serves to mark accumulator overflows. If index is <0 returns 0. $OVFL(V, j) := \text{if}(j < 0, 0, \text{if}(V_{j+1} < V_j, 1, 0))$

Logic section to create vector of divisors based on overflow of the two accumulators. Order of statements is important

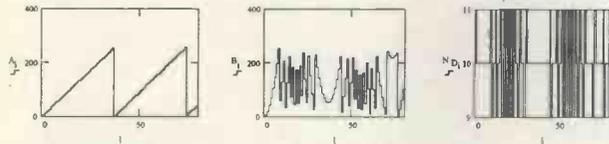
$$N_{D_{i+1}} := N$$

$$N_{D_{i+1}} := \text{if}(OVFL(B, i) = 1, N_{D_{i+1}} + 1, N_{D_{i+1}})$$

$$N_{D_{i+1}} := \text{if}(OVFL(B, i - 1) = 1, N_{D_{i+1}} - 1, N_{D_{i+1}})$$

$$N_{D_{i+1}} := \text{if}(OVFL(A, i) = 1, N_{D_{i+1}} + 1, N_{D_{i+1}}) \quad \text{mean}(N_D) = 10.027818$$

Graphs of accumulator content and loop divider



Create vector of time errors due to non exact division. $T_{error_{i+1}} := T_{error_i} + T_{ref} \frac{N_{D_{i+1}} - 1}{f_{VCO}}$

Create correction voltage from accumulator A. Note that Accumulator is truncated to no of DAC bits.

$$Corr_i := \text{if}(OVFL(B, i) = 1, \text{DAC} - \text{floor}\left(\frac{A_i}{Modulo} \cdot \text{DAC}\right), \text{floor}\left(\frac{A_i}{Modulo} \cdot \text{DAC}\right)) \left(\frac{f_{ref}}{f_{VCO} \cdot \text{DAC}} \right) \cdot 2 \cdot \pi \cdot 1.59$$

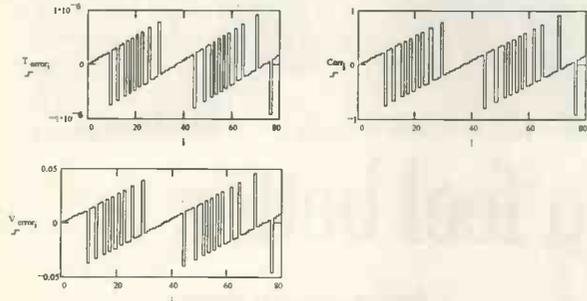
Create error voltage vector at output of phase detector.

$$V_{error_i} := \frac{T_{error_i}}{T_{ref}} \cdot 2 \cdot \pi \cdot 1.59 - 95 \cdot Corr_i$$

Note that correction is assumed to be accurate to 5%

Remove DC component from error voltage $MV := \text{mean}(V_{error}) \quad V_{error} := (V_{error} - MV)$

Graphs of time error, correction, and voltage error

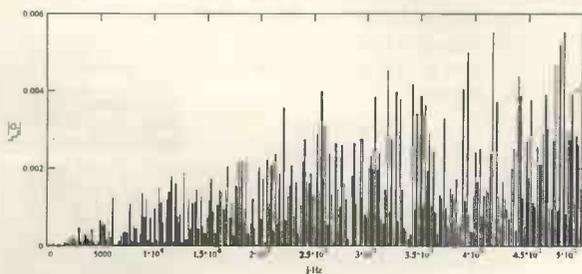


Fourier transform of error voltage. Bin spacing = $f_{REF}/\text{no of samples}$

$$G := 2 \cdot \text{FFT}(V_{error}) \quad \text{length}(V_{error}) = 2.048 \cdot 10^3 \quad \text{length}(G) = 1.025 \cdot 10^3 \quad \text{Hz} := \frac{f_{ref}}{\text{length}(V_{error})}$$

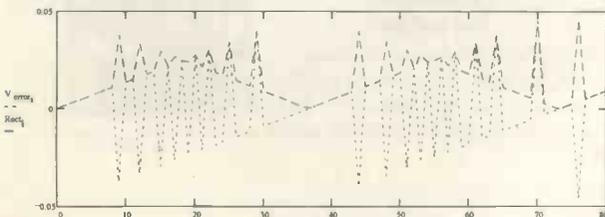
$$j := 0.. \text{length}(G) - 1 \quad \text{WRITE}(freq1) := \text{Hz} \quad \text{WRITEPRN}(freq1) = 0$$

Spectrum of error voltage (y axis in volts eqiv to peak voltage of a transformed sine wave)



Automatic adjustment of error voltage correction by synchronous rectification of V_error

$$Rect_i := \text{if}(OVFL(B, i - 1), -V_{error_i}, V_{error_i})$$



mean(Rect) = 0.0167

Program 2. MathCad routine worksheet for calculating vco sidebands due to the fractional-N synthesiser's mechanism.

Loop fixed parameters:

op amp low frequency pole and open loop gain: $f_0 := 10 \quad A_0 := 100000$

loop filter component values: $R1 := 4700 \quad R2 := 10000 \quad C1 := 10 \cdot 10^{-9} \quad C2 := 5 \cdot 10^{-9}$

Phase detector gain constant (volts per radian) $Kp := 1.59$

VCO gain constant (radians per volt second) $Kv := 50 \cdot 10^3$

VCO modulation pole(Hz) $f_{mod} := 30 \cdot 10^3$

Loop division ratio $N := 10$

Additional low pass pole(Hz): $f_{pole} = 20000$

Data from previous fractional N calculations:

$\text{Hz} := \text{READ}(freq1) \quad \text{SPUR} := \text{READPRN}(freq1) \quad i := 0.. \text{last}(\text{SPUR})$

$f_i := i \cdot \text{Hz} \quad \text{last}(\text{SPUR}) = 1.024 \cdot 10^3 \quad \text{length}(\text{SPUR}) = 1.025 \cdot 10^3$

calculated values $t_0 := \frac{1}{2 \cdot \pi \cdot f_0} \quad t_1 := C1 \cdot R1 \quad t_2 := R2 \cdot (C1 + C2)$

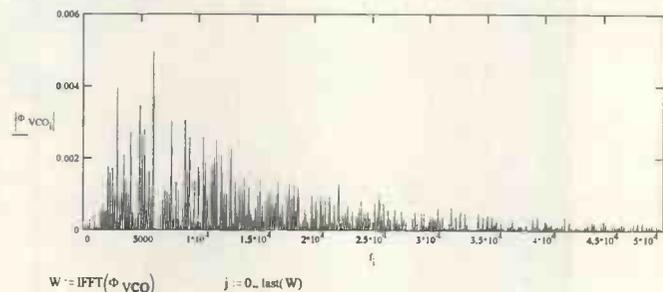
$$t_3 := C2 \cdot R2 \quad t_{mod} := \frac{1}{2 \cdot \pi \cdot f_{mod}} \quad t_{pole} := \frac{1}{2 \cdot \pi \cdot f_{pole}} \quad s_1 := 2 \cdot \pi \cdot j \cdot f_i$$

$$F(s) := \frac{A_0 (1 + s t_2)}{s^2 t_0^2 t_1 t_3 + s^2 (A_0 t_1 t_3 + t_1 t_3 + t_0^2 t_1 + t_0^2 t_2) + s (A_0 t_1 t_1 + t_1 t_2 + t_0) + 1} \left(\frac{1}{1 + s t_{mod}} \frac{1}{1 + s t_{pole}} \right)$$

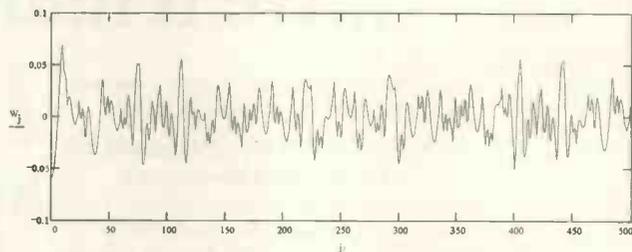
$$G(s) := \frac{Kp \cdot Kv \cdot F(s)}{s} \quad B(s) := \frac{G(s)}{1 + G(s) \cdot H} \quad H := \frac{1}{N} \quad t_{mod} = 5.30516 \cdot 10^{-6} \quad t_0 = 0.01592 \quad t_1 = 4.7 \cdot 10^{-5}$$

$$\Phi_{VCO_i} := \text{if}(i = 0, 0, \left(B\left(\frac{s_1}{Kp}\right) \right)) \quad t_{pole} = 7.95775 \cdot 10^{-6} \quad t_3 = 5 \cdot 10^{-6} \quad t_2 = 1.05 \cdot 10^{-4}$$

This graph is peak phase deviation of the VCO

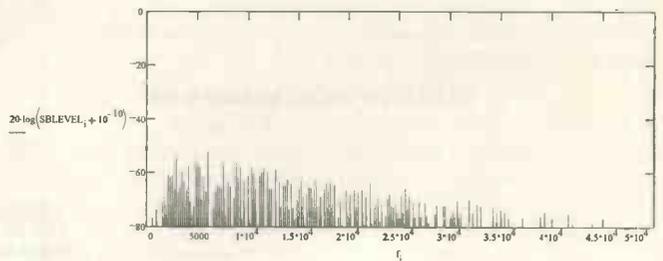


VCO phase deviation in time domain



$$SBLEVEL := \sqrt{11} (|\Phi_{VCO}|)$$

VCO single sideband spur



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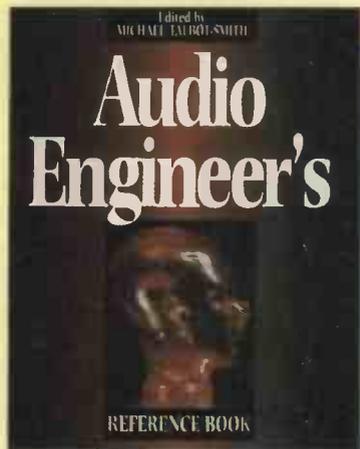
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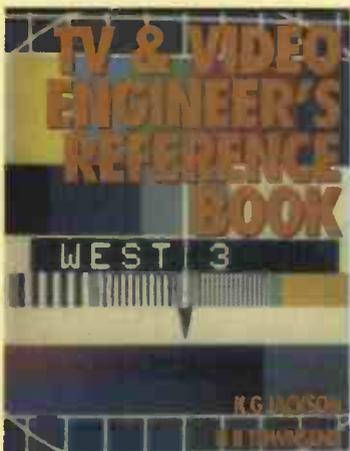
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Hence, the design given here is based on pwm control.

Moreover, this design can generate high duty cycles and the motor switching components have low 'on' resistances. This leads to high efficiency at maximum speed so that nearly full power may be attained.

In this design, motor power is produced via a solid state complementary bridge driver, Tr_{3-6} on Fig. 1.

An optional alteration to the design appears in Fig. 1b. This makes the circuit fully compatible for control via a servo arm.

To keep the cost down, the only relatively expensive components in the design are output mosfets Tr_{3-6} . When choosing these transistors, a compromise needs to be made between low $R_{DS(on)}$ and low price. In practice having a low $R_{DS(on)}$ is more important for motors requiring high currents. This is because whatever the on resistance of a switching mosfet, more power is dissipated in the device as the current through it increases.

With these considerations in mind, the Table 1 gives some possible types for Tr_{3-6} along with prices and corresponding distributors.

Circuit details

Resistors R_{1-6} , together with VR_1 , C_{1-3} , D_1 , Tr_1 and Tr_2 form a linear sawtooth generator. Without R_{4-6} , VR_1 , C_3 , Tr_2 , and with D_1 shorted, the circuit would function as a standard

unijunction transistor relaxation oscillator with a repeating cycle as follows.

Capacitors $C_{1,2}$ charge through $R_{2,3}$, the voltage at point [1] rising towards 12V according to a negative exponential curve.

When the voltage at [1] reaches the peak point voltage of the unijunction device, the emitter becomes forward biased and the dynamic resistance between the emitter and base 1 drops dramatically to a low value. Capacitors $C_{1,2}$ are then rapidly discharged through the emitter.

When voltage at [1] drops below a critical value – usually about half the saturated emitter voltage for the given circuit – the unijunction transistor ceases to conduct and the cycle starts again.

The signal at point [1] is indicated in Fig. 2. With the improved circuit, comprising all the components, as the voltage at [1] rises, the voltages at [3] and [4] fall and rise respectively due to the increased conduction of Tr_2 .

The increase in voltage at [4] is effectively transferred to [5], since C_3 relatively is large. This rising voltage at [5] makes $C_{1\&2}$ charge at a more constant rate, rather than progressively reducing which would occur if the linearisation components were omitted and the charging voltage was the 12V supply.

Linearisation is further increased by the integrating network R_1C_1 which provides second order compensation for the nonlinearity of the wave form. By adjusting VR_1 , a near-linear

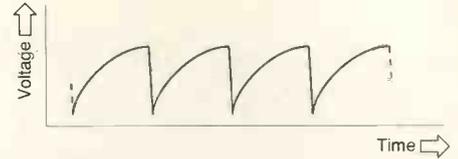


Fig. 2. Waveform obtained using a standard unijunction transistor relaxation oscillator.

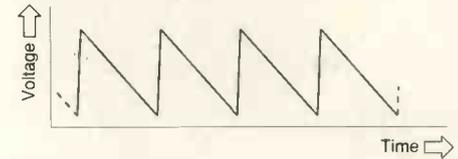


Fig. 3. Enhancing the relaxation oscillator produces a much more linear sawtooth.

sawtooth may be obtained, a positive ramp appearing at [4] and a negative ramp at [3].

A suitable inverting amplifier, IC_{1a} , then inverts, amplifies and shifts the voltage at [4] to a usable level, impedance and amplitude. Voltage at [6] therefore changes as per Fig. 3.

Diode-capacitor pairs D_3C_7 and D_2C_6 find the upper and lower bounds of the voltage at [6] respectively. The upper boundary appears at [8] and the lower at [7]. Op-amps $IC_{1b\&c}$ then act as voltage followers, lowering the impedance of the voltages at [7] and [8]. The voltages appearing at [11] and [12] can therefore be varied between the upper and lower bounds of the sawtooth at [6].

Potentiometer VR_3 is arranged such that when one wiper voltage is at the upper bound the other is at the lower and vice versa.

Op-amps $IC_{2a,b}$ act as Schmitt triggers: for IC_{2a} , when the sawtooth voltage at [6] is greater than the voltage at [11] the output at [13] is high. The same applies for IC_{2b} and points [12] and [14].

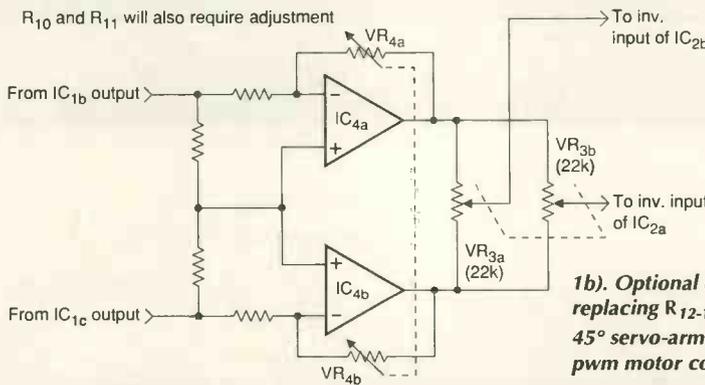
Note that since the thresholds for these schmitt triggers are actually derived from the upper and lower bounds of the sawtooth waveform, very high duty cycles may be achieved at their outputs. This technique offers a significant improvement over the duty cycle attainable using standard monostables. In addition, should the sawtooth voltage at [4] and at [6] change by a small amount in amplitude or level, due to fluctuations in the supply voltage, the performance of the circuit will be maintained. This all helps to fulfil my first and fifth design considerations.

Note that $R_{17\&19}$ were introduced to provide a small amount of positive feedback, and therefore hysteresis in each schmitt trigger, to reduce any change or erratic triggering.

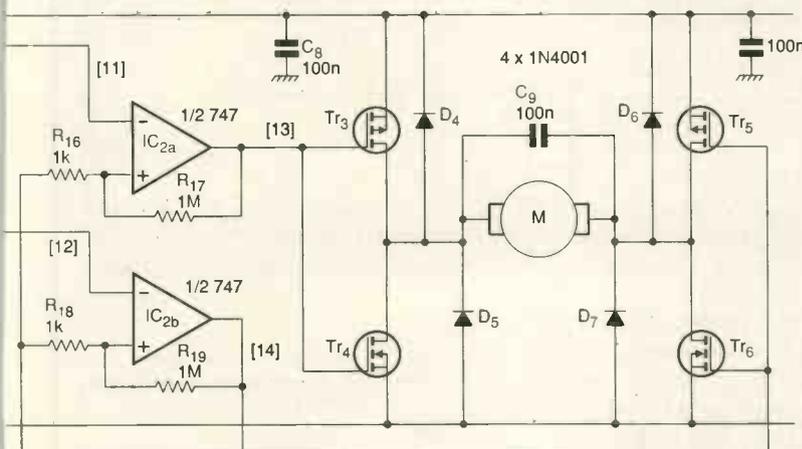
A simple truth table of voltages at points [13] and [14], Table 2, should help you understand the motor drive circuitry around Tr_{3-6} .

Clearly when $Tr_{3\&6}$ conduct the motor will move in one direction and it will move in the opposite direction when $Tr_{4\&5}$ conduct.

Note that D_{4-7} and C_9 are included purely to prevent the mosfets being destroyed by back emf, transient spikes, etc.



1b). Optional circuit extension, replacing R_{12-15} and VR_3 , adds 45° servo-arm facilities to the pwm motor controller.



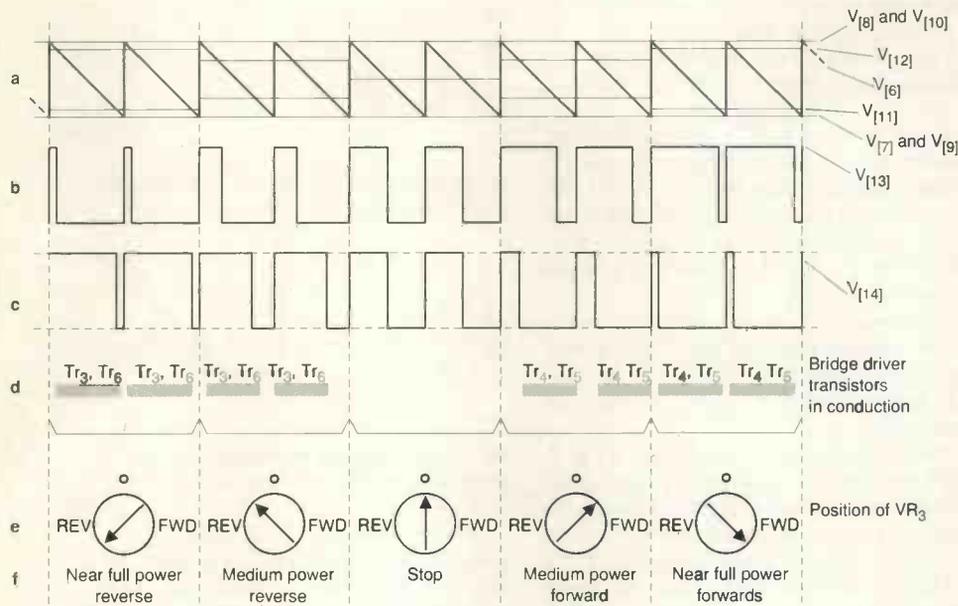


Fig. 4. These timing relationships show how the pwm control sweeps from full-power reverse to full-power forward.

In order to show more detailed operation of the circuit, and to show that VR₃ corresponds to the full control range needed, I have produced a timing diagram/chart, Fig. 5. Sections

d, e, and f are particularly useful in this respect.

Modification for servo actuation

Problems will arise when driving VR₃ from a servo, since the typical angle of swing is +135° for a potentiometer, but only 45° for a servo.

Differential voltage V_[10]-V_[9] therefore needs to be amplified using,

$$\frac{V_{[9]} + V_{[10]}}{2}$$

This is achieved by the modification as in Fig. 1b). Wiper voltages of VR₃ may be varied between the upper and lower bounds of the sawtooth waveform at [6], without using the full range of VR₃'s travel. The amount of travel used is dictated by the position of VR₄, and this can clearly be used to carry out the rimping outlined in the design considerations.

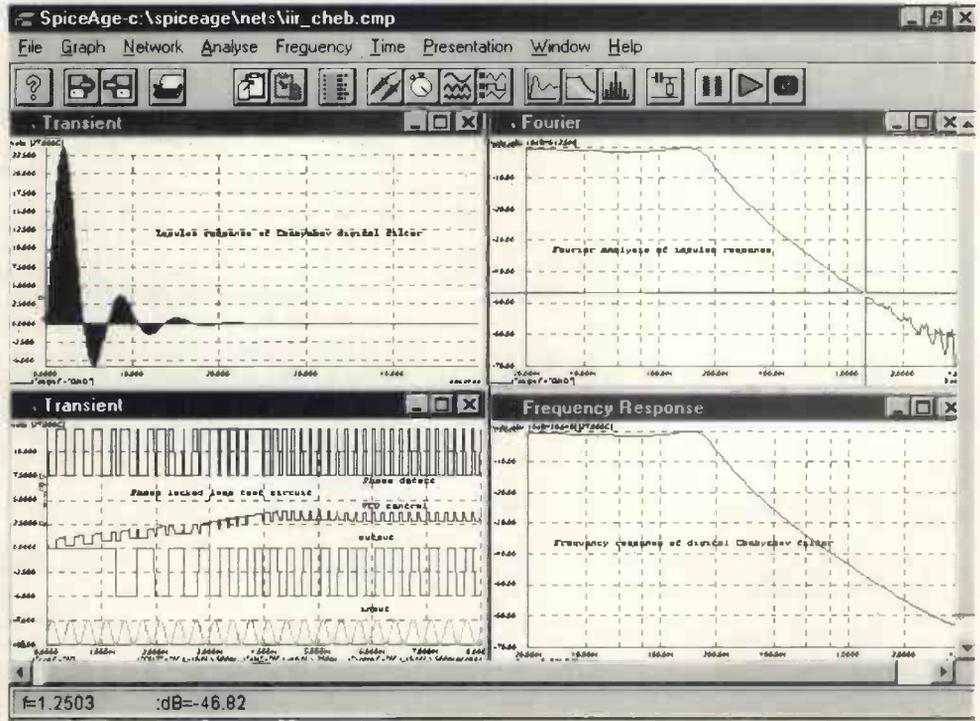
In practice the amplification introduced by IC₃ may need to be so high that the outputs of IC₃ will saturate, in order to provide the correct range of travel for the servo/VR₃.

Under these circumstances R₁₀ or R₁₁ may require adjustment in order to reduce the gain provided by IC_{1a}, allowing V_[9] and V_[10] to lie between the maximum and minimum output voltages of the op-amps IC_{4a} and IC_{4b}. ■

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Gates convert dc to dc

Ian Forster has found that 74AC logic gates – with their high switching speed and low output impedance – make useful building blocks for simple and cheap low-power dc-dc converters.

The 74AC series of logic gates have a number of unique characteristics, including low static power consumption, very high speed and low output impedance. This makes them an useful building block for some simple and inexpensive power conversion circuits as described here.

These power conversion circuits can be considered as high frequency resonant impedance converters. Their Q is determined by the output impedance of the 74AC logic gate and the L/C ratio. This has the effect of multiplying the input 5V swing to a higher level, which is then rectified to produce a stepped up positive or negative – or both – dc supply.

Two basic circuits are shown in Fig. 1(a) and b). Components L₁, C₁ and C₂ are the resonant circuit, with D₁ – or D_{1,2} in the voltage doubling case – providing rectification, with C₃ filtering the output ripple.

Output voltage against load current is shown for both circuits, equivalent to a 1.1kΩ output impedance for the circuit of Fig. 1(a) and 4.5kΩ for Fig. 1(b). Negative voltages can be generated by reversing the diodes.

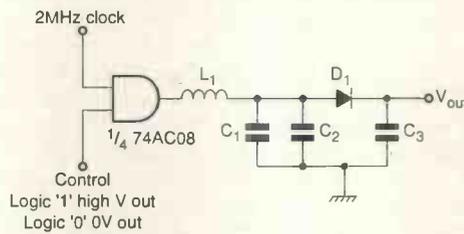
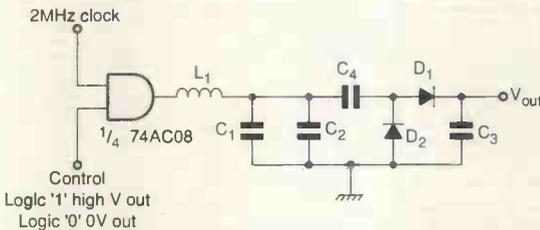
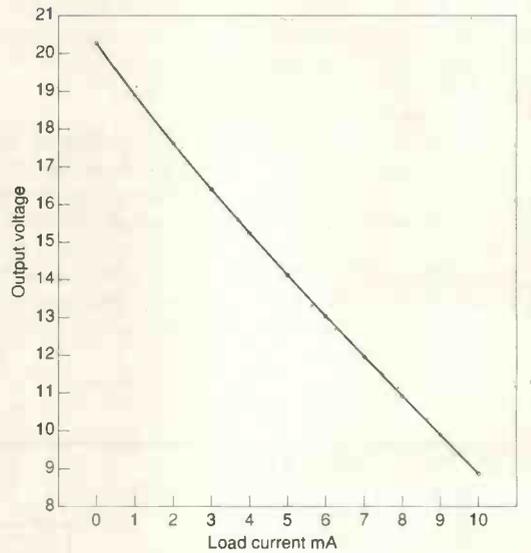
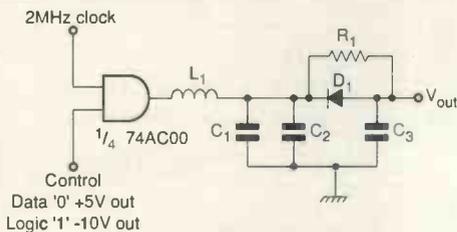
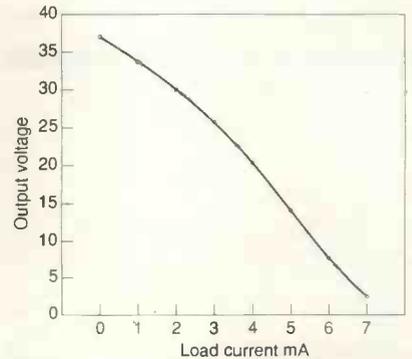


Fig. 1. Basic dc-to-dc converter based on a 74AC logic gate, together with load performance curves. Output impedance of a) is around 1.1kΩ, while that of b) is around 4.5kΩ.

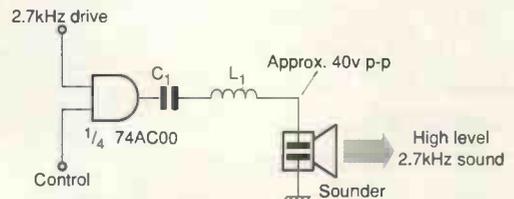


- Component values:
- L₁ 10μH surface mount TOKO 43FS series
 - C₁ 470p ceramic
 - C₂ 120p ceramic
 - C₃ 100n ceramic (see text)
 - C₄ 100n ceramic
 - D₁, D₂ BYV10-40 Schottky diode
 - IC₁ 74AC08



- Component values:
- L₁ 10μH surface mount TOKO 43FS series
 - C₁ 470p ceramic
 - C₂ 120p ceramic
 - C₃ 1n ceramic (see text)
 - R₁ 3k9
 - D₁ BYV10-40 Schottky diode
 - IC₁ 74AC00

Fig. 2. Logic power conversion applications. a) is useful as a serial-port driver while b) can deliver up to 96dBa from a piezo-electric sounder.



- Component values:
- L₁ 220mH TOKO axial
 - C₁ 100n ceramic 100V
 - Sounder - 35mm piezo sound element in Helmholtz resonator

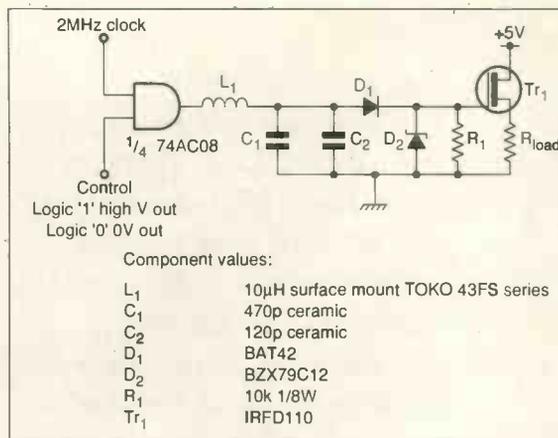


Fig. 3. In a) the 74AC power converter is being used to drive a high-side switch, with a 1.27µs rise time and 3.11µs fall. In b), much higher voltages can be produced - up to ±60V - at the expense of output impedance.

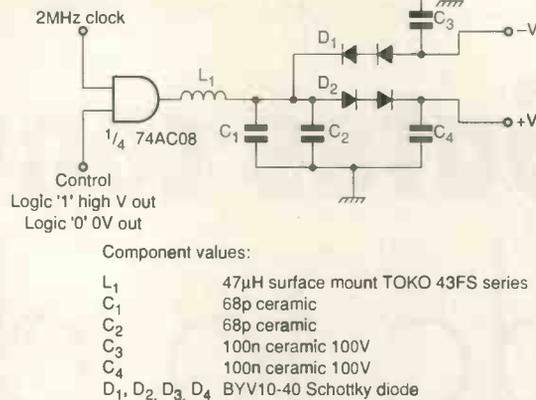


Figure 2a) shows the idea applied to serial data communications. Here a quarter of a 74AC00 nand gate is used to generate either -10V by power conversion or +5V via L_1 and R_1 . The main limitation on speed is the value of C_3 , which includes line capacitance; with the load shown the circuit was able to drive at an equivalent of 9600 baud.

Figure 2b) shows another way the high drive capability of the 74AC gate can be used. Here the tuned circuit is formed by L_1 and the capacitance of the piezoelec-

tric sounder disk. Capacitor C_1 prevents dc being applied to the disc. With the disc mounted in an appropriate Helmholtz resonator, a very high sound level can be produced - in excess of 96dBa.

Delivering up to 60V

Figure 3a) shows a power converter high-side driving an n-channel mosfet as a 5V switch. Mosfets with a p-channel structure tend to have higher on resistance than equivalent n-channel devices and are also more expensive.

In this circuit L_1 , C_1 and C_2 are

the impedance converter with D_1 rectifying the signal. Diode D_2 limits the voltage to avoid damaging the gate. In this example, smoothing of the rectified signal relies on the gate source capacitance, and R_1 provides a discharge path to allow modulation. As shown, with a 10/5Ω load resistor, the on resistance of T_{r1} was measured as 0.54Ω and modulation with a 50kHz square wave at the control input gave a 1.27µs rise and 3.11µs fall time.

Figure 3b) shows a converter using a much higher impedance transformation and hence higher

output voltages. Under no-load conditions $V+$ is 65V and $V-$ is -60V. With a simultaneous 68kΩ load, $V+$ is 43V and $V-$ is -38V.

For interest, I found that it was possible to light a neon lamp with this circuit.

Power conversion circuits using this principle have a number of possible advantages; they are cheap, flexible and, for use in communications circuits, both frequency stable and clock synchronous. This avoids problems associated with jamming of intermediate frequencies. ■

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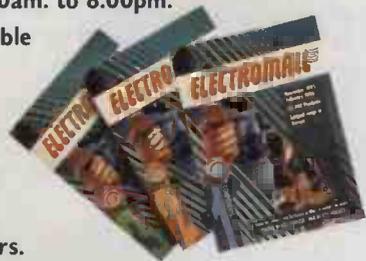
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Low-noise antenna preamp

In D F Conway's low-noise, narrow-band masthead amplifier, input is shared between two mosfets. This eliminates the balun, enhances overload capability and reduces harmonic levels for high-level in-band signals.

This amplifier is specifically designed to be connected to a balanced antenna input for reception of signals over a narrow band of about 2MHz. It was originally designed as a mast-head amplifier for receiving weather satellite transmissions in the range of 137MHz to 138MHz. However, its centre frequency can be varied over a wide range by selecting suitable values for the tuned circuit components.

The amplifier has 28dB gain and a low noise figure. It is powered from a 12-15V supply, fed down the coaxial cable. Mast-head amplifiers are an effective method of enhanc-

Lieutenant Commander D F Conway, ME(Elect), MIEE, MIPENZ, C Eng, psc, RNZN.

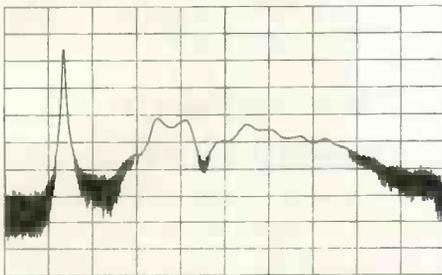


Fig. 1. Output signal amplitude for a signal swept from 0Hz to 1GHz. The hump at about 700MHz is mainly due to self resonance of the dc blocking circuit used to isolate the supply voltage from the spectrum analyser input.

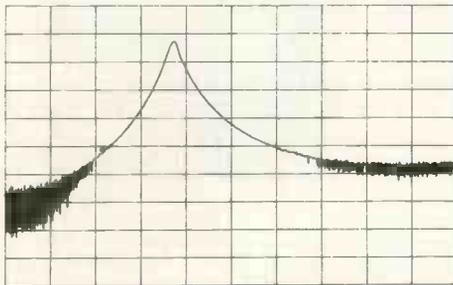


Fig. 2. Output signal amplitude for a signal swept from 100MHz to 200MHz. Frequency response is narrow and shows good roll off.

ing weak signals because they provide a gain stage at the antenna. Any loss that occurs between the antenna and the first gain stage adds directly to the noise figure of the entire antenna/receiver system.

Most single transistor amplifiers have an unbalanced input which requires the insertion of a balun between a dipole antenna and the amplifier. These have insertion losses of about 1dB or more¹ which is greater than the noise contribution of typical low noise amplifiers.

Parallel amplifiers

This amplifier uses two low-noise mosfets as parallel amplifiers to remove the requirement for a balun at the input and improve the dynamic input power range when compared to a single-device amplifier.

Sharing the input signal between two mosfets means that the amplifier will accept a 3dB higher signal before overloading, compared with a single device amplifier. This in turn raises the third-order intercept point by 9dB. The circuit looks similar to a long-tailed pair but the inclusion of by-pass capacitor C_1 decouples the sources of both mosfets to ground.

To ensure even power sharing between the devices, it is important that both signal paths have the same characteristics. Balance is achieved by a symmetrical circuit layout and by the input and output inductors. The transformer action of these inductors compensates for any gain variations between the two mosfets. This is so effective that varying RV_1 over its full range has no observable effect on the amplifier's overall performance.

Implementation details

Correct operation of such a high-gain amplifier depends on good circuit layout and proper shielding. I constructed the amplifier on a double-sided pcb using surface mounted devices to minimise parasitic reactances.

The BF981 device is obsolete and has been repackaged as a BF991. I used the BF981 because the SOT-103 package can be mounted on its 'back' so that the rf signal paths through both mosfets were symmetrical.

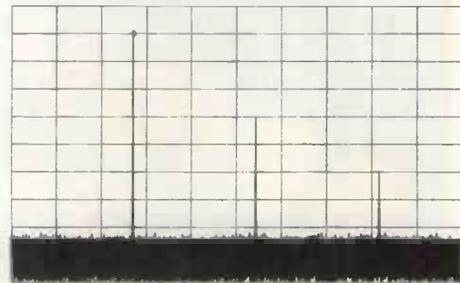


Fig. 3. Output frequency spectrum for a 0dBm input at 137.5MHz. Even at this high input level, the second harmonic is down 30dB while the third is down 50dB. Sweep frequency is 0-500MHz.

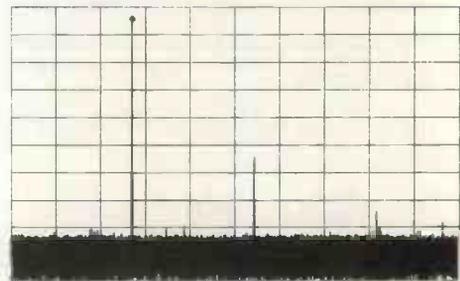


Fig. 4. Output frequency spectrum for a -30dBm input at 137.5MHz. At this lower input level, the second harmonic is down 50dB while the third is down 70dB. Sweep frequency is 0-500MHz.

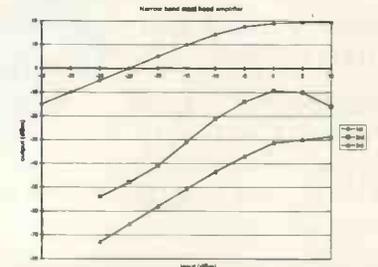


Fig. 5. Plot of input signal level at 137.5MHz versus the output levels of the fundamental, second and third harmonics. The 1dB compression point occurs at about 5dBm while harmonic suppression is good even at high input levels.

Shielding made of sheet steel was used to isolate the input and output signals and prevent oscillation.

Inductors are air cored, space wound using 18swg enamelled wire. The centre tap is a short piece of wire soldered to the centre turn. On the output transformer, the secondary winding is one turn that straddles the centre tap of the primary coil.

The setting up procedure consists of soldering a short link in place of C_3 and measuring the supply current, which should be 10 to 20mA. Replace the link with C_3 and adjust RV_1 so that the supply current is now double the initial reading. Connect the antenna to the amplifier and adjust C_4 and C_5 to achieve maximum amplifier output at the desired input frequency.

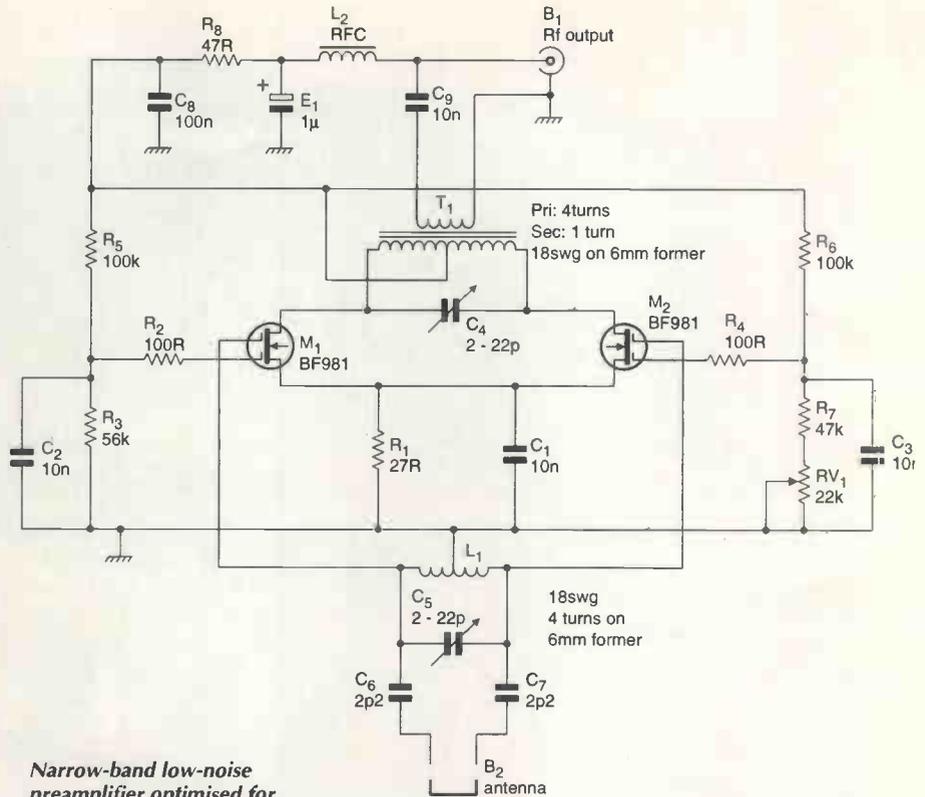
Performance of the amplifier is fully recorded in the plots.

To summarise

This amplifier is suitable for receiving weak signals and is for use with a balanced antenna. Its narrow bandwidth provides good reception in the presence of high level out-of-band interference. As a bonus, using two mosfets to share the load significantly reduces harmonic levels for high-level in-band signals.

Reference

1. Mini-Circuits rf/IF Designers Handbook, '92/'93 Section 14.



Narrow-band low-noise preamplifier optimised for operation at 137.5MHz.

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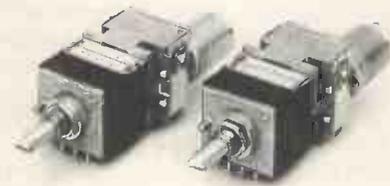
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Relaying transmission line principles

Bill Russell shows how transmission line principles can be explained clearly and simply with the aid of basic measurement equipment and an artificial delay line.

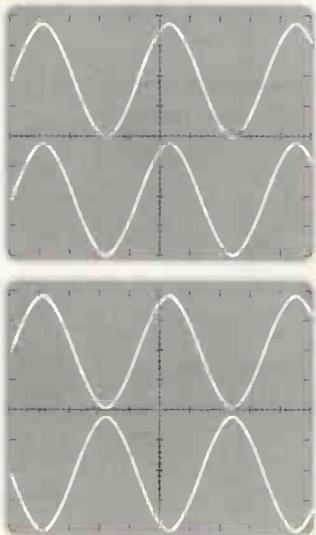


Fig. 3. Signals for a one-wavelength matched line. Output at the load, one wavelength from the input is in phase while at tap 5, half a wavelength from the input, the signal is 180° out of phase. In this and subsequent diagrams, the top waveform in each screen represents input voltage.

The concept of a travelling electromagnetic wave, and the ac voltage and current supporting the wave, has always been a difficult one to present in non-mathematical terms. Even in this age of computer simulation, students remain unconvinced, and ask if there is any simple way that they can establish the basic principles in the laboratory.

In the past, it was usual for most teaching establishments to have a lumped component line, with a delay of 1ms, a Z_0 of around 600Ω, and often with small current-sensing resistors so that progressive voltage and current waveforms could be displayed on an oscilloscope. The real problem – that of simultaneously displaying voltage and current against time and distance along the path of the wave – is not truly solved. However, progressive measurements allow the phase lag and phase velocity to be established. Additionally, current and voltage distribution along the line can be recorded with different terminations.

Unfortunately, the large inductors required for such lines are difficult to accurately fabricate. They also introduce significant losses, which result in discontinuities along the line, heavy attenuation, and a characteristic impedance Z_0 . Although Z_0 may have the correct magnitude, it has a definite reactive component, making such lines unsuitable for simple introductory experiments.

Evaluating the delay line

The discovery of a set of early 1980s computer boards containing 12-pin encapsulated 8μs, 8kΩ delay lines suggested a possible solution. A few simple measurements showed that these lines could be operated into a resistor of 8kΩ at frequencies of 100kHz to 200kHz.

These lines produce negligible standing wave or attenuation. The delay appeared to be generally about 8.4μs. Frequencies of

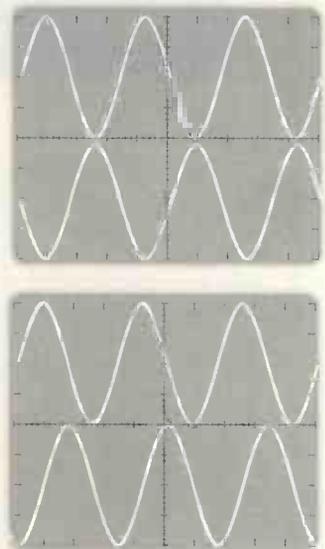


Fig. 4. Ignoring a small capacitive lead, the ratio of input voltage to current gives a nominal resistance equal to that at the termination one wavelength away. This shows that the ratio of line voltage to current in a travelling wave is constant along the length of a matched line. Load is 8kΩ.

Fig. 1. Below, test set-up for the artificial lines. Simple resistive terminations of Z_0 , $2Z_0$ and $0.5Z_0$ are provided for each line.

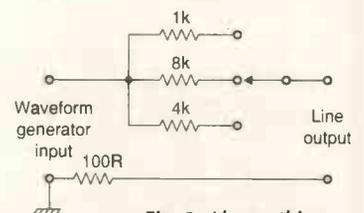
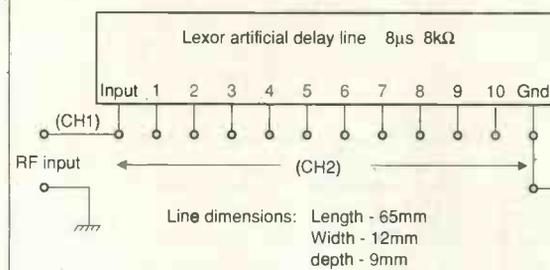


Fig. 2. Above, this ac interface allows 50Ω signal generator output to be matched or mismatched to the line input. The 100Ω resistor in the return line monitors input current when required.

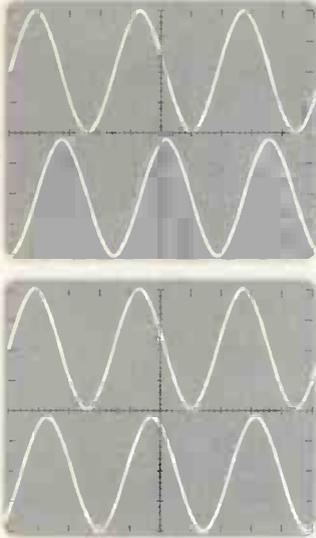


Fig. 5. In the half-wavelength matched line, outputs at the load and at tap 5 are 180° and 90° out of phase respectively.

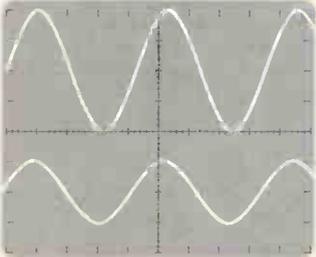
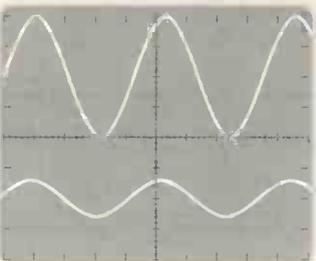


Fig. 6. Half-wavelength matched line outputs at the load, top screen, and at tap 5, bottom screen, are lagging by approximately 90° and 45° respectively. Top waveform in each screen represents input voltage.



approximately 120kHz, 60kHz, and 30kHz allowed the lines to simulate one, half, or quarter wavelengths of uniform line. In addition each delay line had taps at each of the ten sections.

The initial exercise was to calculate the inductance and capacitance per section of the delay line using the simple theory of a lossless line.

Delay per section for $T_d=0.8\mu\text{s}$ is $\sqrt{L_s C_s}$
 Characteristic impedance for $Z_0=8k\Omega$ is $\sqrt{L_s/C_s}$

Note that L_s and C_s are the inductance and capacitance per section. This gives:

$$L_s = T_d Z_0 = 6.4\text{mH} \text{ and } C_s = T_d / Z_0 = 100\text{pF}.$$

A second type of nominal delay, $4\mu\text{s}$, and Z_0 , $4k\Omega$, gave 1.6mH and 100pF. Since these values are easily practicable, discrete lines could be fabricated in place of the encapsulated ones obtained so fortuitously. The simple test panel for the artificial lines is shown in Fig. 1. Simple resistive terminations of Z_0 , $2Z_0$ and $0.5Z_0$ are provided for each line. Open circuit and short circuit options are also available.

Making measurements

Measurements using sine waves are described first, since these can be carried out with a standard 0.2Hz to 2MHz function generator and a 20MHz double beam oscilloscope. An ac interface, Fig. 2, allows the 50Ω output of the signal generator to be matched or mismatched to the input of the line. The 100Ω resistor in the return line monitors input current when required.

Initial measurements are carried out with source and terminating resistors set at 8kΩ. The waveform generator output is adjusted to give a sine wave of 8V peak on the 200 to 200kHz range. Frequency is adjusted around 120kHz so that line input and output voltages on channel 1 and 2 are in phase.

The line is now equivalent to a transmission line of one wavelength and the voltage at each tapping point will establish the progressive phase lag over one wavelength, a slight indication of a standing wave, and a small attenuation over the ten sections. Figure 3 shows the output at the load end, and at the halfway point. This means that, as was to be expected, the artificial line is not entirely loss free, hence Z_0 will have a reactive element.

Channel 2 is now moved to the line end of the 100Ω resistor and the sensitivity increased to display the line input current of the nominally matched line. The ratio V_{in}/I_{in} gives the nominal value of line input impedance, and it will be noted that this is approximately 8kΩ with current leading by a small angle. Figure 4 shows the input voltage and current for the matched one wavelength line.

Readjusting frequency – so that the line behaves as one half, and one quarter wavelength – enables the simple properties of these line lengths to be established. This is provided measurements are carefully recorded and processed. For the purpose of this article, measurements were made using a storage oscilloscope with plotter interface so that a hard copy of the waveforms could be obtained. A selection of these results is shown in Fig. 5 and 6.

Fig. 7. With a wavelength line, a 16kΩ load results in a 2:1 mismatch at the end of the line, and inspection of the voltage at successive taps shows a 2:1 voltage-standing-wave having maxima at each end and in the centre (half-wavelength) and minima at quarter and three-quarter wavelengths from the input. The voltage/current at the input (8V and 0.5mA) gives a nominal input resistance of 16kΩ – identical to that at the load, one wavelength away. Due to the 2:1 standing wave of current, the line voltage/current at the quarter and three-quarter wavelength points will thus be 4V and 1mA, giving a line resistance at these points of 4kΩ. This is more appropriately shown by the measurement on a line of electrical length quarter-wavelength as in Fig.10.

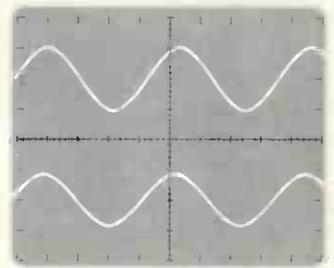


Fig. 8. Loading of 4kΩ also results in a 2:1 mismatch, with a corresponding standing wave. This time minima of 4V appear at each end and in the centre (half-wavelength). Input resistance is again equal to that at the termination i.e. 4kΩ due to the standing wave of voltage and current. Line voltage to current ratio will be a maximum of 16kΩ (8V and 0.5mA) at the quarter- and three-quarter wavelength points, and a minimum of 4kΩ (4V and 1mA) at the half-wavelength point.

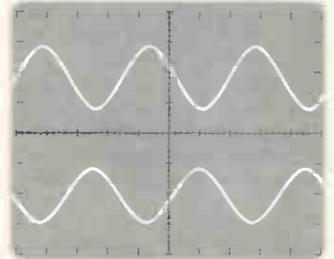


Fig. 9. Due to the standing wave caused by the 2:1 mismatch, the voltages at input and load are minima of 4V but are in antiphase. If monitored, input current would be maximum of 1mA (input resistance 4kΩ). At tap 5 – the quarter-wavelength point – the voltage is a maximum of 8V and lags by 90°. Standing wave current at this point would be 0.5mA (16kΩ).

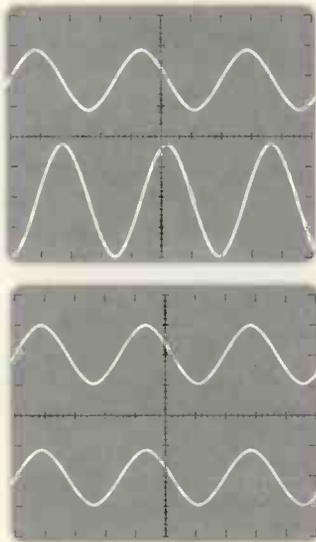


Fig. 10. With a quarter-wave line, 16kΩ loading establishes 2:1 voltage and current standing wave on the quarter-wave line, with a voltage maximum (8V) and a current minimum (0.5mA) at the load end. Voltage minimum (4V) and current maximum (1mA) occur at the input. This is a particular example of the so called 'quarter-wave transformer' in which the ratio of output to input resistance is proportional to the square of the standing-wave ratio, in this case 4:1.

Line properties

In order to establish the properties of a mismatched line, the input is readjusted to the frequency. This makes the line equivalent to one wavelength, and the line is terminated in $2Z_0$ or $Z_0/2$. Monitoring the voltage outputs at successive taps will indicate a 2:1 voltage standing wave on the line in each case with successive maxima and minima spaced at quarter-wave intervals. With a 16kΩ load, voltage maxima occur at each end and in the middle (half-wave point), with minima at the quarter-wave and three-quarter-wave points. With a 4kΩ load, a 2:1 standing-wave ratio is again produced but the relative positions of maxima and minima are interchanged.

Figures 7 and 8 show conditions at the input of these lines and by inference, conditions at successive 14 wave points can be deduced. Figure 9 shows the input, output at the load, and the output at tap 5 for a mismatched line of half-wavelength. Figures 10 and 11 are included to show examples of the traditional quarter-wave matching section.

A second delay line used had characteristics of 4μs and 4kΩ. Connected to the main 8kΩ line and operated at a frequency which makes the 4μs line equivalent to quarter-wave, a 2kΩ load caused a standing wave in the quarter-wave section and acts as a rough match for the main 8kΩ line.

Due to the wide range of measurements provided by this delay line arrangement, the whole exercise is designed to be broken down into a progressive series of laboratory sessions. In this way practical expertise and theoretical understanding gradually increase and are mutually supportive. The use of artificial delay lines to measure the propagation of rectangular pulses will be examined in a further article.

Further reading

1. Ian Hickman, 'RF Reflections', *EW+WW* Oct 1993.
2. D.C.Green, 'Radio Systems Technology', Ch. 3, Transmission Lines.

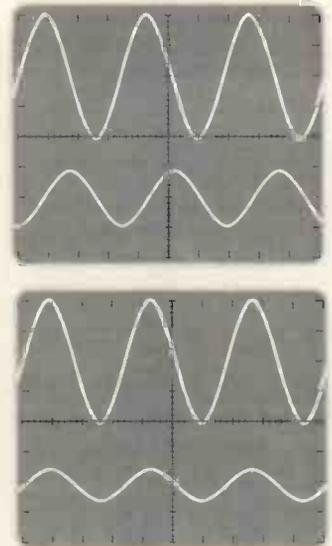


Fig. 11. Again with a quarter-wave line, 4kΩ load establishes a voltage and current standing wave, with a voltage minimum (4V) and current maximum (1mA) at the load, and a voltage maximum (8V) and current minimum (1mA) at the input. In this case, output and input resistances are in the ratio 1:4.

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Cyril Bateman runs through the practicalities of downloading software from the Net - using a working yet almost free demonstration version of PSpice as an example.

Hands-on Internet

Having followed Doug Self's amplifier series, or Owen Bishop's introductory circuit design articles, you might have had a desire to learn PSpice hands on'. Buying the required software can be prohibitively expensive. The cost of downloading an evaluation copy from Internet, however might surprise you.

Over the past year, in the Internet Newsgroups targeted to Electronics and this periodical, interest in using circuit simulation has doubled. An Internet 'Gopher' search against 'PSpice' produced 258 'hits' compared with 143 in February 1995.¹

Fundamental to circuit simulation is the 'device model' used with the simulation software. An op-amp could be described using the exact circuit description, but this would result in very large files and slow simulation, so most semi-conductor makers issue disks of simplified 'MacroModels' extracted using the PSpice 'Parts' software. For most users simulations produced in this way are sufficiently accurate.

Burr-Brown offer a disk - part No AB/E-020F - for the company's op-amp catalogue. It includes four levels of model topologies, 'MacroModels' and three improved levels. These culminate in 'Level IV: Simplified Circuit Models' describing the op-amp at the transistor level for the most accurate simulations. Disks are available on request from the relevant customer service departments for most manufacturers.

This article demonstrates File Transfer Protocol, FTP, using the Internet software included with

OS/2 Warp. Where possible actions needed using Windows or the Mac computer are also indicated - assuming you have access to a local FTP host.

Carrying out the quotes as shown should enable successful transfer of the PSpice software. The indicated costs assume BT costs at 3.5p per minute and Service Provider costs at £3 per hour, using a 14400 baud or faster modem with off-peak transfers typically at 100kbyte/min.¹

The best off-peak times are Saturday and Sunday mornings

while America sleeps, however these times are used for file maintenance in the US (2-5am US local), hence the best times for large file transfers tend to be 10am to 1pm.

Constantly updating

The Internet is always changing, I recently downloaded a copy of PSpice Eval6.1 from,

'ftp.iastate.edu/pub/pc/pspice',

However on preparing this article, I found a message saying PSpice

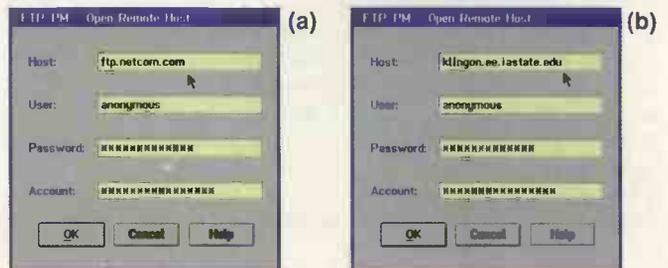


Fig. 1a) Logging on to a remote host, using anonymous ftp. *Microsim.FTP(Windows only) 'ftp.netcom.com'.*
1b) *iastate.edu(dos/Windows/Mac) 'klingen.ee.iastate.edu'.* Hosts automatically respond with the correct drive set to its root directory.

Macro-models for use with PSpice

Analog Devices.	Spice Model Library.
Apex Microtechnology Corp.	Spice Models
Burr-Brown Corp.	Application Note AB/E-020F
Harris Semiconductor.	Analog Products.
Linear Technology Corp.	Spice models.
Texas Instruments Corp.	Mixed Signal Access

Useful Internet sources

Internet Newsgroups	sci.electronics
	sci.electronics.design
	sci.electronics.cad
	sci.electronics.basics
	sci.electronics.equipment
	sci.electronics.components

had moved. An Archie search against 'P Spice' indicated two sources, relocated at 'klingson.ee.iastate.edu' and the new source at Microsim's FTP site 'ftp.netcom.com'.

Microsim's own source at 'ftp.netcom.com' offers the latest Windows evaluation version (6.2) of PSpice. This includes schematic capture, Optimiser and Polaris, in total around 12Mbyte of self extracting zipped files. Its transfer cost is £10.20.

Address 'klingson.ee.iastate.edu' has a slightly older version of 6.2 split into floppy-sized files. These might be easier to download – in total around 14Mbytes – and with a transfer cost of £12.60. To use these Windows versions you must have a suitable 'win32s' file. This is available from either site. Also, a minimum of a 386 with co-processor is needed, together with 8Mbyte of extended memory, Dos 5 and Windows 3.1.

Address 'klingson.ee.iastate.edu' also has Mac and dos evaluation versions of PSpice, including 'pseval5.3' for dos which does not need a co-processor. While the dos versions do not have schematic entry, they have much smaller files, typically 2.5Mbyte costing only £2.25 to download. This is an ideal low-cost start to circuit simulation – especially if your only access to FTP is by 'E-Mail'.

Logging on to the net

Having decided which version of PSpice you want, start up your local FTP host and log-on to the chosen site using 'anonymous' FTP, Figs 1a) and 1b).

Change to the correct 'remote' directory, capture the screen or print this directory for later use, transfer the appropriate text files, save to disk and log-off, Figs 2 and 3.

Read the down-loaded text files; see Table.

Having read the text files, choose

Table. Files on the net relating to PSpice – and where to find them.

Remote host	Microsim.FTP(Windows only)	lastate.edu (Dos/Windows/Mac)
Log-on:	"ftp.netcom.com"	"klingson.ee.iastate.edu"
Remote Directory	"pub/mi/microsim"	"/pub/pspice"
Text File also	"message.txt" "wineval.txt"	(PC only) "pspicftp.txt" (Mac only) "pspicmac.txt"

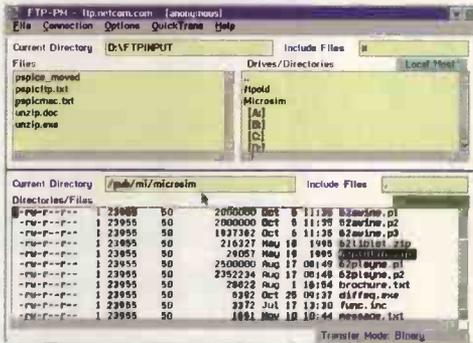


Fig. 2. Logged-on to 'ftp.netcom.com' with remote directory changed to '/pub/mi/microsim'. Change the remote directory by entering '/pub/mi/microsim' into current directory box. Now ready to transfer the highlighted remote file '62pdlb.zip' into the directory. Note the system has been set to binary transfer mode. To initiate transfer, click on 'QuickTrans'.

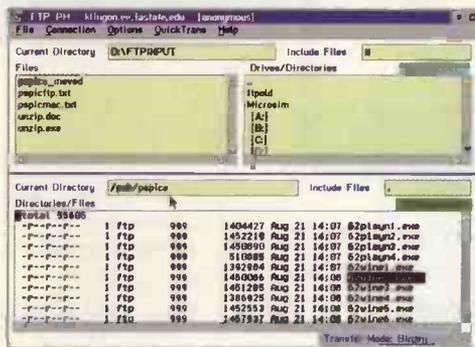


Fig. 3. Logged-on to 'klingson.ee.iastate.edu' with remote directory changed to '/pub/pspice'. Change the remote directory by entering '/pub/pspice' into the current directory box. Now ready to transfer the highlighted remote file '62wine2.exe' into the local directory. Note the system has been set to binary transfer mode. To initiate transfer, click on 'QuickTrans'.

which packages are needed and highlight these on the directory printout. Repeat the above log-on, select Transfer Mode Binary, select the required files and download.

With large file sizes, if for any reason the transfer slows down unduly, wait for 2-3 minutes, or longer if much of the file has already transferred, to see if it restarts and/or regains speed. If not log-off and try for a less busy time

later. Some packages provide auto log-off after a pre-selected inactive period for example five minutes, Fig. 4.

File-Transfer Protocol has two modes of file transfer. Text mode is restricted to printable characters, e.g. text only files, while binary mode transfers the full ascii set and is essential for program or zipped files.

A word of caution. If you are downloading Windows software from the Internet, always ensure you have a valid current backup of your disk before installing the package. From past experience, some sites do not supply the complete configuration routines. The new package works fine, but corrupts the system files, such that none of the old software works. Once more, back-up first so that you can easily restore if needed

Readers wanting more information on requirements for accessing the Internet should see reference 1.

References

1. Surfing with Intent EW&WW June '95 pp. 488-492.
2. Spice – A Guide to Circuit Simulation & Analysis Using PSpice by Paul W.Tuinenga, Prentice Hall ISBN0 13 834607 0.

Limitations of the Pspice evaluation package

'P Spice', by Microsim Corporation, is the industry standard pc simulator and is available in several versions These range in price from £115 for the evaluation system with manuals to £4200 for the standard package.

With the evaluation package, circuit simulation is limited to circuits with up to 64 nodes, 10 transistors, two op-amps, or 65 digital primitive devices, or a combination thereof; 10 ideal transmission lines with not more than 4 non-ideal lines (lossy lines using RLCG parameters), 4 coupled lines; device characterisation limited to diodes; stimulus generation limited to sine waves; sample library of 22 analogue and 140 digital parts. Schematic capture is limited to single A4 page schematics, containing a maximum of 25 symbols, with a reduced symbol and package library set.

Manuals can be purchased from Microsim. Alternatively, should the on-screen 'Help' not suffice and a low cost option is needed, Spice – A Guide to Circuit Simulation & Analysis Using PSpice by PW Tuinenga of Microsim Corporation might well suffice.²

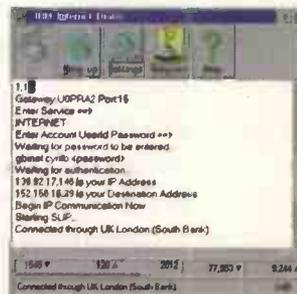
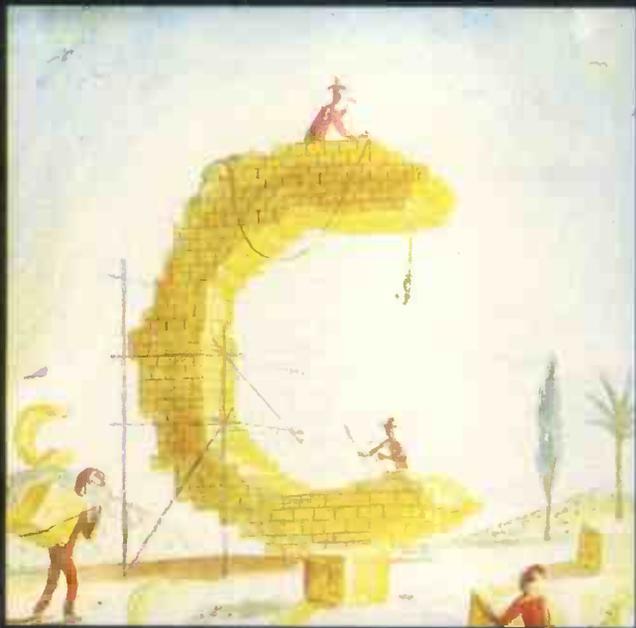


Fig. 4. OS/2 warp internet 'dialler'. Note the horizontal 'speedometer' bar above 'Connected through...' showing the present transfer rate as a percentage of the 'peak' rate attained for this session. Numbers show actual rates in and out, quantity of data transferred and elapsed time 'on-line'. The data in/out rate is continuously updated as prior five seconds averaged, during this session.

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Working with AVALANCHE transistors

On avalanching, a transistor can be made to produce extremely fast, high-current pulses. Ian Hickman has been looking at how ordinary transistors avalanche and at a couple of extremely rare devices specified for avalanche operation.

I have been fascinated by avalanche transistor circuits ever since I first encountered them in the early 1960s. They have probably been known since the earliest days of silicon transistors. I have never heard of them being implemented in germanium.

One important use for avalanche transistors was in creating extremely fast, narrow pulses to drive the sampling gate in a sampling oscilloscope. Such oscilloscopes provided, in the late 1950s, the then incredible bandwidth of 2GHz. At that time, other oscilloscopes were struggling, with distributed amplifiers and special cathode ray tubes, to achieve a bandwidth of 85MHz.

Admittedly, those early sampling oscilloscopes were plagued by aliased responses and, inconveniently, needed a separate external trigger. But they were steadily developed over the years, providing, by the 1970s, a bandwidth of 10-14GHz.

The latest digital sampling oscilloscopes provide bandwidths of up to 50GHz, although like their analogue predecessors they are limited to displaying repetitive waveforms, making them inappropriate for some of the more difficult oscilloscope applications, such as glitch capture.

The basic avalanche transistor circuit is very simple. A version published in the late 1970s apparently produced a 1Mpulse/s pulse train with a peak amplitude of 11V, a half-amplitude pulse width of 250ps and a 130ps rise time¹. It achieved this with a 2N2369 – an unremarkable switching transistor with a 500MHz f_t and a C_{obo} of 4pF.

The waveform, reproduced in the article, was naturally captured on a sampling oscilloscope.

wavered a little after the 1970s. Perhaps this was due to the fact that the limited number of specialised uses for which the devices are appropriate resulted in the spotlight always resting elsewhere.

A problem with designing an avalanche transistor circuit is the absence of transistor types specifically designed and characterised for this application. But this situation has recently changed, due to the interest in high power laser diodes capable of producing extremely narrow pulses. Applications for such lasers include range finding, Pockel cell drivers and streak cameras.

Two transistors specifically characterised for avalanche pulse operation, types ZTX413 and ZTX415, have recently appeared², together with an application note³ for the latter.

The avalanche transistor depends for its operation on the negative resistance characteristic at the collector. When the collector voltage exceeds a certain level, somewhere between V_{ceo} and V_{cbo} , depending on the circuit configuration, the voltage gradient in the collector region exceeds the sustainable field strength, and hole-electron pairs are liberated. These are accelerated by the field, liberating others in their turn. As a result, the current

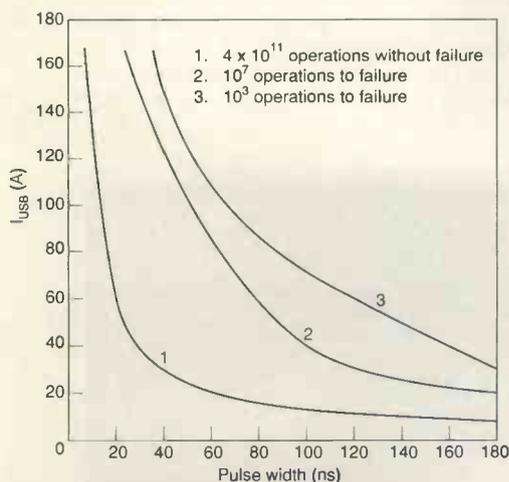


Fig. 1. Maximum permitted avalanche current versus pulse width for the ZTX415, for the specified reliability.

The avalanche circuit revisited
Interest in avalanche circuits seems to have

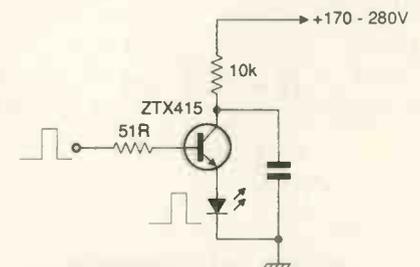


Fig. 2. Simple high current avalanche pulse generator circuit, driving a laser diode.

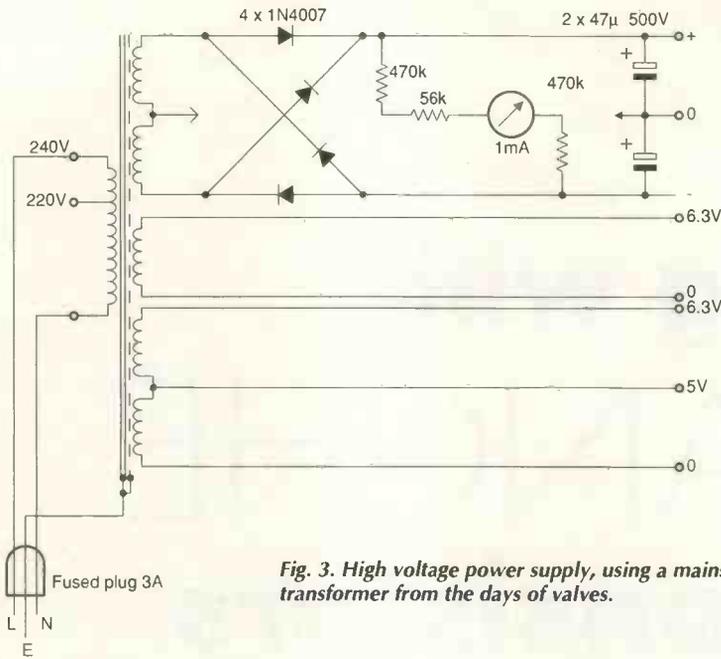


Fig. 3. High voltage power supply, using a mains transformer from the days of valves.

rises rapidly, even though the voltage across the device is falling.

The resultant 'plasma' of carriers results in the device becoming almost a short circuit, and it will be destroyed if the available energy is not limited. If the current in the avalanche mode, I_{USB} , and the time for which it is allowed to flow are controlled, then reliable operation of the device can be ensured, as indicated in Fig. 1 for the ZTX415.

From the diagram, you can see that for 50ns wide pulses, a pulse current of 20A can be passed for an indefinite number of pulses without device failure. This is provided of course that the duty cycle is kept low enough to remain well within the device's 680mW allowable average total power dissipation P_{tot} .

Figure 2 shows a simple high-current avalanche pulse generator, providing positive going pulses to drive a laser diode. Peak current is determined by the effective resistance of the transistor in avalanche breakdown plus the slope resistance of the diode.

As both the preceding parameters are current dependent, it is not easy to determine accurately just what the peak value of current is. In practice however, this is not an insuperable

difficulty. Energy dissipated in the transistor and diode is simply equal to the energy stored in the capacitor. Since, given the value of the capacitor and the supply voltage, the stored charge is known, the pulse width can be measured and the peak current estimated.

If, in a particular circuit, the avalanche- and diode-slope resistances are unusually low, the peak current will be higher than otherwise, but the pulse width correspondingly narrower. Charge passed by the transistor is limited to that originally stored in the capacitor at the applied supply voltage.

Implementing the avalanche device

Having obtained samples of the ZTX415, I decided to investigate the performance in a variant of Fig. 2. This variant provides negative-going pulses, but substitutes a resistive load for the diode to allow quantitative measurements to be recorded.

First I produced a high-voltage source, giving up to 800V off-load. A voltmeter was included, and for versatility, the transformer's low voltage windings were also brought out to the front panel, Fig. 3.

The test set up is shown in Fig. 4a), the

high-voltage supply being adjusted as required by the simple expedient of running the power supply of Fig. 3 from a variable voltage transformer, of the type commonly known under its trade name of Variac.

With the low value of resistance between the base and emitter of the avalanche transistor, the breakdown voltage will be much the same as B_{VCEs} , the collector-emitter breakdown voltage with the base-emitter junction short circuit. With no trigger pulses applied, the high-voltage supply was increased until pulses were produced. With the applied high voltage barely in excess of B_{VCEs} , the pulse-repetition frequency, prf, was low and the period erratic, as was to be expected.

With the voltage raised further, the prf increased, the free running rate being determined by the timeconstant of the collector resistor and the 2nF capacitor. This free running mode of operation is not generally useful, there being always a certain amount of jitter on the pulses due to the statistical nature of the exact voltage at which breakdown occurs. The high voltage supply was therefore reduced to the point where the circuit did not free run, and a 10kHz squarewave trigger waveform applied.

The pulses were now initiated by the positive edges of the squarewave, differentiated by the 68pF capacitor and the base resistor, at a prf of 10kpulse/s. On firing, the collector voltage drops to near zero. This causes a negative-going pulse to appear across the load resistor, which consisted of a 47Ω resistor in parallel with a 50Ω load. The latter comprised two 10dB pads in series with a 50Ω 'through termination' (RS type 456-150). It was mounted at the oscilloscope's Channel 1 input socket and connected to the test circuit by half a metre of low-loss 50Ω coaxial cable. The cable presented a further 50Ω resistive load in parallel with the 47Ω resistor.

Performance observations

Drop in collector voltage can be seen to be almost the full 250V of the supply, Fig. 4 b), lower trace. However, the peak voltage across the load resistor – upper trace – is only around -180V. This circuit provides a negative-going output, unlike that of Fig. 2.

The lower amplitude of the output pulse was ascribed to the esr – equivalent series resis-

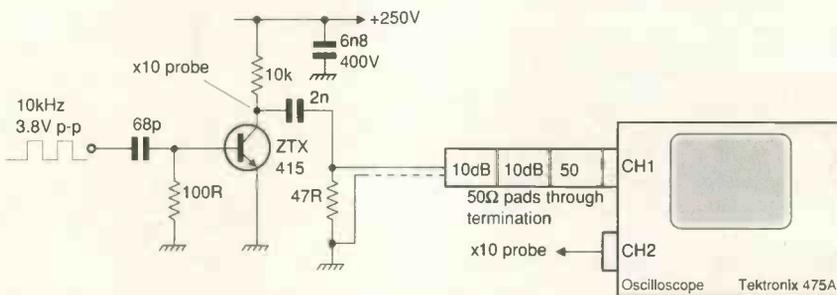
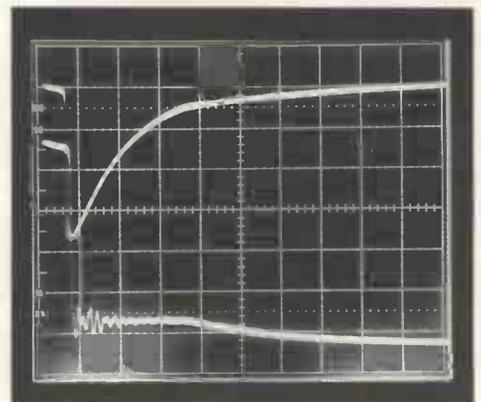


Fig. 4. a) Test set-up used to view the pulse produced by an avalanche transistor.

b) Upper trace, voltage across load, effectively 50V/div (allowing for 20dB pad), 0V = 1cm down from top of graticule, 50ns/div; lower trace, collector voltage, effectively 50V/div (allowing for x10 probe), 0V = 1cm up from bottom, 50ns/div.



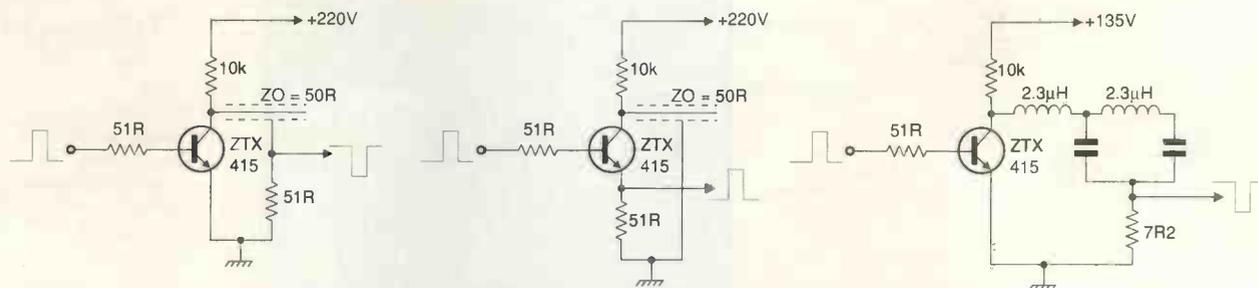


Fig. 5. Circuits producing square output pulses; a), negative-going output pulses and b), positive-going pulses both using coaxial lines; c), negative-going pulses using a lumped component delay line.

tance – of the 2nF capacitor, which was a foil type, not specifically designed for pulse operation. This is confirmed by the shape of the pulse. Its decay is slower than would be expected from the 50ns timeconstant of the capacitor and the 25Ω load, plus transistor slope resistance in avalanche breakdown. This emphasises the care needed in component selection when designing fast laser diode circuits.

Peak pulse voltage across load corresponds to a peak current of 7.25A and a peak power of 1.3kW. However, the energy per pulse is only $\frac{1}{2}CV^2$, where C is 2nF and V is 250V, namely some 63µJ, including the losses in capacitor esr and in the transistor. This represents a mean power of 630mW, most of which will be equally divided between the 47Ω resistor and the first of the two 10dB pads, which is why the prf was restricted to a modest 10kHz.

In Fig. 4b), the lower trace shows the drop across the transistor during the pulse to be about 16V, giving an effective device resistance in the avalanche mode of 16/7.25 or about 2.2Ω. Thus, given a more suitable choice of 2nF capacitor, over 90% of the available pulse energy would be delivered to the load.

In Fig. 2, though, the laser diode slope resistance would probably be less than 25Ω, resulting in a higher peak current, and an increased fraction of the energy lost in the transistor.

Ringing on the lower collector trace in Fig. 4b) is due to the ground lead of the ×10 probe; it could be almost entirely avoided by more careful grounding of the probe head to the circuit. As it also caused some ringing on the upper output-pulse trace, the probe was disconnected when the upper trace was recorded, Fig. 4b) being a double exposure with the two traces recorded separately.

At present, I cannot explain the negative underswing of the collector voltage, starting 200ns after the start of the pulse, before the collector voltage starts to recharge towards +250V.

Squaring the output

The shape of the output pulse from circuits such as Figs 2 and 4a), a step function followed immediately by an exponential decay, is not ideal: for many applications, a square pulse would be preferred. This is simply

arranged by using an open-circuit delay line, in place of a capacitor, as the energy storage element.

When the avalanche transistor fires, its collector sees a generator with an internal impedance equal to the characteristic impedance of the line. Energy starts to be drawn from the line, which becomes empty after a period equal to twice the signal propagation time along the length of the line, as described in Ref. 4.

Figure 5 shows three such circuits, a) and c) producing negative-going pulses and b) positive going. If a long length of line is used,

to produce a wide pulse, then version b) is preferable to a), since it has the output of the coaxial cable earthed. In a), the pulse appears on the outer of the cable, so the capacitance to ground of the outer – which could be considerable – appears across the load.

If a wide negative-going pulse is desired, then an artificial line using lumped components as in c) can be used. Here, the lumped delay line can be kept compact to minimise its capacitance to ground.

Where exceptional pulse power is required, ZTX415 avalanche transistors can be used in series to provide higher pulse voltages as in

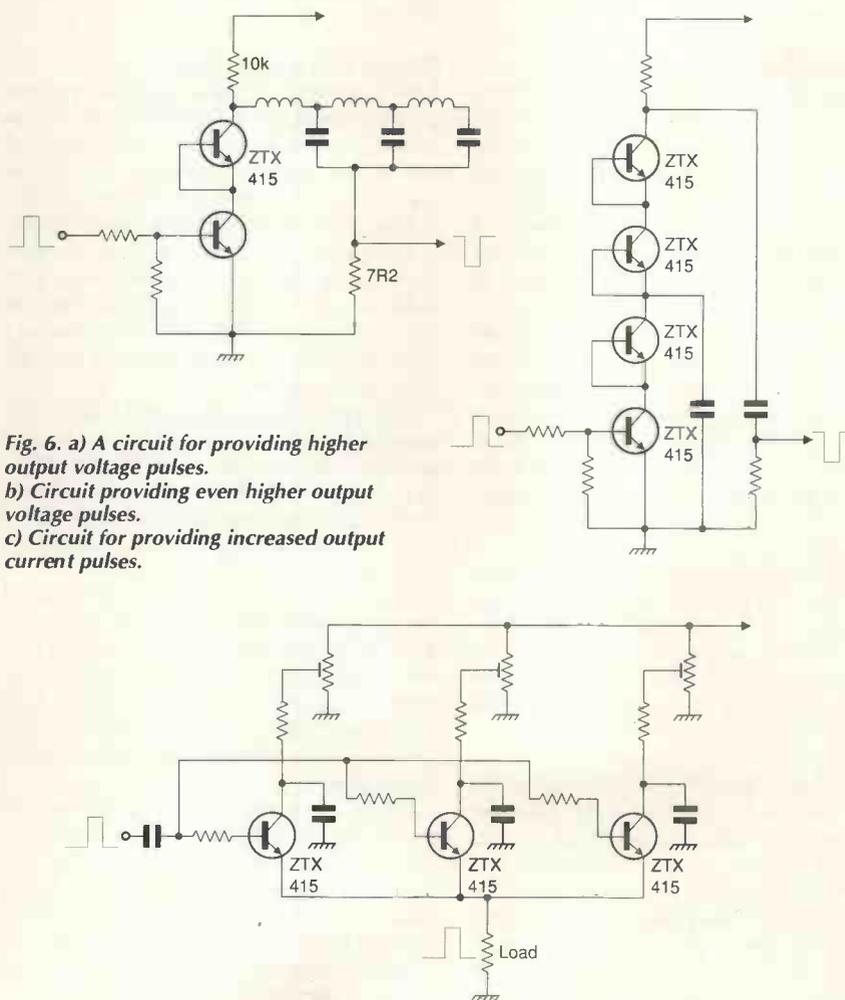


Fig. 6. a) A circuit for providing higher output voltage pulses. b) Circuit providing even higher output voltage pulses. c) Circuit for providing increased output current pulses.

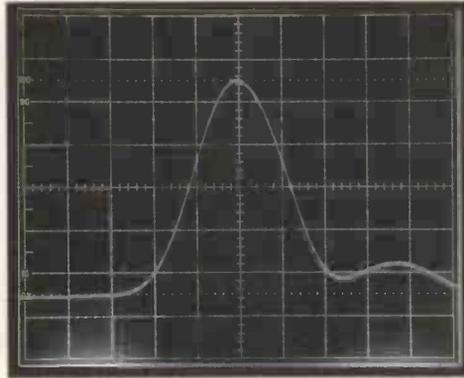
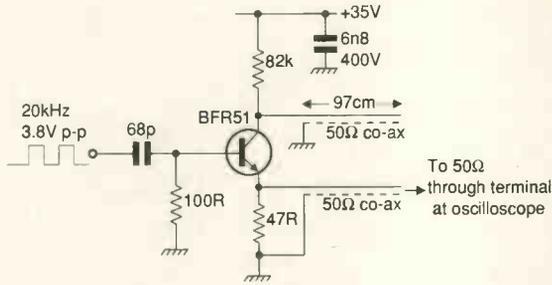


Fig. 7a) Circuit of an avalanche pulse generator using a BFR91 transistor with a 97cm line length. b) Output of a); upper trace, output pulse, 10ns/div, 1V/div, 0V = centre line; lower trace, collector voltage, 20µs/div, 10V/div, 0V = bottom line

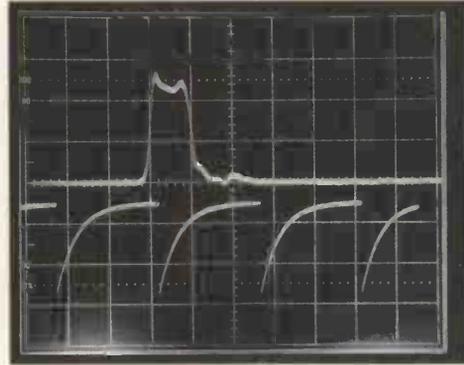
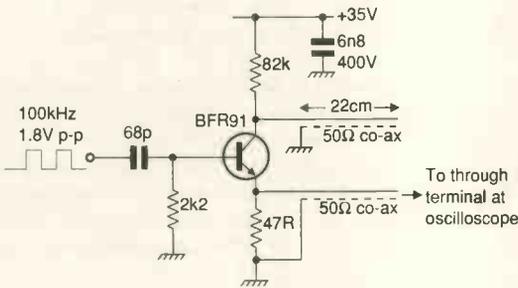


Fig. 8a) Circuit of an avalanche pulse generator using a BFR91 transistor with a 22cm line length. b) Output of a); output pulse, at 1ns/div, >1V/div, indicated rise time 1.5ns.

Figs 6a) and b). Alternatively, they can be used in parallel to provide higher pulse currents as in c).

A high speed version

Rise time of the negative-going edge of the output pulse in Fig. 4b) was measured as 3.5ns, or 3.2ns, corrected for the effect of the 1.4ns rise time of the oscilloscope. This is a speed of operation that might not have expected from a transistor with a minimum f_t of 40MHz and a maximum C_{ob} of 8pF, but this emphasises the peculiar nature of avalanche operation of a transistor.

An obvious question was; could a substantially faster pulse be obtained with a higher frequency device? Low-power switching transistors, being no longer common in these days of logic ICs, the obvious alternative is an rf transistor, which will have a high f_t and a low value of C_{ob} . I therefore decided to experiment with a BFR91, a device with a V_{CEO} rating of 12V and an f_t of 5GHz.

I built the circuit of Fig. 7a) using a length of miniature 50Ω coaxial cable, cut at random from a large reel. It turned out to be 97cm. Given that the propagation velocity in the cable is about 2/3 the speed of light, the cable represents a delay of 4.85ns and so should provide a pulse of twice this length or, in round figures, 10ns.

In the upper trace, Fig. 7b) shows that the circuit produced a pulse of width 10ns and amplitude 5V peak, into a 25Ω load, delivering some 200mA current. Oscilloscope settings were 10ns/div, 2V/div with a centre line of 0V. The lower trace shows – again using a double exposure – the collector voltage at 20µs/div, 10V/div and 0V at the bottom of the graticule. With circuit values shown, at

the 20kHz prf rate used, the line voltage has time to recharge virtually right up to the 35V supply.

Effects of a shorter line

I repeated the experiment, this time with the circuit of Fig. 8a). Line length was reduced to 22cm, some other component values changed and the prf raised to 100kHz. The output pulse is shown in 8b), at 1ns/div horizontal and more than 1V/div vertical, the VARIABLE Y sensitivity control being brought into play to permit the measurement of the 10% to 90% risetime. This is indicated as 1.5ns, but the maker's risetime specification for a Tektronix 475A oscilloscope, estimated from the 3dB bandwidth, is 1.4ns.

Rise times add rms-wise, so if you were to accept these figures as gospel, it would imply an actual pulse rise time of a little over 500ps. In fact, the margin for error when an experimental result depends upon the difference of two nearly equal quantities is well known to be large.

When the quantities must be differenced rms-wise rather than directly, the margin of error is even greater. As a result, no quantitative certainty of the rise time in this case is possible, other than that it is probably well under 1ns. Unfortunately, a sampling oscilloscope does not feature among my collection of test gear.

This raises the intriguing possibility that this simple pulse generator might be suitable as the sample pulse generator in a sampling add-on for any ordinary oscilloscope, extending its bandwidth for repetitive signals to several hundred megahertz, or even a gigahertz.

For this application, it is important that the sample pulse generator can be successfully

run over a range of repetition frequencies. With an exponential approach to the supply voltage at the firing instant, there is the possibility of jitter being introduced onto its timing, due to just how close to the supply voltage the collector has had time to recharge, Fig. 7b), lower trace.

The way round this is to use a lower value of collector resistance returned to a higher supply voltage. This ensures a rapid recharge, but the midpoint of the resistor is taken to a catching diode returned to the appropriate voltage just below the breakdown voltage. The collector voltage is thus clamped at a constant voltage prior to triggering, whatever the repetition rate.

References

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2. 'NPN Silicon Planar Avalanche Transistor ZTX413' Provisional Data Sheet' Issue 2, March 1994 and 'NPN Silicon Planar Avalanche Transistor ZTX415 Data Sheet' Issue 4, Nov 1995.
3. 'The ZTX415 Avalanche Transistor', Zetex plc, Apr 1994.
4. Hickman, I, 'RF Reflections', *EW+WW*, Oct 1993, pp. 872-876.

Figures 1, 2, 5 and 6 are reproduced courtesy of Zetex plc.

Putting the power back in

Design criteria for battery charger circuits are dictated by cell type, the application and operating conditions. In this extract from *Simplified design of micropower and battery circuits* John Lenk examines the characteristics of popular IC controllers.

The first step in designing charging circuits must be to look in detail at an IC that provides fast-charging for popular NiCd/NiMH batteries. Maxim MAX712/713 controllers can fast-charge batteries from a dc source at least 1V higher than the maximum battery voltage.

One to sixteen series cells can be charged at rates between one-third of the battery capacity, i.e. C/3 and four times capacity, or 4C. A voltage-slope-detecting a-to-d converter, timer, and temperature-window comparator determine charge completion.

The ICs (Figs. 1, 2, 3 and 4) are powered by an on-board +5V shunt regulator and draw a maximum of 5µA from the battery when not charging. A low-side current-sense resistor allows the battery-charge current to be regulated while still supplying power to the load.

The MAX712 terminates fast-charge by

detecting zero voltage-slope while MAX 713 uses a negative voltage-slope detection scheme. Both ICs are available in sixteen pin DIP and SO packages. An external power p-n-p transistor, blocking diode, three resistors, and three capacitors are the only required external components.

For high-power charging requirements, the ICs can be configured as a switch-mode battery charger that minimises power dissipation.

Basic operating principles

The ICs provide charging by forcing a constant current into the battery in one of two operating states: fast-charge or trickle-charge. During fast-charge, the current level is high, and once full-charge is detected, the current reduces to the trickle-charge state. The ICs monitor three variables to determine when the battery reaches full charge: voltage slope, bat-

tery temperature and charge time.

Full-charge state are determined by the IC's timer, voltage-slope (V) detector, and temperature comparators, and its voltage and current regulator controls output voltage and current, and senses battery presence.

In a typical battery-charging sequence, Fig. 5, when the batteries are already inserted before application of power, initially the IC draws negligible power from the battery.

When power is applied to DC IN, the power-on reset holds the IC in trickle-charge. Once the power-on reset signal goes high, the IC enters the fast-charge state as long as the cell voltage is above the under voltage-lockout (uvlo) voltage of 0.4V per cell. Fast-charging cannot start until the battery voltage divided by number of cells exceeds 0.4V.

As soon as the cell voltage-slope becomes negative, the fast-charge is terminated and the IC reverts to the trickle-charge state. When power is removed, the device draws negligible current from the battery.

Temperature may be used to control charging too, Fig. 6. The ICs can be configured so that either voltage-slope or temperature detects full-charge.

On a cold day, the battery pack may be too cold for fast charging. During the initial peri-

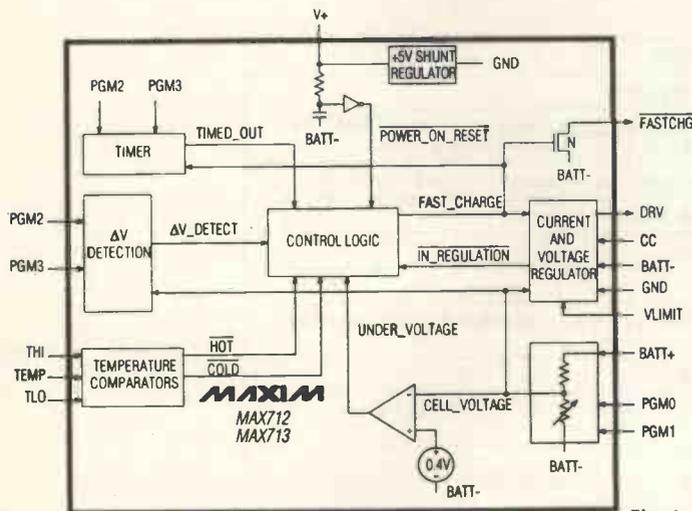


Fig. 1. MAX712/713 used as a fast-charge controller.

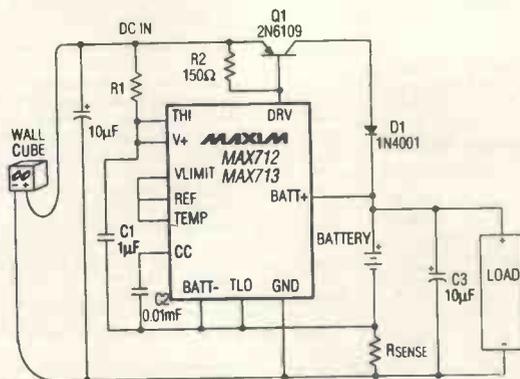


Fig. 2. Typical operating circuit for fast-charge controller. This circuit benefits from a 22kΩ resistor between BATT- and TEMP and 68k between VLIMIT/REF and TEMP. A 10nF capacitor over R2 may also be necessary.

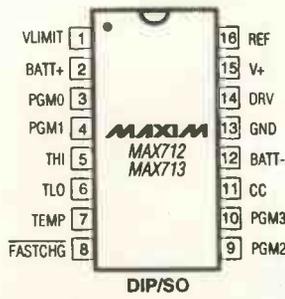


Fig. 3. Pin connections for fast charging with a MAX712/713.

of the ICs will remain in the trickle-charge state, and when the battery temperature exceeds the limit set by THI (pin 5), the ICs will again revert to trickle-charge.

It is possible that the battery will be inserted into an already powered-up IC (Fig. 7). To begin with, the charger output voltage will be regulated at the number of cells times V_{LIMIT} (pin 1) and the IC will be in the trickle-charge state. But on insertion of the battery, the IC will detect current flow into the battery and switch to fast-charge.

Once full-charge is detected, the IC will revert to trickle-charge. If the battery is removed, the IC will remain in trickle-charge, and the output will once again be regulated.

Powering the ICs

The ICs are inactive with the wall cube unplugged, Fig. 2, drawing a maximum of 5µA from the battery. Diode D_1 prevents the collector-base junction of Tr_1 from conducting

current into the DRV (pin 14). With the wall cube connected, R_1 charges C_1 . When C_1 charges to +5V, the internal shunt regulator sinks current to regulate $V+$ to 5V, and fast-charge commences. The IC remains in fast-charge until one of the three fast-charge termination conditions is triggered.

If DC IN exceeds 20V, a cascode connection can be added in series with the DRV pin Fig. 8. This will prevent the DRV pin absolute-maximum ratings from being exceeded. Resistor R_1 in Fig. 8 is selected to pass 5mA at the minimum DC IN voltage.

Total power dissipation of the IC must not exceed the absolute maximum specifications for the device, Table 1. The difference between maximum DC IN voltage and minimum DC IN voltage determines power dissipation in the IC.

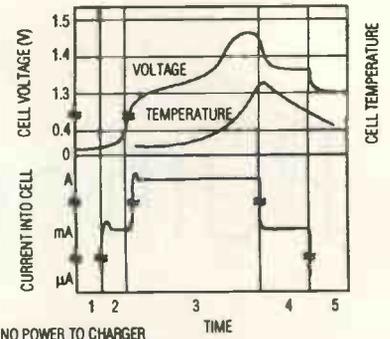
Maximum current into $V+$ is maximum DC IN voltage - 5V divided by R_1 . Dissipation, due to the shunt regulator, is five times the maximum current into $V+$. Sink current into the DRV pin also causes power dissipation.

Fast charging

The IC enters the fast-charge state under one of the following conditions:

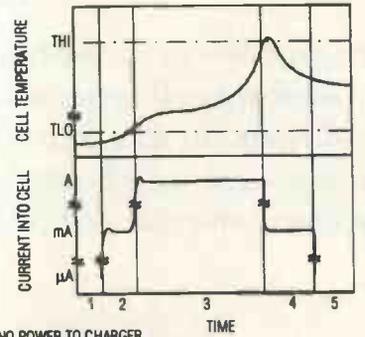
Upon application of power with battery-current detection (GND, pin 13) voltage less than BATT-, pin 12; voltage and TEMP, pin 7, higher than TLO, pin 6; and the cell voltage higher than the WLO voltage.

Upon insertion of a battery with TEMP, pin 7, higher than TLO, pin 6, and lower than



1. NO POWER TO CHARGER
2. CELL VOLTAGE LESS THAN 0.4V
3. FAST CHARGE
4. TRICKLE CHARGE
5. CHARGER POWER REMOVED

Fig. 5. Charging sequence when batteries are already inserted. Once the power-on reset signal goes high, the IC enters the fast-charge state.



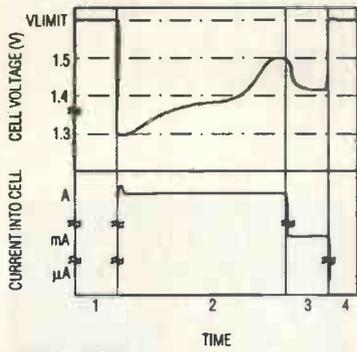
1. NO POWER TO CHARGER
2. CELL TEMPERATURE TOO LOW
3. FAST CHARGE
4. TRICKLE CHARGE
5. CHARGER POWER REMOVED

Fig. 6. Temperature can be used to detect a full charge.

PIN	NAME	FUNCTION
1	VLIMIT	Sets the maximum cell voltage. If VLIMIT is tied to $V+$, the battery terminal voltage ($BATT+ - BATT-$) will not exceed $1.85V \times$ (number of cells); otherwise, it will not exceed VLIMIT \times (number of cells). Do not allow VLIMIT to exceed +2.5V, unless tied to $V+$.
2	BATT+	Positive terminal of battery
3, 4	PGM0, PGM1	PGM0 and PGM1 set the number of series cells to be charged. The number of cells can be set from 1 to 16 by connecting PGM0 and PGM1 to any of $V+$, REF, or BATT-, or by leaving the pin open (see Table 2).
5	THI	Trip point for the over-temperature comparator. If the voltage on TEMP rises above THI, fast charge ends.
6	TLO	Trip point for the under-temperature comparator. If the MAX712/MAX713 powers on with the voltage on TEMP less than TLO, fast charge is inhibited and will not start until TEMP rises above TLO. TLO must be set below the minimum operating temperature of the charger.

PIN	NAME	FUNCTION
7	TEMP	Sense input for temperature-dependent voltage from thermistors
8	FASTCHG	Open-drain fast-charge status output. While the MAX712/MAX713 fast charges the battery, FASTCHG sinks current. When charge ends and trickle charge begins, FASTCHG stops sinking current.
9, 10	PGM2, PGM3	PGM2 and PGM3 set the maximum time allowed for fast charging. Timeouts from 33 minutes to 264 minutes can be set by connecting to any of $V+$, REF, or BATT-, or by leaving the pin open (see Table 3). PGM3 also sets the fast-charge to trickle-charge current ratio (see Table 5).
11	CC	Compensation input for constant current regulation loop
12	BATT-	Negative terminal of battery
13	GND	System ground. The resistor placed between BATT- and GND is used to monitor the current into the battery.
14	DRV	Current sink for driving the external PNP current source
15	$V+$	Shunt regulator. The voltage on $V+$ is regulated to +5V with respect to BATT-, and the shunt current powers the MAX712/MAX713.
16	REF	2.0V reference output. Sources up to 1mA.

Fig. 4. Pin descriptions for a fast-charge controller.



1. BATTERY NOT INSERTED
2. FAST CHARGE
3. TRICKLE CHARGE
4. BATTERY REMOVED

Fig. 7. Charging sequence when batteries are inserted in powered-up IC. When the battery is inserted, the IC detects the current flow and switched to fast charge.

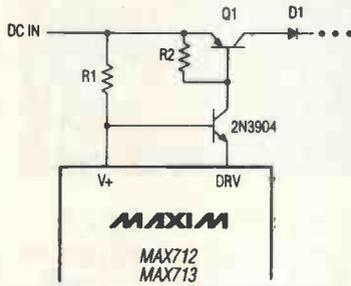


Fig. 8. DRV pin cascode connections where DC IN exceeds 20V.

THI, pin 5; and the cell voltage higher than the UVLO voltage.

Resistor R_{SENSE} sets fast-charge current into the battery. In fast-charge, the voltage difference between the BATT-, pin 12, and GND, pin 13, is regulated to 250mV. Sink current to DRV, pin 14, increases if the difference between pins 13 and 12 falls below 250mV.

Fast-charge current I_{FAST} is 250mV divided by R_{SENSE} . For example, with an R_{SENSE} of 250mΩ, fast-charge current is 1A.

Trickle charge

Circuits within the IC set trickle-charge current by increasing the current amplifier gain.

When a fast-charge (IFAST) rate C/2, C, 2C, or 4C is used, a C/16 trickle-charge rate is selected automatically, Table 2. Other fast-charge rates can be used, but the trickle-charge current will not be exactly C/16. For simplified design, use a rate of C/2, C, 2C, or 4C, depending on recommendations of the battery manufacturer.

Some manufacturers recommend a lower trickle-charge rate for NiMH batteries. Figure 9 shows a circuit that can reduce trickle-charge to less than C/16. When the circuit is in trickle-charge mode, some of the current is shunted around the battery because Tr_2 is turned on. Select the value of R_7 as follows:

$$R_7 = (V_{BATT} + 0.4V) / (I_{TRICKLE} - I_{BATT})$$

where V_{BATT} equals battery voltage when charged; $I_{TRICKLE}$ equals the IC trickle-charge setting and I_{BATT} equals the desired battery trickle-charge current.

Table 1. Maximum ratings and characteristics for fast-charge controllers.

ABSOLUTE MAXIMUM RATINGS

V+ to BATT-	-0.3V, +7V	REF Current	10mA
BATT- to GND	±1V	Continuous Power Dissipation (TA = +70°C)	
BATT+ to BATT-		Plastic DIP (derate 10.53mW/°C above +70°C)	842mW
Power Not Applied	+20V	Narrow SO (derate 8.70mW/°C above +70°C)	696mW
With Power Applied	The higher of ±20V or ±2V x (programmed cells)	CERDIP (derate 10.00mW/°C above +70°C)	800mW
DRV to GND	-0.3V, +20V	Operating Temperature Ranges:	
FASTCHG to BATT-	-0.3V, +12V	MAX712_C	0°C to +70°C
All Other Pins to GND	-0.3V, (V+ + 0.3V)	MAX712_E	-40°C to +85°C
V+ Current	100mA	MAX712_MJE	-55°C to +125°C
DRV Current	100mA	Storage Temperature Range	-65°C to +150°C
		Lead Temperature (soldering, 10sec)	+300°C

Stresses beyond those listed under 'Absolute Maximum Ratings' may cause permanent damage to the device. These are stress ratings only, and functional operation of the device at these or any other conditions beyond those indicated in the operational sections of the specifications is not implied. Exposure to absolute maximum rating conditions for extended periods may affect device reliability.

ELECTRICAL CHARACTERISTICS

(Iv+ = 10mA, TA = TMIN to TMAX, unless otherwise noted. Refer to Typical Operating Circuit. All measurements are with respect to BATT-, not GND.)

PARAMETER	CONDITIONS	MIN	TYP	MAX	UNITS
V+ Voltage	5mA < Iv+ < 20mA	4.5		5.5	V
Iv+ (Note 1)			5		mA
BATT+ Leakage	V+ = 0V, BATT+ = 17V			5	μA
BATT+ Resistance with Power On	PGM0 = PGM1 = BATT-, BATT+ = 30V			30	kΩ
C1 Capacitance			0.5		μF
C2 Capacitance			5		nF
REF Voltage	0mA < IREF < 1mA	1.96		2.04	V
Undervoltage Lockout	Per cell	0.35		0.5	V
External VLIMIT Input Range		1.25		2.5	V
THI, TLO, TEMP Input Range			0	2	V
THI, TLO Offset Voltage (Note 2)	0V < TEMP < 2V, TEMP voltage rising	-10		10	mV
THI, TLO, TEMP, VLIMIT Input Bias Current			-1	1	μA
VLIMIT Accuracy	1.2V < VLIMIT < 2.5V, 5mA < IDRV < 20mA, PGM0 = PGM1 = V+		-30	30	mV
Internal Cell Voltage Limit	VLIMIT = V+	1.6	1.65	1.7	V
Fast-Charge VSENSE		225	250	275	mV
Trickle-Charge VSENSE	PGM3 = V+	1.5	3.9	7.0	mV
	PGM3 = open	4.5	7.8	12.0	
	PGM3 = REF	12.0	15.6	20.0	
	PGM3 = BATT-	26.0	31.3	38.0	
Voltage-Slope Sensitivity (Note 3)	MAX713		-2.5		mV/tA per cell
	MAX712		0		
Timer Accuracy			-15	15	%
Battery-Voltage to Cell-Voltage Divider Accuracy			-1.5	1.5	%
DRV Sink Current	VDRV = 10V		30		mA
FASTCHG Low Current	VFASTCHG = 0.4V		2		mA
FASTCHG High Current	VFASTCHG = 10V			10	μA
A/D Input Range			1.4	1.9	V

- Note 1: The MAX712/MAX713 are powered from the V+ pin. Since V+ shunt regulates to +5V, R1 must be small enough to allow at least 5mA of current into the V+ pin.
 Note 2: Offset voltage of THI and TLO comparators referred to TEMP.
 Note 3: tA is the A/D sampling interval (see Table 3).

Regulation loop

The regulation loop controls the output voltage between the BATT+ and BATT- terminals, and the current through the battery, via the voltage between BATT- and GND.

Sink current from the DRV pin is reduced when the output voltage exceeds the number of cells times V_{LIMIT} or when the battery current exceeds the programmed charging current. The regulation loop provides the following functions:

When the charger is powered, the battery can be removed without interrupting power to the load.

If the load is connected as shown in Fig. 2, the battery current is regulated regardless of the load current (provided that the input power source can supply both).

Voltage loop

The voltage loop sets the maximum output voltage between the BATT+ and BATT- pins. If V_{LIMIT} is set to less than 2.5V, then the

Table 2. Trickle-charge rate for fast-charge controllers.

PGM3	Fast charge rates	Trickle charge current (ITRICKLE)
V+	4C	$I_{FAST}/64$
OPEN	2C	$I_{FAST}/32$
REF	C	$I_{FAST}/16$
BATT-	C/2	$I_{FAST}/8$

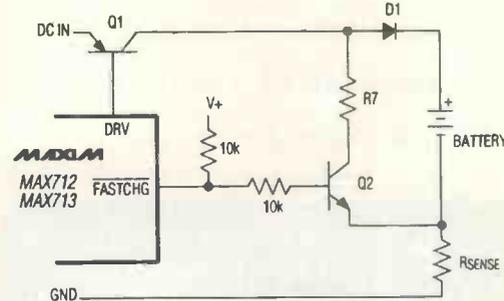
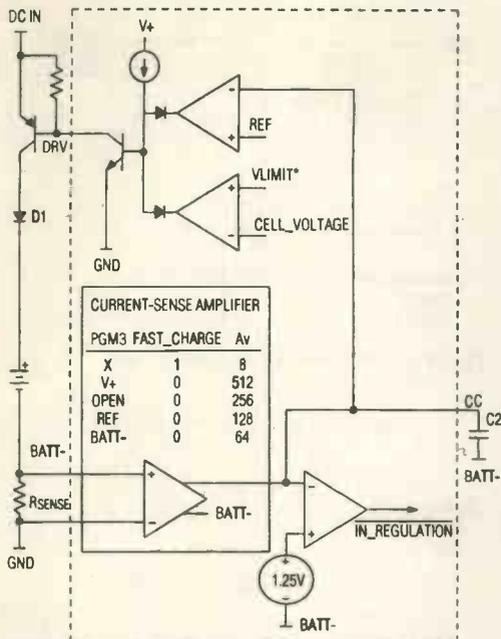


Fig. 9. Reducing trickle charge for NiMH batteries



* IF THE VLIMIT PIN IS TIED TO V+ THEN THIS NODE ACTUALLY EQUALS 1.65V

Fig. 10. Current regulation loop for fast charge controller.

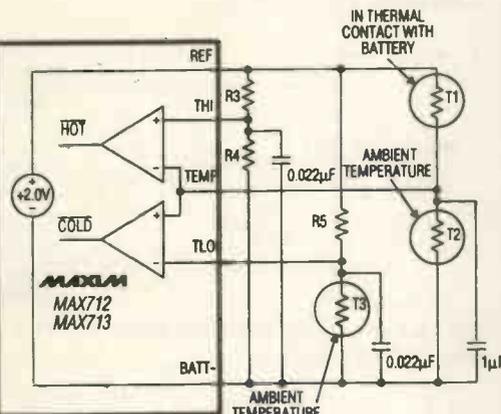


Fig. 11. Controlling charge cut-off with NTC thermistors.

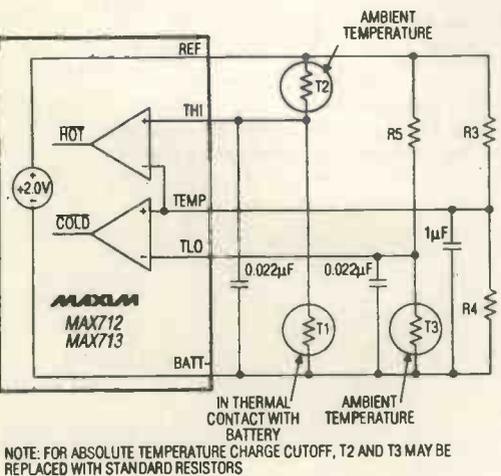


Fig. 12. Controlling charge cut-off with battery thermistor.

maximum BATT+ voltage, referred to BATT-, is V_{LIMIT} multiplied by the number of cells as determined by PGM0, PGM1 connections. If V_{LIMIT} is tied to $V+$, then the maximum BATT+ voltage, referred to BATT-, is 1.65 V multiplied by the number of cells, PGM0, PGM1.

When the battery is removed, the IC does not provide a constant current. Instead, it regulates BATT+ to the maximum voltage, as determined above.

The voltage loop is stabilised by the C_3 filter capacitor. A large filter capacitor is required only if the load is supplied by the IC in the absence of a battery. In this case, set C_3 , in farads, as follows:

$$C_3 = (50 \times I_{LOAD}) / (V_{OUT} \times BWRL)$$

where BWRL equals loop bandwidth in Hz (10,000 is recommended); C_3 is greater than 10 μ F; I_{LOAD} equals external load current in A, and V_{OUT} equals programmed output voltage (V_{LIMIT} times the number of cells).

Current loop

In the current-regulation loop, Fig. 10, stability of the current loop is set by capacitor C_2 at the CC terminal. To get the exact value for C_2 , calculate the current-regulation loop bandwidth (BWCRL) using transistor characteristics (β , f_T , g_m , etc.). For simplified design, use the transistor T_1 types and capacitor C_2 values shown in Fig. 2.

The ICs dissipate power because of the current-voltage product at the DRV pin, which is part of the current loop. Power dissipation – shown in the absolute maximum ratings, Fig. 3 – must not be exceeded.

Voltage-slope cut-off

The a-to-d converter inside both the MAX712 and MAX713 has 2.5mV resolution and stores cell voltage at sampling intervals (t_A) determined by the PGM2/PGM3 connections.

At two t_A intervals, the voltage difference between the cell voltage versus time. Each a-to-d conversion is averaged over five, to filter out noise. Because the battery current is kept constant by the regulation loop – even when there is a varying external load – the conversion results are accurate.

The MAX712 terminates fast-charge when a conversion result is equal to, or less than, its predecessor. The MAX713 terminates when a conversion is at least 2.5mV less than its predecessor. This is the only difference between the two.

Temperature-charge cut-off

Charge cut-off can be controlled with ntc thermistors, Fig. 11. The same-model thermistor should be used for T_1 and T_2 , so that both have the same nominal resistance. Voltage at TEMP is 1V (referred to BATT-) when the battery is at ambient temperature.

The threshold chosen for TLO determines the temperature below which fast-charging is inhibited. If TLO is greater than TEMP, when

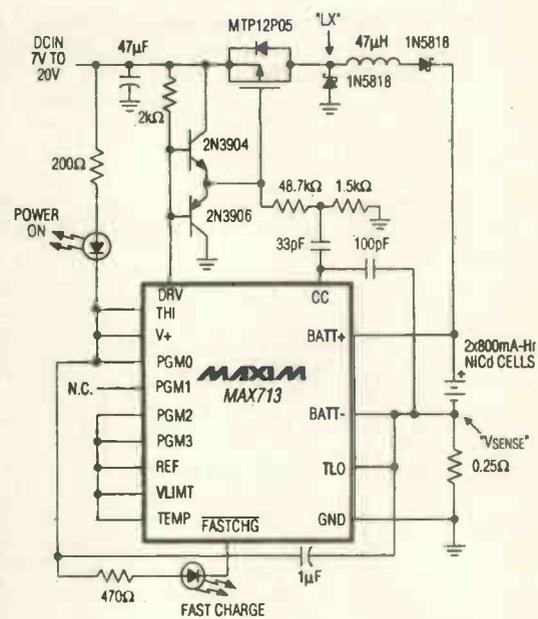


Fig. 13. Basic switch-mode connections.

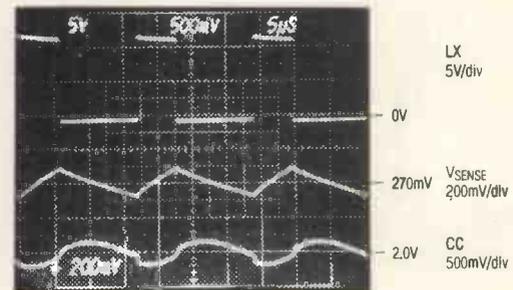


Fig. 14. Switching wave-forms for basic charger circuit.

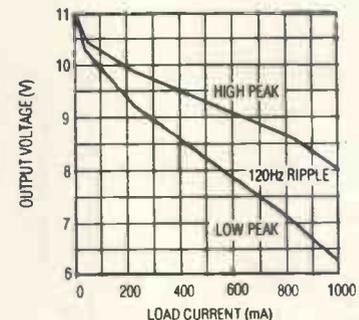


Fig. 15. Sony Radio ac adapter AC-190, 9V dc, 800mA.

the IC starts up, fast-charge will not start until TLO goes below TEMP. The threshold chosen for THI sets the point at which fast-charging terminates. As soon as the voltage on TEMP rises above THI, fast-charge ends and will not restart after TEMP falls below THI.

Cold-temperature charge inhibition may be disabled by removing R_5 , T_3 , and the 0.022 μ F capacitor and tying TLO to BATT-.

To disable the entire temperature-comparator charge-cut-off mechanism, remove T_1 , T_2 , T_3 , R_3 , R_4 , R_5 , and the associated capacitors then make the following connections: TEMP to REF, THI to $V+$, and TLO to BATT-.

Some battery packs may come with a tem-

perature-detecting thermistor connected to the negative terminal of the battery pack. In this case, use the connections shown in Fig. 12. Thermistors T_2 and T_3 may be replaced with standard resistors if absolute temperature charge cutoff is acceptable.

Switch-mode operation

Switch-mode operation, Fig. 13, is used for

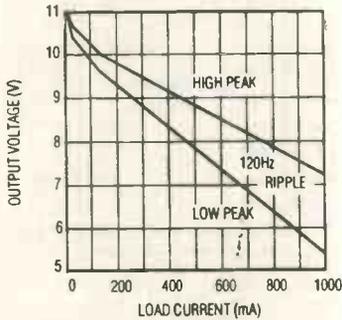


Fig. 16. Sony CD player ac adapter AC-96N, 9V dc, 600mA.

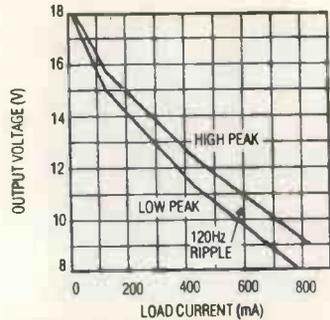


Fig. 17. Panasonic modern ac adapter KX-A11, 12V dc, 500mA.

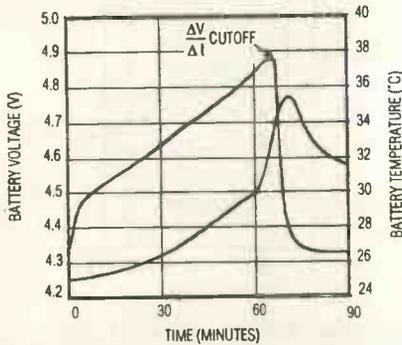


Fig. 18. Three NiMH cells charged with MAX712.

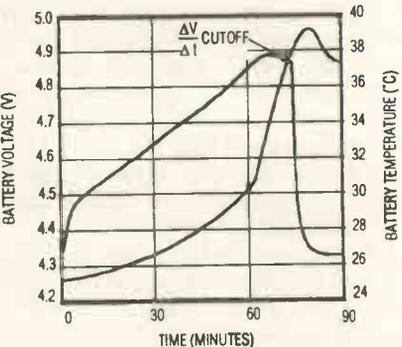


Fig. 19. Three NiMH cells charged with MAX713.

applications where power dissipation of the pass transistor cannot be tolerated. An example is where heat-sinking is not feasible or is too costly. The appropriate circuit uses the error amplifier at the CC pin as a comparator, with a 33pF capacitor adding hysteresis.

Figure 13 is configured to charge two cells at 1A. Higher charge currents and greater numbers of cells can be accommodated by changing the 0.25Ω sensing resistor, and connections PGM0-3.

Switching waveforms, Fig. 14, associated with the circuit show that the arrangement cannot service a load while charging.

Switching frequency can be decreased by increasing the value of the capacitor connected between CC and BATT. Note that the two capacitors connected to the CC pin must be placed as close as possible to the pin and the leads must be as short as possible. The CC node is a high-impedance point – capable of producing high voltages – so logic lines must not be routed near the CC pin.

Line-voltage operation

Consumer-product ac-to-dc wall cubes typically consist of a transformer, a full-wave bridge rectifier, and a capacitor.

Typical characteristics, Figs 15-17, show substantial 120Hz output-voltage ripple. So when selecting an adaptor for use with the 712/713, the lowest dip in the wall-cube voltage during fast-charge should be at least 1V higher than the maximum battery voltage.

Battery charging example

Figures 18 and 19 show the results of charging three AA 1000mAh NiMH batteries from Gold Peak (part number GP1000AAH) GP Batteries at a 1A rate using the 712 and 713.

The circuit from Fig. 2 is used, but with temperature control as shown in Fig. 11. Conditions were:

DC IN uses a Sony AC-190 which is a 9V dc at 800mA ac-to-dc adaptor, Fig. 15. PGM0 is V+, PGM1 is RTEF, PGM2 is REF, and PGM3 is REG. $R_1=200\Omega$, $R_2=150\Omega$, $R_{SENSE}=250m\Omega$, $C_1=1\mu F$, $C_2=0.01\mu F$, $C_3=10\mu F$, $V_{LIMIT}=REF$, $R_3=10k\Omega$, $R_4=15k\Omega$, T_1 and T_2 are both part # 13A1002 (Alpha Thermistor 00 1 800-235-5445). If R_5 and T_3 are omitted then $TLO=BATT-$.

Absolute maximum rating for the BATT+ input voltage must be limited by external circuits, Fig. 20, when DC IN is not applied.

Current-sense resistor R_{SENSE} causes a small efficiency loss during battery use. The efficiency loss is significant only if R_{SENSE} is much greater than the internal resistance of the battery pack. The circuit in Fig. 21 can be used to shunt R_{SENSE} whenever power is removed from the charger.

Figure 22 shows a circuit used to indicate charger status, with logic-level outputs, and Fig. 23 shows a led drive circuit that indicates charger status.

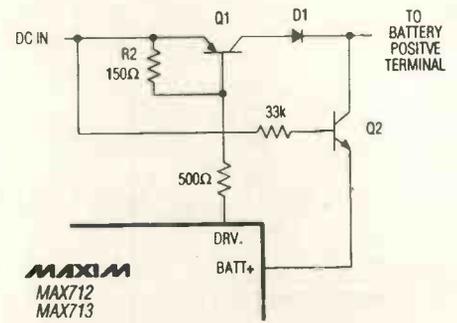


Fig. 20. Connections for high cell counts.

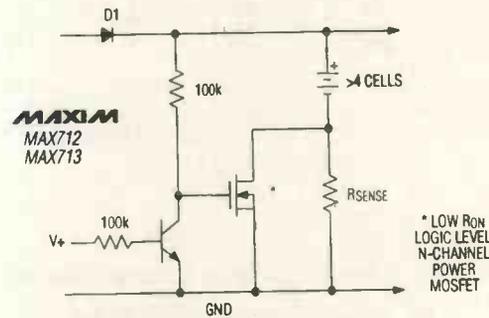


Fig. 21. Connections to shunt-sense resistor when output power is removed.

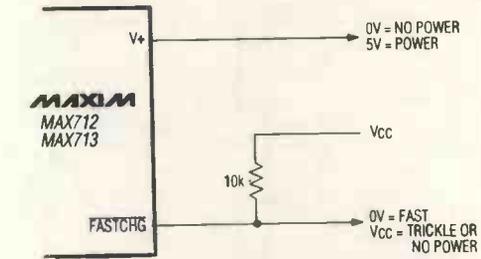


Fig. 22. Connections to indicate charger status via external logic circuitry.

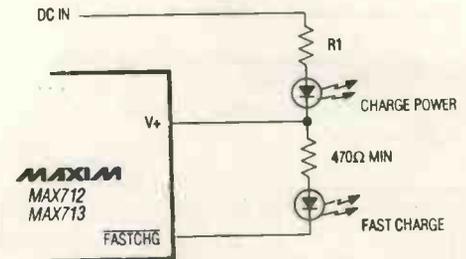


Fig. 23. Connections needed to indicate charger status via light-emitting diodes.

Part II will show how to use the MAX712/713 and other ICs in practical circuits designed for specific charging applications and cell

Building blocks of time

Traditionally, Radio-Code time signal receivers have been expensive, self-contained units, but now there are low-cost modules providing access to the raw time signal for well under £30.

Radio controlled clocks – allowing access to a time reference accurate to a second in a million years – are proving more and more popular. A considerable fall in price has led to the sale of over 6,000,000 radio controlled clocks and radio time receiver chips in Europe – in the last twelve months alone.

In the near future, it is likely that radio controlled time receivers will replace traditional clocks in many commercial, personal time keeping and control applications.

Currently, radio controlled time is used for, heating system timing, accurate time stamping of data, encryption of data and digital signatures and synchronising clocks to an accuracy of milliseconds. It is also used for setting the time on fax machines, video recorders, wrist watches, etc.

What is time?

Time is no longer calculated by observing the stars but 'synthesised' in a laboratory. In Britain this is the responsibility of the National Physics Laboratory.

Every minute, the National Physical Laboratory, NPL, transmits a time telegram as a binary coded decimal signal. This contains the time, the date and the calendar day. NPL is part of an international network, each making a contribution to co-ordinated universal time. This means that world-wide, all clocks agree to the nanosecond.

Earth loses around 5ms a day. To ensure that solar and atomic times remain the same, laboratory time is reset approximately once a year.

MSF is the call sign of the NPL time code transmitter. Anyone can make use of the signal transmitted. Equipment require is a facility to receive and condition the time telegrams, and a receiver with a suitable tuned antenna. A receiver module measuring just 7mm by 24mm is shown in the photograph at the end of this article.

MSF's time telegram

Every minute, the MSF transmitter at Rugby transmits a data stream containing time information. The first second of every minute contains 'fast code' information relating to the minute in which it is transmitted. Timing is governed by Co-ordinated Universal Time.

The remaining 59 seconds contain 'slow code' information as UK time relating to the minute following that in which it was transmitted.

Table 1. Time telegram of MSF transmitter – binary code for seconds 17 to 30 and data for March 1996.

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

Fig. 1. Example of how date information is encoded in the MSF transmission. This represents March 1996.

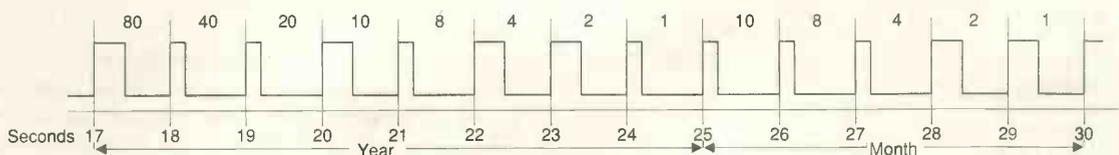
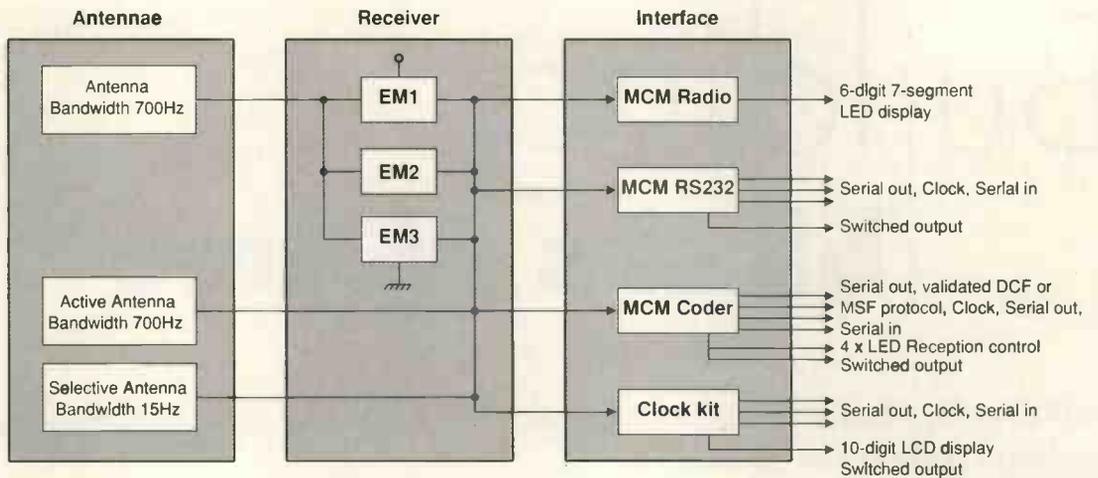


Fig. 2. Key building blocks of the Radio-Code receive module set. These can be used to build anything from a basic received producing the raw data stream up to a stand-alone Radio Code clock.



Carrier frequency of the time telegram is 60kHz. The amplitude is switched off at the beginning of each second, for 100ms or 200ms; these periods are the so-called second markers. The short ones, at 100ms, correspond to a zero bit, i.e. binary 0, and the long ones, 200ms, to one bit, binary 1.

At the switch over to the next minute bits 52 to 59 are set permanently to 10000001₂. This sequence is unique and identifies the following minute marker, second 00. Table 1 shows the format of the MSF time telegram.

The Radio-Code receiver modules discussed in this article only decode the slow code, using the data starting at second 17 and extending to second 59. Numbers are represented in the binary-coded decimal form, i.e. decimal 9₁₀ becomes binary 1001₂.

For transmitting the complete time and date, it is sufficient to have 35 bits. The hour decimals for example only require two bits. They can only have the value 0₁₆, 1₁₆ or 2₁₆ = 00₂, 01₂ or 10₂. For the year, this has been confined to two digits, so 96 represents 1996. Figure 1 shows the seconds 17 to 30 with the data March 1996.

Decoding the time signal

There are many circuits for decoding the UK's MSF time signal. The modules outlined in Fig. 2 have the advantages of small size, low cost and high sensitivity. They also have a highly tuned antenna. The receiver modules contain all the support electronics, tuned crystal, etc, required to receive the MSF time signal.

All that is needed to reconstitute the slow code, exactly as broadcast by Rugby, is an antenna.

As an alternative to the receiver module, it is possible to buy the chip at the heart of the module. This allows the user to select a quartz filter for receiving time signals from the UK's MSF transmitter, Germany's DCF signal, USA's WWVB, Switzerland's HBG and Japan's JG2As. Antennas tuned for these signals are also available.

Receiving MSF

Receiver modules contain a very sensitive straight through receiver IC with an open-collector n-p-n output. A block diagram of the receiver module IC is shown in Fig. 3.

The very low bandwidth of about 10Hz is achieved by using, a 60kHz crystal filter. This guarantees good suppression of disturbances on other frequencies. The high impedance input represents a very small load for the antenna which maintains its quality.

Internally, the radio frequency is rectified and used for both gain regulation and to represent the envelope of the radio frequency. Rectified radio-frequency voltage is compared to a threshold level and the comparator result output.

Antenna design

At 500km from the radio station, field strength of the MSF signal is about 1mV/m. At this distance the antenna gives an output voltage of about 50µV, but the receiver needs only 1µV for good signal demodulation.

The connection between receiver IC and antenna should be as short as possible with the bar axis of the ferrite antenna oriented at right angles to the direction to the radio station. Antenna output voltage is directly related to the cosine of this angle. This means there is a wide range of angle in which good reception is possible, but only a narrow range with no reception when the axis points directly to the radio station.

As mentioned, the antenna is best mounted close to the receiver. Where this is impractical an active antenna is used. The active antenna contains a receiver module and a ferrite antenna in a plastic package.

The antenna should be as short as possible. If the length of the antenna is over 20cm, the cable capacitance increases the antenna resonant circuit capacitance and lowers the resonant frequency.

Microcontroller modules

A radio receiver module is available for producing a battery powered radio clock unit or an alarm-clock radio. It is designed to drive an liquid-crystal or led display via static shift registers

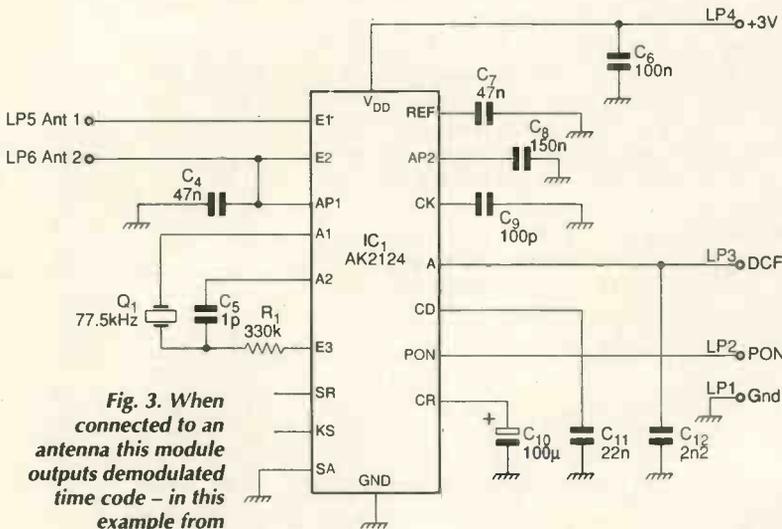


Fig. 3. When connected to an antenna this module outputs demodulated time code - in this example from Frankfurt's DCF transmissions.

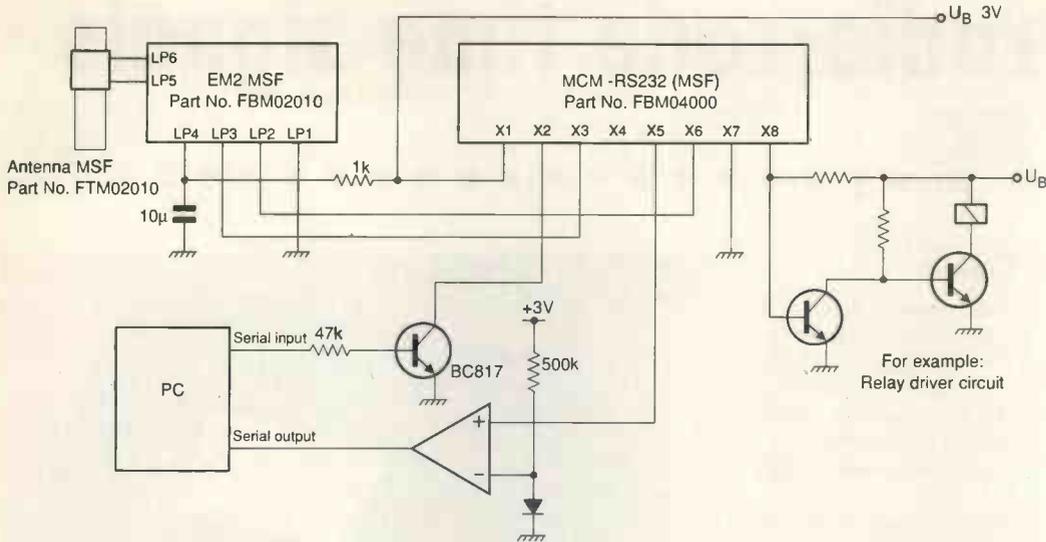


Fig. 6. PC-interfacing Radio-Code clock incorporating an active antenna for increased sensitivity and enhanced MSF reception.

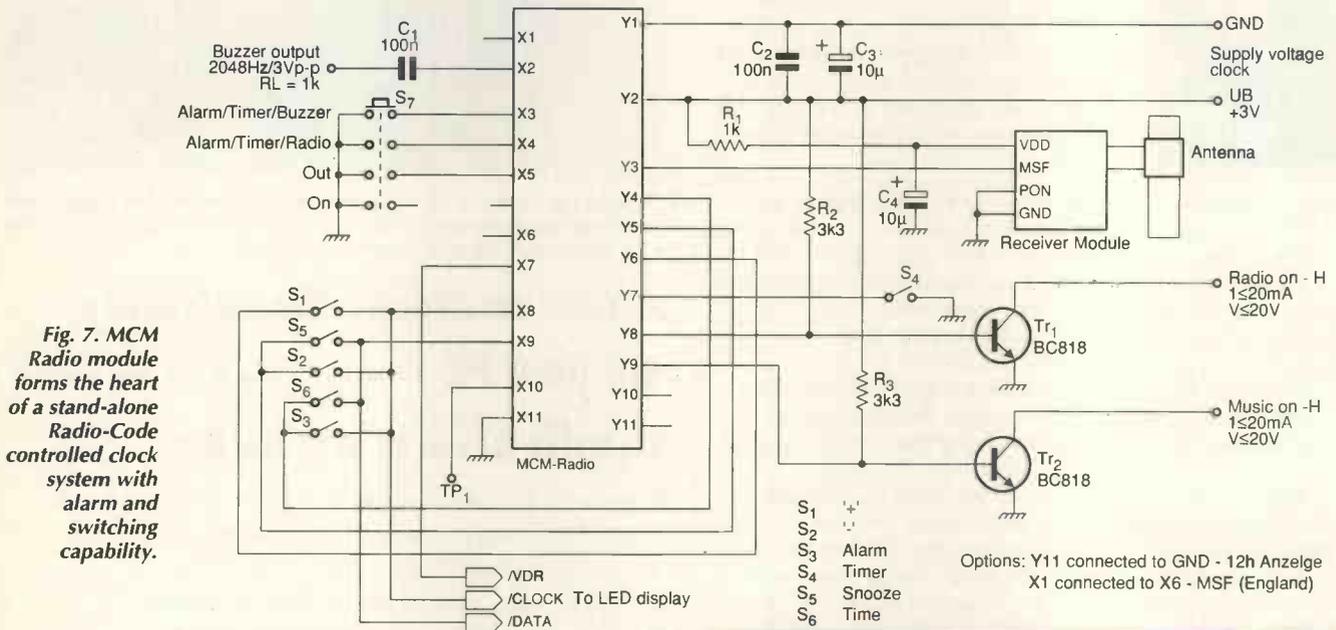


Fig. 7. MCM Radio module forms the heart of a stand-alone Radio-Code controlled clock system with alarm and switching capability.

without multiplexing.

The micro controller within the module is four bit device and it decodes the demodulated time telegram from the receiver modules. The controller also translates the time signal for different formats.

Serial i/o via RS232

A second module is available to facilitate accessing of the time signal on a computer. The MCM RS232 is designed for applications in which a host computer receives the exact time information via a serial interface.

The advantage of this module compared to direct host decoding of receiver output is the presence of exact time information all the time. Once synchronised, the controller predicts the incoming time information so that it can send out the translated RS232 data stream synchronously with the incoming rf signal.

There is also a switching output. Switch on and off times can be set by the host computer. If the switching output is connected to a mains switch the host computer can turn itself on and off by setting the appropriate switch on and switch off times.

An application showing the MCM RS232 serial interface and the EM2 MSF receiver modules is shown in Fig. 6.

An application circuit for the radio-controlled clock kit with active antenna is shown in Fig. 7



Typical MSF receiver module with a footprint of only 7 by 24mm.

Receive radio-code time signals on your PC.....

Modules for Radio Code

The EM1, connected to the MSF passive antenna, receives the 60kHz Rugby signal and outputs the slow code comprising seconds 17 to 59. Operating from a 3V supply, the EM1 has an antenna input, supply pins, a keying input and an open-collector MSF output.

Quiescent current in standby is less than 1µA.

MCM RS232, combined with an EM1 receiver and antenna, feeds decoded time information to a computer via RS232 via its internal microcontroller. MCM Radio, designed for stand-alone applications, forms the heart of a Radio-Code clock with LCD or LED read-out. This module includes alarm facilities.



Plugging a radio-code receiver into your PC's COM port and running the dos and windows software supplied gives you access to the atomic-clock referenced 60kHz time signal transmitted from Rugby. This signal is accurate to a second in a million years and corrected automatically for summer/winter time.

Based on a highly-tuned and reliable receiver module with antenna, the system automatically updates the PC's clock at switch on and at any other desired time. Under Windows, an icon is available signalling to the operator that the receiver is receiving the Rugby signal, and indicating the current time and date.

Time data received by the PC is via standard RS232 and well documented, allowing you to use atomic-clock referenced timing and date stamping in your own applications. Sending the ASCII letter o for example returns a 15-character string representing hours, minutes, seconds, day of week, day of month, month, year and summer-time and receiver status.

Normally, the receiver module together with dos and windows software costs £69.50, or £99.50 for a version with in-built liquid-crystal display for time and date display. Until 15 April 1996, Galleon is offering these two products to *EW* readers at special 25% discount prices of £52.13 and £74.63 respectively.

All prices quoted are inclusive of VAT, but excluding £2 postage.



This module receives radio-code data and sends it to the PC via an RS232 link. From the COM port, RS232 time information can easily be incorporated into user applications. Software supplied as standard synchronises the PC clock and - under windows - displays time/date together with confirmation of the received signal.



The desk-top version of the radio-code PC clock features two alarms and an integral display showing hours/mins secs or hours/mins day together with day+month, signal strength indications.

Access to atomic time accuracy via your PC.....

Applications include:

- research and development
- synchronising encryption-key changes
- broadcast transmission sync
- timing video monitoring in security systems
- controlling public and distributed clocks
- distributed timing for remote communications
- access control
- production monitoring
- linking remote networks
- data security

Ordering details

Product	Normal price	EW discount price
EM1	£16.99	£12.74
MCMRS232	£19.99	£14.99
MCM Radio	£19.99	£14.99
Pass antenna	£4.99	£3.74
Act antenna	£34.99	£26.24
Rec. w disp	£99.50	£74.63
Rec. w-RS232	£69.50	£52.13

Please send your order to Galleon Systems Ltd, together with payment. Alternatively fax credit card number, type and expiry date with order on 0121 608 4477 or telephone Galleon on 0121 359 0981. If possible, include a daytime telephone number with your order. Address orders and all correspondence relating to this order to Galleon Systems Ltd, Aston Science Park, Love Lane Birmingham B7 4BJ.

*Overseas readers can also obtain this discount but details vary according to country. Please ring, write or fax Galleon Systems.

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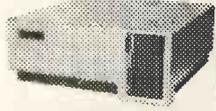
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Order as HIGRADE 286 **ONLY £129.00 (E)**

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Massive purchases of standard 5 1/4" and 3 1/2" drives enables us to present prime product at industry beating low prices! All units (unless stated) are **BRAND NEW** or removed from often brand new equipment and are fully tested, aligned and shipped to you with a 90 day guarantee and operate from standard voltages and are of standard size. All are IBM-PC compatible (if 3 1/2" supported on your PC).

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Converts your colour monitor into a QUALITY COLOUR TV!!



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- For overseas PAL versions state 5.5 or 6 MHz sound specification. *For cable / hyperband reception Telebox MB should be connected to a cable type service. Shipping code on all Teleboxes is (B)

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One of the highest specification monitors you will ever see - At this price - Don't miss it!!

Mitsubishi FA3415ETKL 14" SVGA Multisync colour monitor with fine 0.28 dot pitch tube and resolution of 1024 x 768. A variety of inputs allows connection to a host of computers including IBM PC's in CGA, EGA, VGA & SVGA modes, BBC, COMMODORE (including Amiga 1200), ARCHIMEDES and APPLE. Many features: Etched faceplate, text switching and LOW RADIATION MPR specification. Fully guaranteed, supplied in EXCEL-

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VGA cable for IBM PC included.
External cables for other types of computers CALL

5000 Monitors from stock !!!
HERCULES, EGA, CGA, VGA, SVGA - 6 to 26"
+ Many special items - CALL with your needs!

Just In - Microvitec 20" VGA (800 x 600 res.) colour monitors. Good SH condition - from £299 - CALL for info

PHILIPS HCS35 (same style as CM8833) attractively styled 14" colour monitor with both RGB and standard composite 15.625 KHz video inputs via SCART socket and separate phono jacks. Integral audio power amp and speaker for all audio visual uses. Will connect direct to Amiga and Atari BBC computers. Ideal for all video monitoring / security applications with direct connection to most colour cameras. High quality with many features such as front concealed flap controls, VCR correction button etc. Good used condition - fully tested - guaranteed
Dimensions: W14" x H12 1/2" x 15 1/2" D. **Only £95 (E)**

PHILIPS HCS31 Ultra compact 9" colour video monitor with standard composite 15.625 KHz video input via SCART socket. Ideal for all monitoring / security applications. High quality, ex-equipment fully tested & guaranteed (possible minor screen burns). In attractive square black plastic case measuring W10" x H10" x 13 1/2" D. 240 V AC mains powered. **Only £79.00 (D)**

KME 10" 15M10009 high definition colour monitors with 0.28" dot pitch. Superb clarity and modern styling. Operates from any 15.625 khz sync RGB video source, with RGB analog and composite sync such as Atari, Commodore Amiga, Acorn Archimedes & BBC. Measures only 13 1/2" x 12" x 11". Good used condition. **Only £125 (E)**

20" 22" and 26" AV SPECIALS

Superbly made UK manufacture. PIL All solid state colour monitors, complete with composite video & optional sound input. Attractive teak style case. Perfect for Schools, Shops, Disco, Clubs, etc. in EXCELLENT little used condition with full 90 day guarantee.

20"....£135 22"....£155 26"....£185 (F)

SPECIAL INTEREST ITEMS

- MITLS FA3445ETKL 14" Industrial spec SVGA monitors 2Kw to 400 kW - 400 Hz 3 phase power sources - ex stock £245
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- Stanecol STA15 15kW RF induction heater system £750
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- IBM 53F5501 Token Ring ICS 20 port lobe modules £750
- IBM MAU Token ring distribution panel 8228-23-5050N £95
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- Trend DSA 274 Data Analyser with G703(2M) 64 i/o £950
- HP APOLLO R7X00 system units £1800
- HP6621A Dual Programmable GPIB PSU 0-7 V 160 watts £1800
- HP6264 Rack mount variable 0-20V @ 20A metered PSU £975
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- EG-NG Brookdeal 9503EC Precision lock in amp £650
- OTC Ltd 1550 SM Stabilised IR laser calibration source £2250
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- Andrews LARGE 3.1 m Satellite Dish + mount (For Voyager) £375
- Thurby LA 160B logic analyser £950
- Sekonic SD 150H 18 channel digital Hybrid chart recorder £1995
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- System Video 1152 PAL waveform monitor £485
- Test Lab - 2 mtr square quietised acoustic test cabinets £300
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Please call for further details on the above items

19" RACK CABINETS

Superb quality 6 foot 40U
Virtually New, Ultra Smart
Less than Half Price!

Top quality 19" rack cabinets made in UK by Optima Enclosures Ltd. Units feature designer, smoked acrylic lockable front door, full height lockable half louvered back door and louvered removable side panels. Fully adjustable internal fixing struts, ready punched for any configuration of equipment mounting plus ready mounted integral 12 way 13 amp socket switched mains distribution strip make these racks some of the most versatile we have ever sold. Racks may be stacked side by side and therefore require only two side panels to stand singly or in multiple bays. Overall dimensions are: 77 1/2" H x 32 1/2" D x 22" W. Order as:

- OPT Rack 1 Complete with removable side panels. £335.00 (G)
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32U - High Quality - All steel RakCab

Made by Eurocraft Enclosures Ltd to the highest possible spec, rack features all steel construction with removable side, front and back doors. Front and back doors are hinged for easy access and all are lockable with five secure 5 lever barrel locks. The front door is constructed of double walled steel with a 'designer style' smoked acrylic front panel to enable status indicators to be seen through the panel, yet remain unobtrusive. Internally the rack features fully slotted reinforced vertical fixing members to take the heaviest of 19" rack equipment. The two movable vertical fixing struts (extras available) are pre punched for standard 'cage nuts'. A mains distribution panel internally mounted to the bottom rear, provides 8 x IEC 3 pin Euro sockets and 1 x 13 amp 3 pin switched utility socket. Overall ventilation is provided by fully louvered back door and double skinned top section with top and side louvres. The top panel may be removed for fitting of integral fans to the sub plate etc. Other features include: fitted castors and floor levelers, prepunched utility panel at lower rear for cable / connector access etc. Supplied in excellent, slightly used condition with keys. Colour Royal blue. External dimensions mm=1625H x 635D x 603 W. (64" H x 25" D x 23 1/2" W)

Sold at LESS than a third of makers price !!

A superb buy at only £195.00 (G)

Over 1000 racks - 19" 22" & 24" wide
3 to 44 U high. Available from stock !!
Call with your requirements.

TOUCH SCREEN SYSTEM

The ultimate in 'Touch Screen Technology' made by the experts - MicroTouch - but sold at a price below cost !! System consists of a flat translucent glass laminated panel measuring 29.5 x 23.5 cm connected to an electronic controller PCB. The controller produces a standard serial RS232 or TTL output which continuously gives simple serial data containing positional X & Y co-ordinates as to where a finger is touching the panel - as the finger moves, the data instantly changes. The X & Y information is given at an incredible matrix resolution of 1024 x 1024 positions over the entire screen size !! A host of available translation software enables direct connection to a PC for a myriad of applications including: control panels, pointing devices, POS systems, controllers for the disabled or computer un-trained etc etc. (Imagine using your finger with "Windows", instead of a mouse !! (a driver is indeed available !!) The applications for this amazing product are only limited by your imagination!! Complete system including Controller, Power Supply and Data supplied at an incredible price of only: **£145.00 (B)**
Full MICROTUCH software support pack and manuals for IBM compatible PC's £29.95 RFE - Tested

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Order as: XT RAM UG. 256k. £34.95 or 512k £39.95 (A1)

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- Shipping on all fans (A). Blowers (B). 50,000 Fans Ex Stock CALL

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CIRCLE NO. 126 ON REPLY CARD

CIRCUIT IDEAS

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Recognising the importance of a good idea, Thurlby Thandar Instruments will be giving away one of these excellent instruments once every six months. This incentive is in addition to our monthly £100 'best circuit idea' award and £25 awards for each circuit published.

£100 WINNER

Inductance on a capacitance meter

Using this circuit, you can measure inductance by means of a capacitance meter. It offers the advantage over a gyrator in that the inductor is earthed.

$$Y = j\omega L / (R_8 + j\omega L) \times (V + Y)$$

$$Y = j\omega L V / R_8$$

$$U = 2V - (V + Y) = V - j\omega L V / R_8$$

Since $V/I_{in} = (R_8 R_1) / j\omega L$, which is a capacitive impedance,

$$I_{in} R_1 = V - U$$

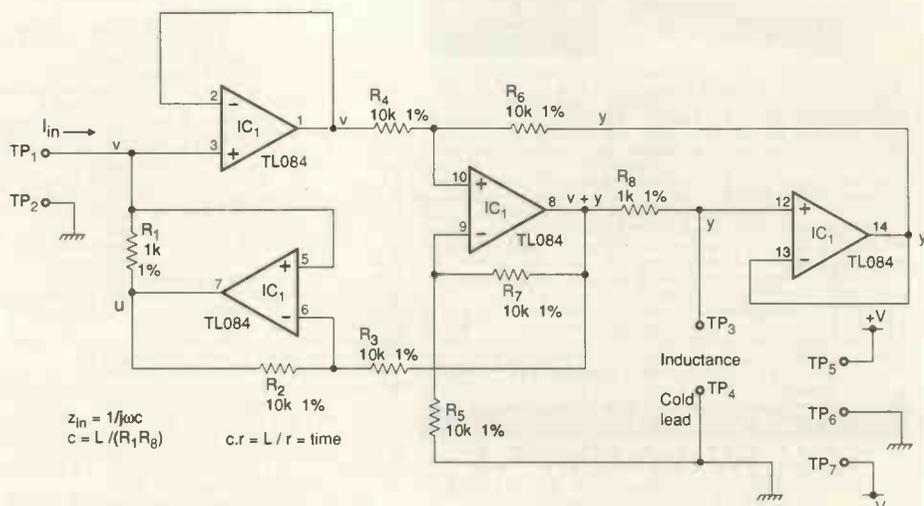
$$I_{in} = j\omega L V / (R_8 R_1)$$

$$1 / j\omega C = (R_8 R_1) / j\omega L$$

$$C = L / (R_8 R_1)$$

Making R_8 and R_1 1kΩ presents a 1H inductor as 1µF, 10mH as 10nF, etc.

Marco Trinci
Montecatini Terme
Italy



Without calculation, this circuit measures inductance in terms of capacitance, while allowing the inductor to be earthed.

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 TEK 494AP 1Kc/s - 21GHz - £7k.
 TEK 496P 1KHz-1.8GHz - £4k.
 TEK 5L4N 0-100KHz - £400.
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 TEK 7L12 - 100KHz-1800Mc/s - £1000.
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 Marconi TF1245 Circuit Magnification meter + 1246 & 1247 Oscillators - £100-£300.
 Marconi microwave 6600A sweep osc., mainframe with 6650 PI - 18-26.5GHz or 6651 PI - 26.5-40GHz - £1000 or PI only £600. MF only £250.
 Marconi distortion meter type TF2331 - £150. TF2331A - £200.
 Tektronix Plug-ins 7A13 - 7A14 - 7A18 - 7A24 - 7A26 - 7A11 - 7M11 - 7S11 - 7D10 - 7S12 - S1 - S2 - S6 - S52 - PG506 - SC504 - SG502 - SG503 - SG504 - DC503 - DC508 - DD501 - WR501 - DM501A - FG501A - TG501 - PG502 - DC505A - FG504 - 7880 + 85-7892A
 Gould J3B test oscillator + manual - £150.
 Tektronix Mainframes - 7603 - 7623A - 7613 - 7704A - 7844 - 7904 - TM501 - TM503 - TM506 - 7904A - 7834 - 7623 - 7633.
 Marconi 6155A Signal Source - 1 to 2GHz - LED readout - £400.
 Barr & Stroud Variable filter EF3 0.1Hz - 100Kc/s + high pass + low pass - £150.
 Marconi TF2163S attenuator - 1GHz - £200.
 Farnell power unit H60/50 - £400 tested. H60/25 - £250.
 Rascal/Dana 9300 RMS voltmeter - £250.
 HP 8750A storage normalizer - £400 with lead + S/A or N/A Interface.
 Marconi TF2330 - or TF2330A wave analysers - £100-£150.
 Tektronix - 7S14 - 7T11 - 7S11 - 7S12 - S1 - S2 - S39 - S47 - S51 - S52 - S53 - 7M11.
 Marconi mod meters type TF2304 - £250.
 HP 5065A rubidium vapour FX standard - £1.5k.
 Systron Donner counter type 6054B - 20Mc/s - 24GHz - LED readout - £1k.
 Rascal/Dana 9083 signal source - two tone - £250.
 Systron Donner - signal generator 1702 - synthesized to 1GHz - AM/FM - £600.
 Tektronix TM515 mainframe + TM5006 mainframe - £450 - £850.
 Farnell electronic load type RB1030-35 - £350.
 Rascal/Dana counters - 9904 - 9905 - 9906 - 9915 - 9916 - 9917 - 9921 - 50Mc/s - 3GHz - £100-£450 - all fitted with FX standards.
 HP4815A RF vector impedance meter c/w probe - £500-£600.
 Marconi TF2092 noise receiver. A, B or C plus filters - £100-£350.
 Marconi TF2091 noise generator. A, B or C plus filters - £100-£350.
 Marconi 2017 S/G 10KHz - 1024MHz.
 HP180TR, HP182T mainframes £300-£500.
 Philips panoramic receiver type PM7900 - 1 to 20GHz - £400.
 Marconi 6700A sweep oscillator + 18GHz PI's available.
 HP8505A network ANZ + 8503A S parameter test set + 8501A normalizer - £4k.
 HP8505 network ANZ 8505 + 8501A + 8503A.
 Rascal/Dana VLF frequency standard equipment. Tracer receiver type 900A + difference meter type 527E + rubidium standard type 9475 - £2750.
 HP signal generators type 626 - 628 - frequency 10GHz - 21GHz.
 HP 432A - 435A or B - 436A - power meters + powerheads - Mc/s - 40GHz - £200-£1000.
 Bradley oscilloscope calibrator type 192 - £600.
 HP8614A signal generator 800Mc/s - 2.4GHz, new colour £400.
 HP8616A signal gen 1.8GHz - 4.5GHz, new colour £400.
 HP 3325A syn function gen 20Mc/s - £1500.
 HP 3336A or B syn level generator - £500-£600.
 HP 3586B or C selective level meter - £750-£1000.
 HP 3575A gain phase meter 1Hz - 13Mc/s - £400.
 HP 8683D S/G microwave 2.3 - 13GHz - opt 001 - 003 - £4.5k.
 HP 8660 A-B-C syn S/G. AM + FM + 10Kc/s to 110Mc/s PI - 1Mc/s to 1300Mc/s - 1Mc/s to 2600Mc/s - £500-£2000.
 HP 8640B S/G AM-FM 512Mc/s or 1024Mc/s. Opt 001 or 002 or 003 - £800-£1250.
 HP 86222BX Sweep PI - 01 - 2.4GHz + ATT - £1750.
 HP 8629A Sweep PI - 2 - 18GHz - £1000.
 HP 86290B Sweep PI - 2 - 18GHz - £1250.
 HP 86 Series PI's in stock - splitband from 10Mc/s - 18.6GHz - £250-£1k.
 HP 8620C Mainframe - £250. IEEE - £500.
 HP 8615A Programmable signal source - 1MHz - 50Mc/s - opt 002 - £1k.
 HP 8614A Sweep generator. 1 - 110Mc/s - £300.
 HP 8488A HP - IB switch control unit - £500 + control modules various - £175 each.
 HP 8160A 50Mc/s programmable pulse generator - £1000.
 HP 853A MF ANZ - £1.5k.
 HP 8349A Microwave Amp 2 - 20GHz Solid state - £1500
 HP 3585A Analyser 20Hz - 40Mc/s - £4k.
 HP 8569B Analyser .01 - 22GHz - £5k.
 HP 3580A Analyser 5Hz - 50KHz - £1k.
 HP 1980B Oscilloscope measurement system - £600.
 HP 3455A Digital voltmeter - £500.
 HP 3437A System voltmeter - £300.
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 HP 5335A Universal counter - 200Mc/s - £500.
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 HP 3763A Error detector - £500.
 HP 3764A Digital transmission analyser - £600.
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 HP 3780A Pattern generator detector - £400.
 HP 3781A Pattern generator - £400.
 HP 3781B Pattern generator (bell) - £300.
 HP 3782A Error detector - £400.
 HP 3782B Error detector (bell) - £300.
 HP 3785A Jitter generator + receiver - £750-£1k.
 HP 8006A Word generator - £100-£150.
 HP 8016A Word generator - £250.
 HP 8170A Logic pattern generator - £500.
 HP 59401A Bus system analyser - £350.
 HP 59500A Multiprogrammer HP - IB - £300.
 Philips PM5390 RF syn - 0.1 - 1GHz - AM + FM - £1000.
 S.A. Spectral Dynamics SD345 spectrascope 111 - LF ANZ - £1500.
 Tektronix R7912 Transient waveform digitizer - programmable - £400.
 Tektronix TR503 + TM503 tracking generator 0.1 - 1.6GHz - £1k - or TR502.
 Tektronix 576 Curve tracer + adaptors - £300.
 Tektronix 577 Curve tracer + adaptors - £300.
 Tektronix 1502/1503 TDR cable test set - £1000.
 Tektronix AM503 Current probe + TM501 m/frame - £1000.
 Tektronix SC501 - SC502 - SC503 - SC504 oscilloscopes - £75-£350.
 Tektronix 465 - 465B - 475 - 2213A - 2215 - 2225 - 2235 - 2245 - 2246 - £250-£1000.
 Kikusui 100Mc/s Oscilloscope COS6100M - £350.
 Nicolet 3091 LF oscilloscope - £400.
 Rascal 1991 - 1992 - 1988 - 1300Mc/s counters - £500-£900.
 Fluke 80K-40 High voltage probe in case - BN - £100.
 Rascal Recorders - Store 4 - 4D - 7 - 14 channels in stock - £250 - £500.
 Rascal Store Horse Recorder & control - £400-£750 Tested.
 EIP 545 microwave 18GHz counter - £1200.
 Fluke 510A AC ref standard - 400Hz - £200.
 Fluke 355A DC voltage standard - £300.
 Wiltron 610D Sweep Generator + 6124C PI - 4 - 8GHz - £400.
 Wiltron 610D Sweep Generator + 61084D PI - 1Mc/s - 1500Mc/s - £500.
 Time Electronics 9814 Voltage calibrator - £750.
 Time Electronics 9811 Programmable resistance - £600.
 Time Electronics 2004 D.C. voltage standard - £1000.
 HP 8699B Sweep PI YIG oscillator .01 - 40GHz - £300. 8699B MF - £250. Both £500.
 Schlumberger 1250 Frequency response ANZ - £1500.
 Dummy Loads & power att up to 2.5 kilowatts FX up to 18GHz - microwave parts new and ex equipt - relays - attenuators - switches - waveguides - Yigs - SMA-APC7 plugs - adaptors.
 B&K Items in stock - ask for list.
 W&G Items in stock - ask for list.
 Power Supplies Heavy duty + bench in stock - Farnell - HP - Weir - Thurlby - Rascal etc. Ask for list.

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CIRCLE NO. 127 ON REPLY CARD

Remote multichannel resistance measurement

This low-power circuit measures the value of remote resistive sensors using any type of connection such as wire, infrared or ultrasonics; I use a vhf radio link.

With a UMC UM3758-108AM encoder/decoder, it is possible to transmit 8-bit data combined with 10 address bits. Depending upon the logical level on the mode input, the circuit acts as a decoder or encoder. In this application, the address bits are hard-wired high and the measuring resolution is limited to 6-bit, for reasons explained later.

In encoder mode, the Tx/Rx pin transmits the address/data information as long as power is supplied. A built-in RC oscillator, needing only R_9 and C_9 , provides clocking; matching of the encoder/decoder's oscillator frequencies is not too critical; 5% resistors will suffice.

Serial data is supplied to the Rx input in decoder mode, where it is examined bit by bit as received. Only if two successive address/data combinations match is data transferred to the output data pins D_{1-8} , which latch the data until the next valid data is being received. The Tx/Rx out pin switches low if data matches, returning high after two successive unmatched address words.

System timing is controlled by IC_1 , a 74HC4060. The positive-going edge of the Q5 output of IC_1 triggers an enable signal, the duration of which is adjustable by P_1 about

every 60s, depending upon R_1 , R_2 , C_1 . At the start of this enable time, the counter in IC_4 is reset to zero by the positive-going edge of the oscillator enable signal via C_3 , R_4 . Diode D_8 gates the IC_4 clock signal, pulling down the clock input to disable counting. During the enable time, a sensor-dependent count is reached on the outputs Q_{4-9} .

Be careful to avoid overflow, by selecting the right combination between the enable time and the IC_4 oscillator frequency, which depends on the sensor used.

At the end of the oscillator enable time, the falling edge of IC_{5b} Q output triggers IC_{5a} to switch on the power supply for both the UM3758 and the transmitter. To be sure that a minimum of three address/data codes are transmitted, the transmitter on time, adjustable by P_2 , is 250-300ms. Raising the UM3758 oscillator frequency makes it possible to use an even shorter transmitter time, but bear in mind that the receiver must be capable of detecting this signal. A low-loss, dual P-channel mosfet, a Siliconix SI9933DY with $R_{gs} < 0.2\Omega$, switches the transmitter/encoder supply.

To prevent any current flow through the data input of IC_2 when it is powered down, the counter outputs of IC_4 are connected via diodes D_{1-6} . Because of the internal architecture of the UM3758, an open input will be seen as logical 1, so there is no need

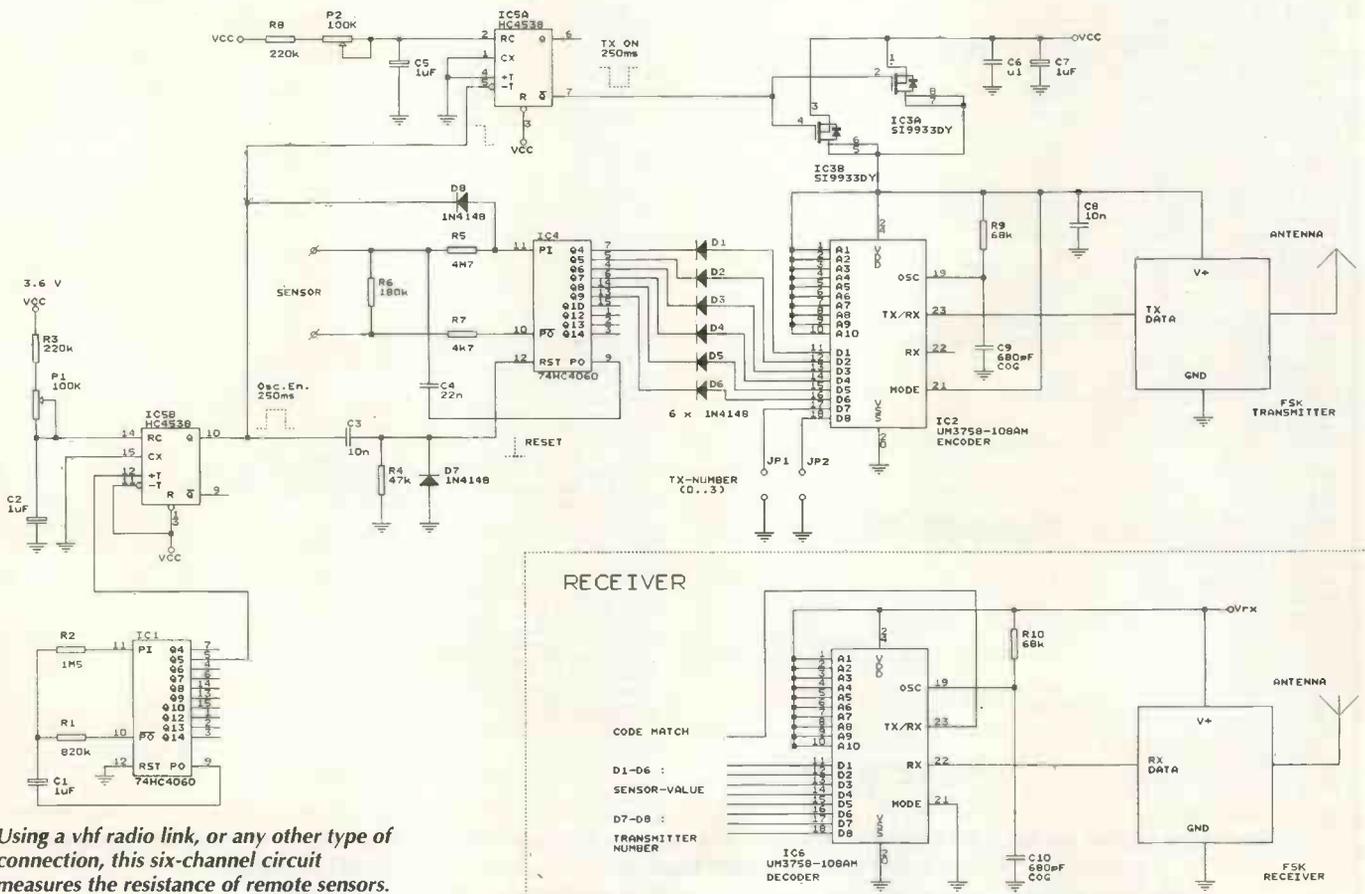
to use pull-up resistors.

Logic levels of data bits D7 and D8 are user selectable to allow the use of four transmitters on the same frequency without changing decoder address lines in the receiver. A small difference in transmitting interval will prevent most of the interference when more than one transmitter is used.

Powered by a 3.6V lithium cell, the transmitter logic draws less than 100µA, transmitting current depending upon the type of transmitter used, although overall power consumption is still low because of the short transmitting time. A telemetry system based on this design has been in use by our institute since early 1994 to study activity and feeding behaviour of the red deer. As the resistive element we use an electrolytic tilt sensor. I built the circuit using surface-mounted components but, except for the SI9933DY, they are also available in through-hole form. **Willem van der Veer**
Institute for Forestry and Nature Research Wageningen
The Netherlands.

References

UM3758 series data sheet, UMC Europe, Amsterdam, The Netherlands. Telephone: 0031-20-6970766.
Little Foot series manual, Siliconix Ltd, Newbury, Berks RG14 5UX. Tel. 01344-485757.



Using a vhf radio link, or any other type of connection, this six-channel circuit measures the resistance of remote sensors.

COMPUTER ICs

TMS 9900NL-40 PULLS	£20 ea
S9900 NEW AMD EQUIVALENT	£30 ea
MC6802 PROCESSOR	£2 ea
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MK48202-20 ZERO POWER RAM EQUIV 6116LP	£4
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LM323K 5V 3A METAL	£3
LM350K (VARIABLE 3A)	£3
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LM317H TO5 CAN	£1
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14M3818 16M00 17M625600 18M00000 18M432 19M050 19M2	
19M440 20M000 20M0150 21M676 22M1184 23M587 24M0000	
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CRYSTALS

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21M400M15A 24M000 25M000 26M995 BN 27M045 RD 27M095 OR	
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100n 600V Sprague axial 10/£1	100/£5 (£1)
2µ2 160V rad 22mm, 2µ2 100V rad 15mm	100/£10
10n/33n/47n 250V AC x rated 15mm	10/£1
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1µ 100V rad 15mm, 1µ 22mm rad	100/£6
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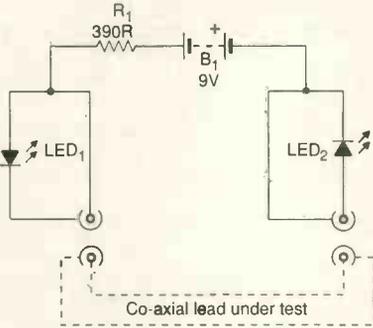
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100n ax long leads	100/£5
100n 50V oil package 0.3	

They don't come much simpler than this one. Three electrical components test a coax. cable for shorts and continuity, provided you have remembered to isolate the sockets.



Short and continuity test for coaxial cable

With three components, you can test a coaxial cable for inner and outer continuity and short circuits between them. Led 1 lights for inner continuity and led 2 when the outer is in one piece. If there is a short, neither led lights, since they are both across the

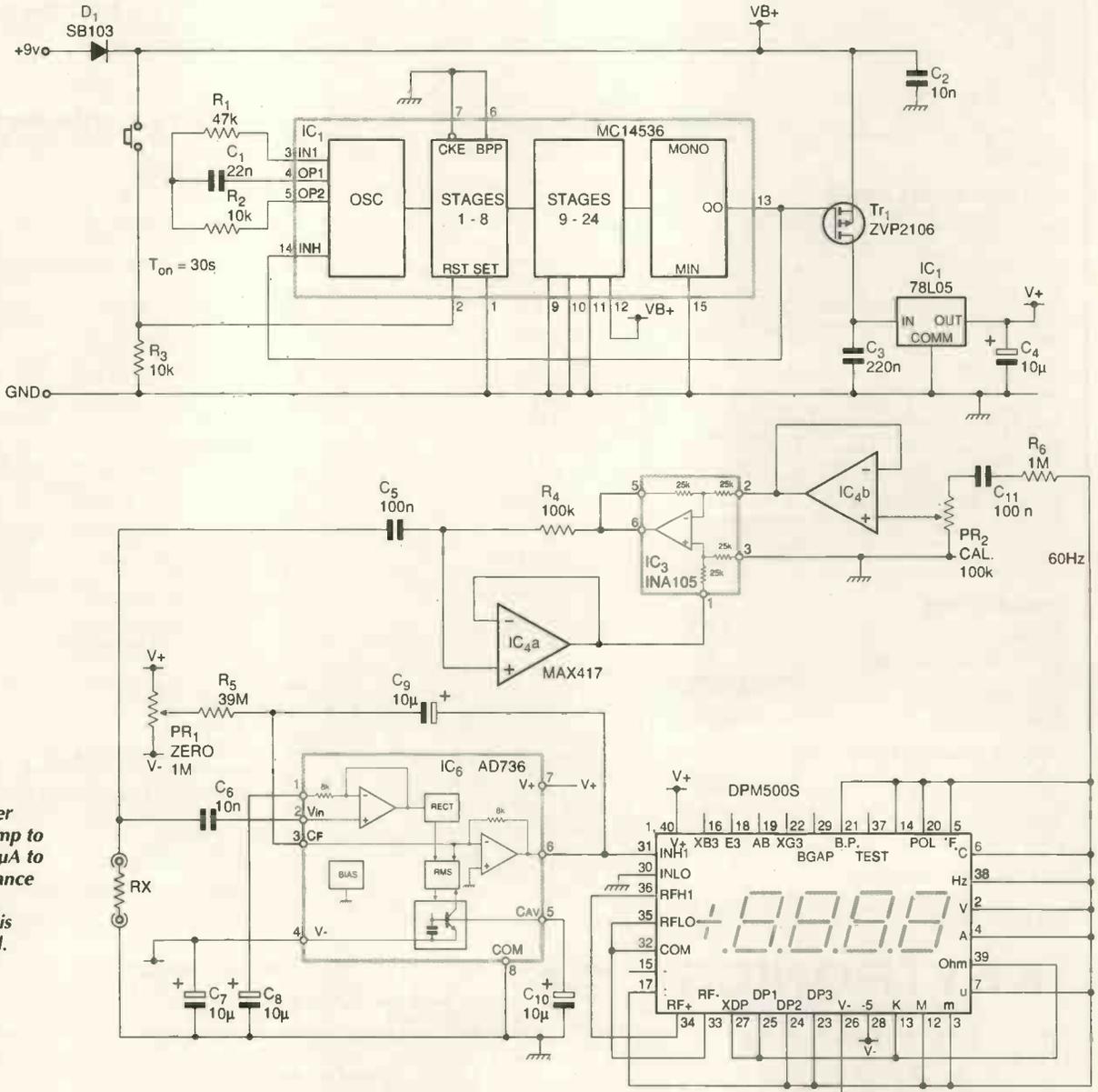
cable, short-circuit current being limited by R_1 . To resolve the ambiguity of both leds being out for shorts and open circuits in inner and outer, disconnect the cable and touch the outs to the socket shells. **S Roberts**
Bude, Cornwall

Measuring conductivity

To measure the contact resistance of physiological electrodes, current density must be kept below the threshold of feeling and be low enough to avoid polarising the electrodes.

In the instrument shown here, the reference current comes from the Howland current pump based on the *INA105*, which takes its 60Hz input from the display backplane drive, PR_2 setting the output current at $1\mu A$.

Voltage across the electrodes (R_X) is rectified by the *AD736* and displayed. Measuring range is $0.1k\Omega$ to $199.9k\Omega$. **A G Birkett**
London SE22



Conductivity meter uses a current pump to supply constant $1\mu A$ to the contact resistance to be measured. Resulting voltage is digitally displayed.

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ECC35	7.50	Z80	3.50	UCH81	2.50	6C5	3.50	12BE6	2.50
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ECL80	1.50	OB2	2.70	ZD21	3.50	6F6	3.50	812A	65.00
ECL82	3.00	CC3	2.50	3828	15.00	6FQ7	17.50	813	27.50
ECL83	3.00	OD3	2.50	4C250B STC	55.00	6GK6	4.00	833A	85.00
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EF37A	2.50	PCF86	2.50	SV4G	4.00	6J5	3.00	931A	25.00
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EF85	1.50	PCL85	2.50	6AL5	1.00	6K6GT	3.00	6080	7.50
EF86	10.00	PCL86	2.50	6AM6	2.00	6K7	4.00	6146B GE	25.00
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EL81	5.00	PL802	6.00	6B7	4.00	6SC7	2.00		
EL84	2.25	PY81	1.50	6B8	4.00	6SG7	3.00		
EL84 Mull	6.00	PY88	2.00	6BA6	1.50	6S7	3.00		

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300	34.32	24.02	18.09	13.38	12.98	12.58
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625	53.09	41.36	31.14	23.02	21.24	20.57
750	58.39	44.23	33.30	24.62	23.89	23.17
1000	78.80	55.16	41.54	30.70	29.80	28.89
1200	82.45	57.72	43.46	32.12	31.17	30.23
1500	105.10	73.63	55.40	40.94	39.74	38.53
2000	114.45	96.13	72.39	53.51	51.93	50.36
2500	163.04	114.13	85.94	63.51	61.64	59.79

These prices are for 240volt primary and two equal secondaries with 8" colour coded fly leads. Each transformer is supplied with a mounting kit (steel dish washer pads, nut and bolt) Mail order prices include vat and postage. Please do not hesitate to telephone or write with your particular requirements.



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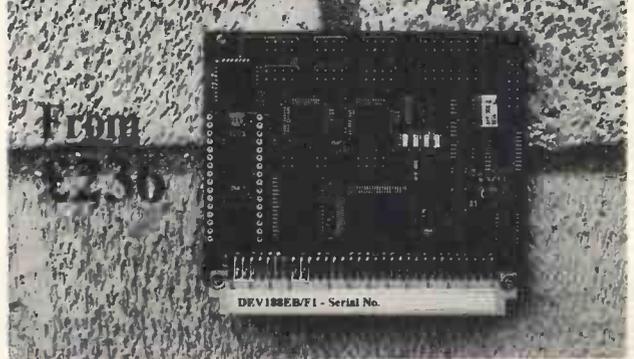
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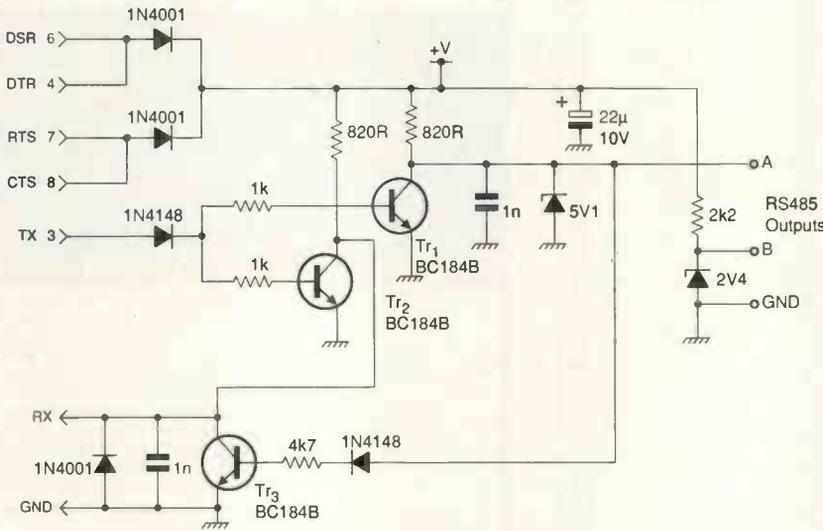
RS232-to-RS485 PC-powered converter

needed to control settings on machines, which have RS485 ports, by means of a laptop PC, which has

RS232. Since no power supply was available, it had to come from the serial port but, since there is

insufficient power on most serial ports to supply an RS485 driver I developed this circuit.

This simple circuit enables a PC RS232 serial port to control machines with RS485 drivers, supplying the power from the PC port.



Transistor Tr_1 inverts the signal on the RS232 TX pin, buffering it from the A line. B is at around 2.4V, held by the zener, since the RS485 is differential. Transistor Tr_3 is a level converter to RS232 and Tr_2 prevents data from echoing to the RX pin when the PC transmits.

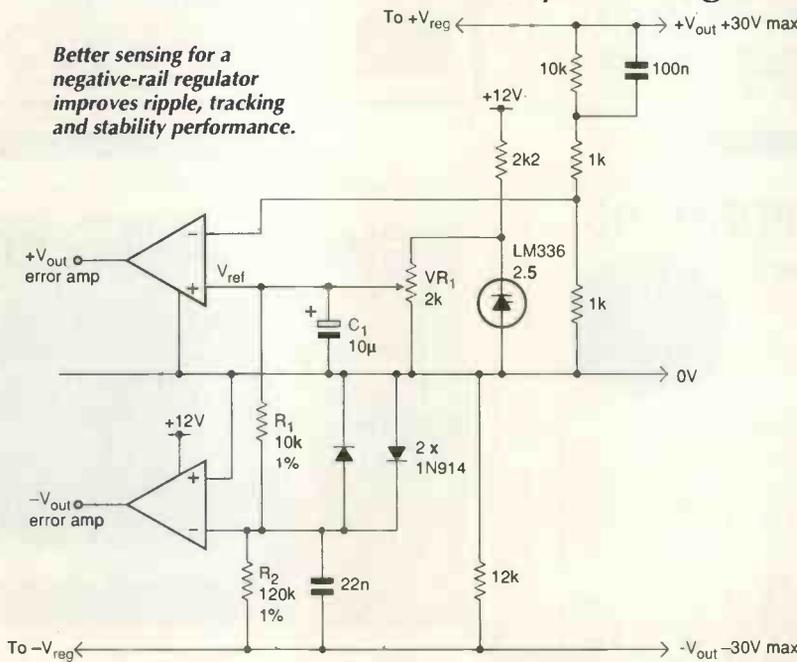
Power from DTR and CTS is smoothed by the 22µF capacitor and the 1nF capacitors reduce noise, where this is necessary.

Eight controls can be daisy-chained at 19200baud, if MAX487 low-power RS485 drivers are used; with 75ALS176 drivers it still works at 19200baud on three controls. In fact, it was intended for only one driver.

K Guy Wilkinson
Bromsgrove, Worcester

Sense, but not sensibility for negative rails

Better sensing for a negative-rail regulator improves ripple, tracking and stability performance.



A small modification to the standard negative half of a dual power supply regulator improves ripple performance and tracking.

Normally, the sensing voltage for the negative regulator comes from a resistive chain strung between positive and negative rails. Here, it is returned via R_1 to the common reference point. With $R_{1,2}$ scaled suitably, an adjustment of the potentiometer between 0V and 2.5V gives a V_{out} change from 0V to 30V.

With this circuit, the negative output showed increased stability with reference to both 0V and the positive rail and, on a 2A load, ripple is less than 0.5mV and a 1kHz switching waveform came out square. Tracking between the two rails is within 50mV. The filter formed by R_1 and C_1 removes any chance of ripple being injected into the reference from the negative rail.

Gregory Freeman
Nairne, South Australia

No-compromise oscilloscope probe

In most situations, the standard 10:1 oscilloscope probe, equalised by a few picofarads, works well but, when looking at

fast waveforms, it can distort the picture, as shown in the bottom trace on a 20ns pulse.

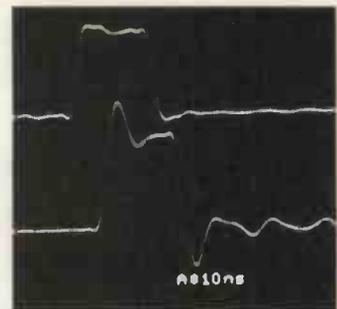
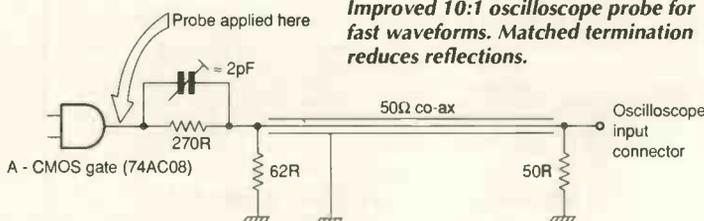
For this kind of work, a

properly matched transmission line is the answer; anything else produces reflections.

Part of the circuit is a resistive pad, still dividing the input by a factor of ten, but terminating in 50Ω to absorb reflections; the 270Ω resistor swamps the gate output impedance. The top trace shows the result.

Nick Wheeler
Sutton, Surrey

Improved 10:1 oscilloscope probe for fast waveforms. Matched termination reduces reflections.



Compensated pad, top trace, shows much improved performance, over standard oscilloscope probe. Extra 8µs delay on standard probe is due to greater length of coax.

New Mini Camera & Special Offers

New mini waterproof TV camera 40x40x15mm requires 10 to 16 volts at 120mA with composite video output (to feed into a video or a TV with a SCART plug) it has a high resolution of 450 TV lines vertical and 380 lines horizontal, electronic auto iris for nearly dark (1 Lux) to bright sunlight operation and a pinhole lens with a 92 degree field of view, it focuses down to a few CM. It is fitted with a 3 wire lead (12v in gnd and video out)..... £89.50 + VAT = £109.95 or 10+ £89.52+VAT = £104.95

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Power interface 4A kit..... £36.00

Power interface 8A kit..... £46.00

Stepper kit 4 (manual control) includes 200 step stepping motor and control circuit..... £23.90

Hand held transistor analyser it tells you which lead is the base, the collector and emitter and if it is NPN or PNP of faulty..... £35.45

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1.5D 3mm or 5mm red or green..... 7p each yellow 11p

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Standard charger charges 4 AA cells in 5 hours or 4Cs or Ds in 12-14 hours * 1XPP3 (1, 2, 3 or 4 cells may be charged at a time)..... £5.95

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7 segment common anode led display 12mm..... £0.45

LM557A TO3 case variable regulator..... £1.95

GaAs FET low leakage current S8875..... £12.95 each

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BC547A transistor..... 20 for £1.00

74LS05 hex inverter..... £10.00 per 100

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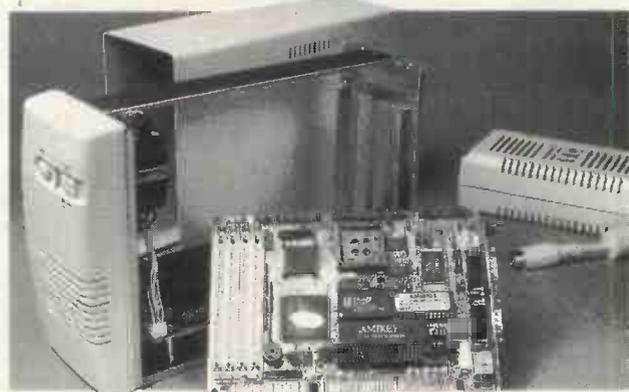
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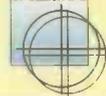
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CIRCLE NO. 135 ON REPLY CARD

High-power, isolated switch

This switching circuit is controlled by logic-level input, handles high power and the output is isolated from the control circuit.

Optical isolation is normal, but a separate supply would have been needed for the mosfet drive; for that reason, a transformer provides the isolation, since drive is now provided by the control circuit.

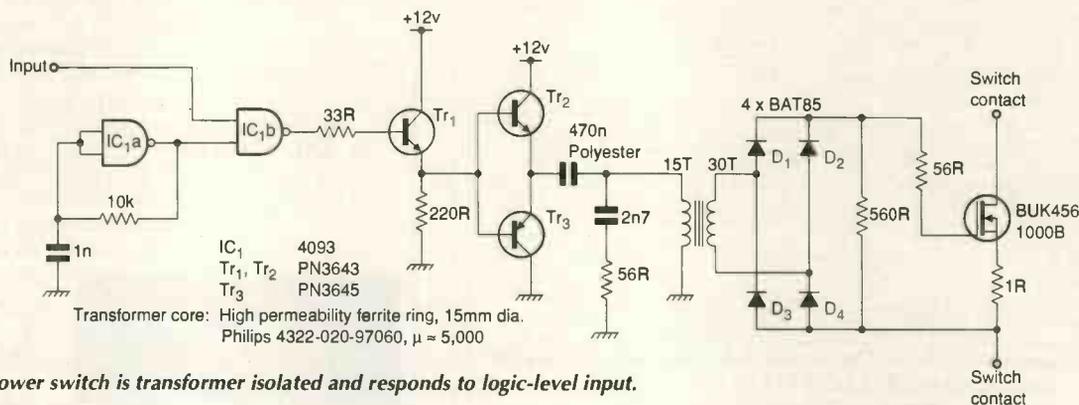
Output from the oscillator $IC_{1(a)}$ is

gated by $IC_{1(b)}$ in response to the input. Transistor Tr_1 and the output stage $Tr_{2,3}$ drive the transformer, which is on a small ferrite toroid, its secondary providing the mosfet drive, after rectification by the schottky diodes, which have a low forward voltage. The 560Ω resistor presents a low source impedance for rapid mosfet switching.

Keep the transformer primary and

secondary well separated for a high breakdown voltage; leakage inductance is unimportant, since losses can be offset by an increased turns ratio.

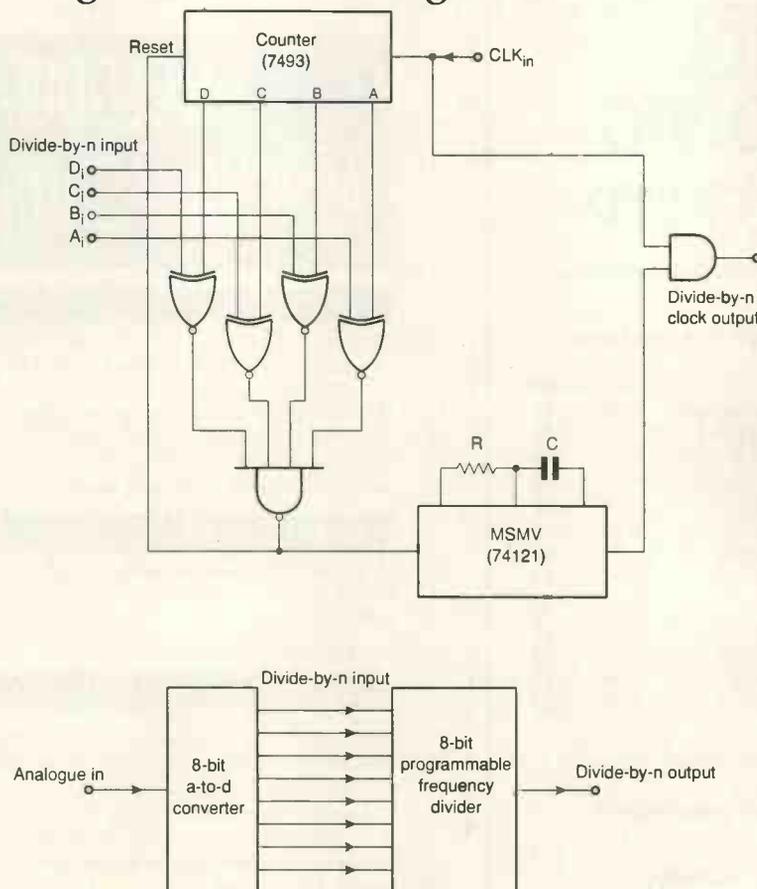
Phil Denniss
 Department of Plasma Physics
 University of Sydney
 NSW
 Australia



Power switch is transformer isolated and responds to logic-level input.

Programmable voltage-to-time converter

Programmable frequency divider used to form a voltage-to-time converter, in which output period represents input voltage.



As a step towards a v-to-t converter, Fig. 1 shows a programmed frequency divider that produces an output frequency of f_{in} divided by the binary input, also generating a repetitive output sequence depending on the binary input.

The circuit shows a four-bit circuit. Clock signals drive the counter, whose output at some point coincides with the binary input, coincidence being detected by the gating circuit, which produces a rising pulse edge to trigger the 74121 multivibrator for a wider pulse. Timing components are chosen to cover the duration of the clock period. Output, the clock frequency divided by the binary input, consists of the Anded clock and multivibrator output.

To form a voltage-to-time converter, use an 8-bit circuit of the form in Fig. 1, as shown in Fig. 2, where the 8-bit analogue-to-digital converter provides the binary input to the frequency divider. Output is the period of the divided clock frequency, which corresponds to the input voltage of the a-to-d converter.

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CIRCLE NO. 136 ON REPLY CARD

Coaxial cable choices

There's a variety of coaxial cable and connector types available. Nick Wheeler discusses how to choose the right combination.

The simple, flexible, laboratory connector uses one of a small number of cable types. Some 49 types are listed in the American Radio Relay League (ARRL) handbook. These vary widely in overall diameter and performance. This article is intended to help in selecting the right type of cable, together with appropriate end connectors.

General considerations

While a coaxial connector assembly has much in common with a good-quality screened audio connector – and is often used for this purpose – it is designed for use as a radio-frequency transmission line. Its characteristic impedance, Z_0 , is proportional to $\log(b/a)$, where b is the inner diameter of the outer conductor and a is the outer diameter of the inner conductor. If a coaxial connector assembly is driven from a source whose output impedance is Z_0 and terminated in Z_0 , then the input sig-

nal will appear almost unchanged at the termination.

There will be a slight reduction in amplitude due to losses, and the signal will be delayed by the time it takes to traverse the length of the line. This is longer than the time taken to traverse the length of the line in free space. Thus in free space the velocity of propagation is some 30cm/ns whereas in typical lines it is 20-27cm/ns.

Coaxial structures

The centre conductor of a coaxial cable may be solid or stranded. For anything other than permanent fixed installations such as television down leads, stranded types are essential. All other things being equal, a solid conductor will be slightly less lossy than a stranded one.

Air is the ideal insulator, but this can only be achieved over short lengths. Next best is air-spaced PE (polyethylene). Suitably spaced PE

Table 1. Velocity factors of common coaxial cable insulators

Air space PE washers	0.89
Air space PE voids	0.86
Foam PE	0.79-0.80
Solid PTFE	0.7
Solid PE	0.66

washers keep the central conductor in position.

In a different form of air spacing, the insulator has longitudinal voids so that in cross section it looks like a spoked wheel. This is very nearly as good as the washer approach and is much cheaper to make. It is commonly used for satellite down leads, and also for tv down leads.

Down leads used for reception usually have a Z_0 of 75 Ω . This can be shown to be less lossy than otherwise physically similar 50 Ω cable. The lower impedance cable has better power handling capability and is nearly always used for transmitting and receiving applications. It is also easier to match 75 Ω cable to typical antennas.

Next in order of merit is foam PE, followed by PTFE and finally the cheapest – solid PE. Velocity factors associated with these materials are shown in Table 1.

There are three types of outer conductor. The first is a solid metal tube, invariably of copper, and in large sizes over 1cm, corrugated. Clearly such cables are not designed to be flexed repeatedly.

The next – and the most common – type uses copper braid, sheathed with PVC to minimise corrosion and damage.

Braided sheath is not completely leak-proof. Small amounts of power can escape and be radiated, constituting a loss and possibly an EMI problem. Also, strong incident radiation may permeate the cable, causing interference.

Cables are available which include a layer of copper foil wound without gaps between the braid and the insulator. These can achieve sig-

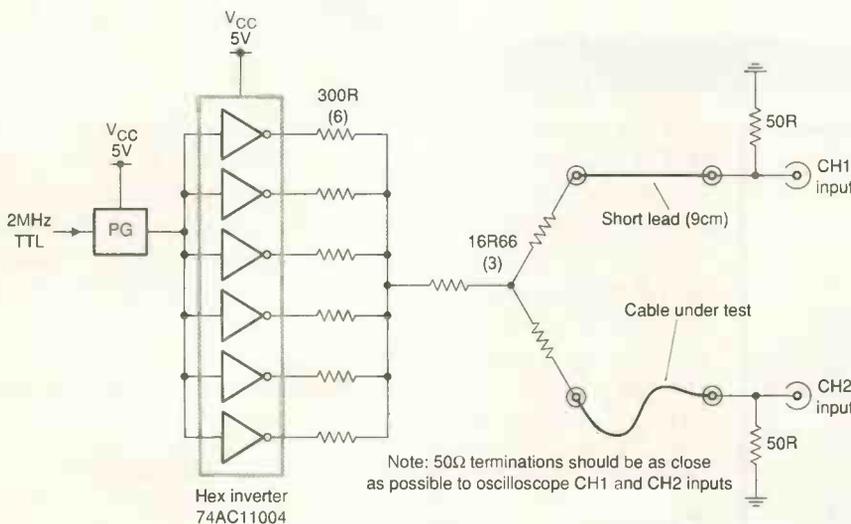


Fig. 1. Simple circuit providing a reliable 50 Ω , 10ns pulse source for assessing the performance of a transmission line. Connecting a 100MHz oscilloscope provides good test results.

nificantly better performance than braid-only cables – but at considerably higher cost.

Limitations of coaxial cable

Coaxial cable has an upper limit of usable frequency other than that imposed by insulator losses. This is due to higher order propagation modes, analogous to those found in waveguides. These set in at wavelengths shorter than

$$\lambda = \pi(a+b)/(2\sqrt{K}) \quad (\text{Ref. 2})$$

Where K is the dielectric constant of the insulator.

Applying this, approximately, to RG58 cable, which is commonly used for laboratory hook-up connectors, gives,

$$\begin{aligned} a &= 1\text{mm} \\ b &= 2.92\text{mm} \\ K &= 2.3 \text{ for PE} \end{aligned}$$

As a result, λ is 4.06mm, or about 65GHz. It is unlikely that this phenomenon will be encountered in ordinary laboratory work. Note that in some electronic warfare search receivers, this may be a frequency of interest. Applying this formula to PTFE-insulated RG405 cable yields a slightly higher frequency, and is specified to 18GHz.

There is another limitation, more likely to be encountered, but only in connection with transmitters or other high power rf sources. As we are dealing with transmission lines, mismatch will produce current and voltage nodes and antinodes along the line. Under extreme conditions these can lead to voltage breakdown or local overheating.

An open circuited line is just as liable to cause damage as a short-circuited one.

Connectors for coaxial cables

Three main professional series of connectors are SMA, BNC and N-Type. Electrically, TNC is a close equivalent to BNC. The SMB and SMC series are miniature SMA types, but note that SMB connectors are not positively retained. A circular spring provides a weakly retained push-fit action, useful for laboratory work.

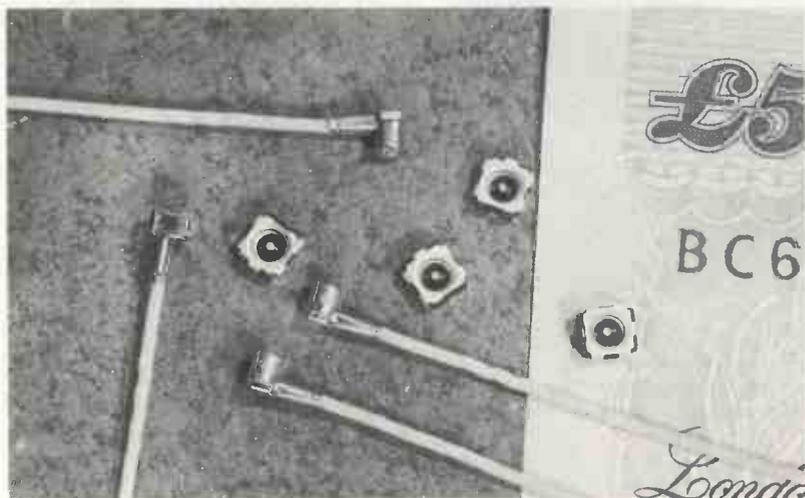
Ideally, the following examples should not be chosen for rf work.

- UHF connectors: poor performance in all respects.
- Phono: as the name implies, these are intended for audio but often used for video.
- Television down-lead types: these are not a good choice. Most rely on a solderless connection between the centre conductor and the tubular centre pin.

There is a comprehensive range of 'Inter-series adaptors'. These allow you to access virtually every type from BNC.

The full range of BNC accessories and connectors is only available in 50Ω form, but the more commonly used parts also come in 75Ω form. These look identical, except that the 50Ω plug and socket is slightly larger than the

World's smallest surface-mount coaxial connectors



Designed for use in communications applications, these ultra-miniature low-cost SC connectors require only 11.5mm² of board space and stand no more than 3mm high when connected.

These 50Ω Murata connectors accept two coaxial cable diameters of 0.8 or 1.2mm and their maximum vswr is 1.2 in the DC to 3GHz range. Maximum contact resistance is 15mΩ while minimum insulation resistance is 500MΩ.

75Ω alternative. Note that mixing connectors of different impedances can cause damage.

In addition to the wide range of adapters, there are Z₀ terminators. These can be either 'stop end' or 'through' types. The latter can save a lot of trouble when terminating at, say, an oscilloscope input. In addition 'through' attenuators are available.

Crimped cable attachments are quick and simple. Provided a cable strain relief is used these terminations are also durable. Clamp types impose only gentle forces on the outer braid and a good life is obtainable without the need for a strain relief. If a fault does occur, the connector can be disassembled and remade.

The most common clamp and crimp connectors suit 5mm cable outside diameter. This should be regarded as the default size. Stripping coaxial cable is a difficult process but special tools are available.

Crimp tools cost upwards of £70. Such tools form the sleeve into a uniform hexagonal shape. Much cheaper tools designed for fitting spade-type connectors to power cables will not suffice.

N-Type connectors appears on equipment intended for use above 1GHz. The Type 1 can only be used with URM67 or RG 213U cable. They are both reasonably flexible at 10.3mm outside diameter.

For laboratory bench use the Type 2 free plug can be used with 5mm URM43 or URM 76. These parts are specially made to introduce the minimum possible Z₀ discontinuities. They are characterised up to 10GHz.

SMA connectors are physically small and are available in crimp or clamp form. They are also made for assembly by soldering to RG402 and RG405 semi-rigid cable. In this form of assembly they are rated up to 18GHz.

Much of the circuitry of professional equipment operating above 1GHz consists of modules. These modules are often from different manufacturers, neatly plumbed together with semi-rigid coaxial cable and terminated with SMA connectors. The small size of these connectors often justifies their use in lower-frequency applications.

There are several useful parts in the SMA, series intended for direct mounting on pcbs. These can be used to provide detachable links to front-panel mounted connectors.

Testing for performance

In my view, there is no better way of testing a transmission line than feeding it with a pulse sourced from the correct Z₀ and observing the effect across a correct termination at the other end. A 100MHz or better oscilloscope is advisable to avoid measurement errors.

There are many sources of suitable pulses. I used a variant of the circuit on page 61 of *EW*, Jan. 96. This is the box marked PG on Fig. 1.

A logic-one pulse of less than 10ns duration in response to ttl drive is needed. A problem here is that a single logic gate cannot drive a 50Ω line directly. The device used must be able to sink or source 100mA.

In Fig. 1, PG produces logic-one pulses of about 10ns duration at a pulse-repetition frequency of 2MHz. These are applied to all six

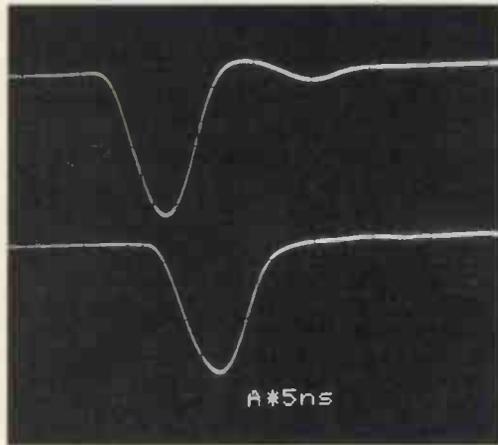


Fig. 2. Test results for a length of PTFE insulated miniature coaxial cable. Upper trace, input signal of approximately 1.2V. Lower trace, showing acceptable results: a loss of 0.37dB and delay of 5ns.

inputs of a hex inverter. The outputs are all paralleled via 300Ω chip resistors.

Making the assumption that the output impedance of a gate in the one or zero state is zero, we then have a 50Ω source. A possible part is 74AC11004.

A 74AC004 will also suffice. If it is loaded with 50Ω, the transmitted signal will be inverted with a peak amplitude of $V_{cc}/2$. This is passed down a short length of 50Ω line to a resistive splitter comprising a star connection

of three 16.666Ω resistors (3×50Ω in parallel). One of the two outputs feeds an oscilloscope input in parallel with a 50Ω termination. The second port goes to the connector assembly under test. The other end of the assembly under test feeds the other oscilloscope input, where it is properly terminated with 50Ω.

Figure 2 is the result of such a test, on an 87cm length of PTFE insulated miniature coaxial cable terminated with SMA 'free' plugs. The upper trace is the input and is of a peak amplitude of 1.2V. This result is close to $V_{cc}/4$, accounted for by the effect of the paralleled matching resistors and the splitter.

The output has a peak amplitude of 1.1V, representing a loss of 0.37dB, or 4.3dB/10m. This falls between the 3.6dB/10m at 100MHz and 19dB/10m at 1000MHz quoted for this cable, which seems reasonable.

There is a delay – estimated at 0.4ns – between the output of the splitter and the channel-one input of the oscilloscope. This must be added to the 5ns delay scaled from the oscillogram.

Making the necessary corrections leads to a velocity factor of 0.53, which is too low for the known figure of 0.7 for PTFE. However, it is near and making measurements to an accuracy of 1ns with a 100MHz oscilloscope is questionable anyway.

In any case the cable is clearly fit for laboratory use. Undulations in the upper trace are obviously due to reflections. The amplitude of these undulations is approximately 0.1 that of

the main pulse, suggesting a standing wave ratio of about 1:1.2, which is normally insignificant.

In summary

Although there is a bewildering array of cables, plugs and sockets available, new designs should be based on the SMA, BNC and N-Type connectors. As for the cable, this should be RG174A or RG316/U, which suit SMA and appropriate BNC types.

Unless compactness is important, then URM76 or the slightly cheaper RG58C, 5mm outside diameter, are the cables to choose. These suit BNC clamp and crimp connectors and also N-Type, type 2 free plugs. All other N-Type free parts require the use of URM 67 or RG213U. Most other cable sizes upwards of 5mm outside diameter can be terminated in BNC, but none of these has any pronounced advantage over those already mentioned.

There are 75Ω cables specifically intended for television and satellite down-lead applications. A limited range of BNC plugs and sockets are available in 75Ω form. Almost all 75Ω cables are solid-core, intended for one-off installation.

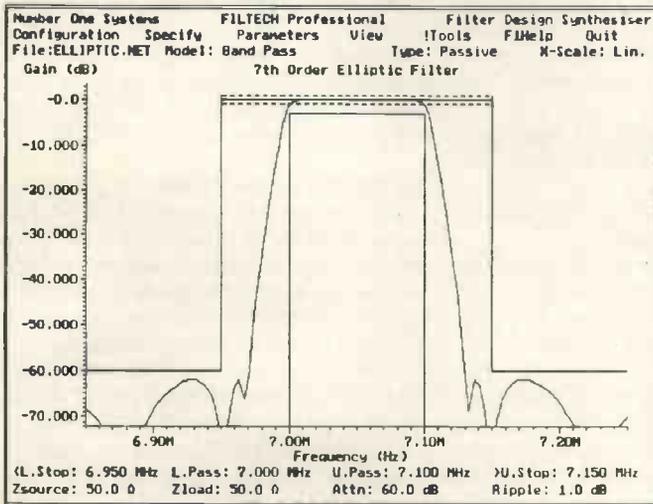
Further reading

American Radio Relay Handbook. (A good source on this subject, held in the reference section of most large libraries)

Services Textbook of Radio, Vol 5 (Long out of print, but good on line theory).

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ACTIVE

Discrete active devices

High-voltage igbt. The first device to be available from Motorola's non-punchthrough, high-voltage, insulated gate bipolar transistor family will be the *MHPM1A1200A120C5*, which will cost about \$1000 in small quantities. The family will contain devices rated at up to 1200A at 1200V and is intended for motor drives, power conversion and welding. The package consists of a copper baseplate with multiple copper substrates; since these devices exhibit a positive temperature coefficient, they can be paralleled for higher currents, so that a number of modules can be produced using only a few, smaller, die sizes. Motorola Inc. Tel., 001 602 244-3831; fax, 001 602 244-6002.

Digital signal processors

Audio processor. *TDA1548T* from Philips is an audio processor incorporating Bitstream filters for digital de-emphasis, volume and tone control, with Bitstream d-to-a and headphone amplifiers, providing a single-chip solution to all the functions needed for CD, MD and DCC personal stereo players. It is compatible with I²S or I²Sb-justified serial input data and puts out a 1.7Vpk-pk audio signal into 32Ω. Gothic Crellon Ltd. Tel., 01734 788878; fax, 01734 776095.

Electronic trimmers. AKM *AK9813/9844* are single-chip electronic trimmers for voltage control input, replacing mechanical types or eeprom/multiple d-to-a circuitry. 9813 has 12 channels (9844 four), control being effected by a simple microcontroller. Output range is ground to supply voltage and there is a power-down mode. DIP International Ltd. Tel., 01223 462244; fax, 01223 467316.

Linear integrated circuits

Video d-to-a. *SAA7167 YUV-to-RGB* video digital-to-analogue converter is an addition to Philips's Desk-Top-Video chipset and converts digital video to analogue rgb, mixing the output with an external analogue rgb channel to merge digital video sources with standard VGA graphics, thereby doing away with the need to access the computer's frame store memory. Together with the *SAA7131* FMPEG decoder the new chip allows

the production of cheap video playback cards for Pcs. In addition, the chip will handle various input formats, on-chip colour keying between video and rgb inputs, I²C-bus control and direct drive to a monitor. Philips Semiconductors (Eindhoven). Tel., 00 31 40 722091; fax, 00 31 40 724825.

High-current driver. Driving high-speed signals into low-impedance loads, Analog's *AD815* driver has differential input and output, enabling it to replace multi-chip designs in asynchronous and high bit-rate digital subscriber links and other twisted-pair cable driver functions, as well as in video distribution. Output is 400mA at 40Vpk-pk differential and the device will deliver up to 1Apk. Analog Devices Ltd. Tel., 01932 266000; fax, 01932 247401.

Memory chips

Fast srams. EDI's snappily named *ED18F32128LPBAC* is a low-power, 55-100ns, 128 by 32 sram in a JEDEC 68-pin pack. It is pin-compatible with the *ED117F32128* flash unit, but has a battery backup and an industrial-temperature version. EDI (UK). Tel., 01276 472637; fax, 01276 473748.

Microprocessors and controllers

Low-cost, 64 bits. IDT's *Orion R4640* embedded microprocessor, available in 80MHz, 100MHz and 133MHz versions, is designed for 64-bit performance 'at a 32-bit price', having a 64-bit core and 32-bit interface. Its features include a two-way associative, 8Kb instruction cache, an 8Kb data cache, power saving, a static core and wait instructions for stand-by. Speed is measured at 175dhrystone. Associated chips are the *R4761* memory controller and *R4762* PCI bridge. Integrated Device Technology. Tel., 01372 363734; fax, 01372 378851.

6x86 100MHz processor. The *Cyrix 100MHz 6x86* superscalar, superpipelined processor is available for immediate delivery. It was formerly known as the M1 processor, is compatible with the x86 instruction set and shows a benchmark performance figure of 678 on the Norton System Information v8.0, said to be considerably in excess of that for the 133MHz Pentium. Flashpoint, the supplier, also offers a motherboard from DTK with up to 128Mb of ram, three PCI slots and five ISA slots. Flashpoint Technology. Tel., 01753 538715; fax, 01753 538733.



PASSIVE

Passive components

S-m electrolytics. Sanyo's *SM Series OS-CON* surface-mounted electrolytic capacitors are meant for automatic insertion and reflow soldering. They use an organic electrolyte, which affords a life expectancy of 220 years (!), a frequency range approaching that of film types, self regeneration after excessive loads, low esr and stable leakage current. Values are 1-150μF at 6.3-20V, the largest can size being 8.8/8.8/13mm. Inelco Ltd. Tel., 01734 810799; fax, 01734 810844.

Uhf chip capacitors. Syfer high-Q surface-mounted chip capacitors cover the 0.47pF-1nF range, a 10pF component in the 0805 size taking up 2mm by 1.2mm of board space, although rated at 100V and having a ±5% tolerance. Esr at 1GHz is 0.2Ω; Q at this frequency is 50 and rises to 200 at 100MHz. Flint Distribution. Tel., 01530 510333; fax, 01530 510275.

Accurate Al-foil capacitors. Resin-dipped aluminium-foil capacitors from Sang Jing are claimed to be the ideal replacement for tantalums, now experiencing a shortage, since they have tight specifications, have good temperature characteristics and are reliable. Leakage current is 0.01CV or 0.5μA, whichever is larger; working voltages are 6.3-50V; and values 0.1-

Optical devices

Blinking leds. Elcos offers surface-mounted leds that blink and four-pad multicolour types — the *CR* range, all on ceramic substrates. Blinking versions contain a blink ic whose frequency is trimmable by resistance or capacitance. All types have a wide viewing angle and a flat-top package for use in light-pipe application. There are also the red/green/blue types which, when correctly driven, will display any colour, including white. Flint Distribution. Tel., 01530 510333; fax, 01530 510275.

470μF in tolerances of 10% or ±20%. Europa Components & Equipment plc. Tel., 0181-953 2379; fax, 0181-207 6646.

Precision ceramic capacitors. NPO multilayer ceramic capacitors offer very narrow tolerances for use in impedance matching in hf communications equipment. New production techniques have reduced the cost of these devices and this series provides tolerances of ±0.1pF in values of less than 2.7pF, with ±0.05pF available on request. They are suitable for wave or reflow soldering. Philips Components. Tel., 00 31 40 722790; fax, 00 31 40 724547.

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Cermet trimmer. BI Technologies introduces the *Model 23 Series* of 4mm surface-mounted, single-turn cermet trimmer pots in sealed packages, which withstand infrared or convection reflow processing and cleaning. T-cross slots are provided for adjustment and the devices come in both J-hook and gull-wing styles. Standard range is 10 Ω -2M Ω at 20% tolerance. BI Technologies Ltd. Tel., 01384 442393; fax, 01384 440252.

Audio products

Ceramic speaker. Meant for use in cellular telephones, Pedoka's *ST-20PR* piezoelectric speaker has a sensitivity of 104dB \pm 2dB at 1kHz, 1Vrms at the input and 1.6k Ω series resistance. It is 2.1mm thick and 21mm in diameter and has an HP series connector on 30mm of 32awg wire; other connectors can be supplied. Pedoka Ltd. Tel., 01462 422433; fax, 01462 422233.

Notebook audio. From IBM, the *PCMCIA Advanced Audio Adaptor*, a PCMCIA device to add good quality stereo recording and playback to portable computers with a Type II expansion port. It provides output to headphones or speakers, accepts joystick input and has an electret microphone for recording; it is compatible with most games that need Soundblaster support. DIP Systems. Tel., 01483 202070; fax, 01483 202023.



Hardware

Emc racks. Barton Engineering has a range of emc racks for best performance at 1GHz of 70dB. They have full-length, lockable metal doors with detachable hinge pins and four-point locking; and drilled air wave guides for up to four fans. These racks consist of only ten parts. Rfi seals are beryllium copper, low-compression, clip-on contacts on the doors and monel mesh on the sides. Finish is textured paint over nickel and the frame is nickel-plated. Barton Engineering and Export Ltd. Tel., 01227 272141; fax, 01227 771653.

Test and measurement

150MHz digital oscilloscope. Hitachi Denshi's newest digitising oscilloscope, the *VC-7524*, has a bandwidth of 150MHz and samples at 100Msample/s on each of four channels simultaneously. Memory is 2Mbyte per channel. Additions to the facilities of its forerunner, the *VC-7124*, are long and split memory, compressed display, 4000 times magnification, storage of 2000 triggered events, single-shot and slow-roll modes and glitch function to enable amplitude and time frame of an event to be measured. In the long memory mode, the whole 2Mbyte memory is compressed and displayed to a screen capacity of 1Kbyte. Other features, too numerous to mention here, are incorporated. Hitachi-Denshi (UK) Ltd. Tel., 0181 202 4311; fax, 0181 202 2451.

A/d multimeter. Two ranges of handheld multimeters by AVO have a dual display to give simultaneous viewing of volts/frequency, current/frequency, conductance and resistance. Two models in the *M7000* series are in tough cases with rubberised holsters for industrial use, one of them being of 0.2% accuracy and the other 0.3%, with a switchable choice of 4000 or 40000 counts. *M8000* models provide true rms dc+ac and ac voltage readings and both are accurate to within 0.08%. Capacitance and diode test are included on all models. Avo International Ltd. Tel., 01304 502101; fax, 01304 207342.

Digital handheld multimeter. Di-loG's *DL-2977* is a 400-count digital multimeter with a 43-segment bar graph and a X10 zoom. It autoranges to provide V, I, R, true rms, frequency, temperature, continuity and diode testing capability. Accuracy when reading voltage is within \pm 0.3% +1 digit. Facilities are offered for max/min measurement against stored references. Di-loG Ltd. Tel., 01707 375550; fax, 01707 393277.

Digital delays. An improved version of the *9650A Digital Delay Generator* is announced by EG&G. There is a



better communications link and a new panel with an alphanumeric display; a shut-down memory retains current settings. Trigger is internal or external to 2MHz and produces four independently variable delays from zero to 100ms and two difference outputs; pulse widths are 30ns-1ms. A "scan" mode produces increasing delays and a burst of pulses can be generated. EG&G Instruments Ltd. Tel., 01734 773003; fax, 01734 773493.

Thermal imager. *ThermaCAM* is a small, portable infrared imager for testing boards, components and assemblies. It uses a focal-plane array camera, has a 256 by 256 pixel display in colour, 12-bit digital storage in PCMCIA cards, colour viewfinder and interchangeable lenses. Video or TIFF files can be sorted and viewed inside the camera and the software allows Windows-based analysis. The camera array is maintained at a constant temperature by a microcooler. Inframetrics Infrared Systems Ltd. Tel., 01256 50533; fax, 01256 50534.

Transducer calibration. For the rapid calibration of transducers, HBM has the *DMCplus Digital Measuring Amplifier*, with which the input quantity and output signal are processed to present on a PC screen the calibration curve, using HBM software. The curve is printable and saved as archive. HBM United Kingdom Ltd. Tel., 0181-420 7170; fax, 0181-420 7336.

Flatbed recorders. Yokogawa's *LR120 Series* of flatbed recorders are available with one or two pens and only take up 320 by 323mm of bench space. Chart width is 200mm and recording speed is selected or programmed in the 1cm/h-60cm/min range. Input is 1mV fsd to 300V dc and versions are available for current measurement, ac and temperature measurement. Operation is Y/T or X/Y and an RS 232 interface allows remote control and data logging, a

Multi-purpose tools. Three very desirable SOG toolkits from Jensen. The *Paratool* is the top one, with pliers, wire-cutter, file, awl, knife blade, serrated blade, scale and a can-opener, all in one folding tool. The other two, *Toolclip* and *Micro Toolkit* are scaled-down versions with screwdrivers, but they all fold up very small. So far as we are aware, they are not associated with the Swiss Army. Jensen Tools. Tel., 0800 833246 (free); fax, 01604 785573.

software package giving viewing and analysis by PC. Martron Instruments Ltd. Tel., 01494 459200; fax, 01494 535002.

Literature

CPC. Combined Precision Components has produced a monthly product supplement and a weekly offer list of 20 pages containing details of reduced prices. The 1996 catalogue of 1600 pages is also available. Combined Precision Components plc. Tel., 01772 654455; fax, 01772 654466.

Electrospeed. Connectors are a major content of *Electrospeed's* new colour catalogue, including those from Molex, Cinch, Hirschmann, Thomas & Betts and Multi-Contact. Other increased sections are those for emc and filtering, batteries and chargers. Electrospeed. Tel., 01703 644555; fax, 01703 610282.

IDT. Company and product data from IDT now comes in a number of ways: CD-ROM; WWW; and fax. The CD uses Adobe Acrobat to display and print pages, including data sheets; the Web (<http://www.idt.com>) presents almost as much data and is updated daily, more information being also retrieved using anonymous ftp to <ftp.idt.com/docs/docid.ext>. Sales are on eurossales@idt.com. For

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Neanderthals not on line, there is Fax-on-Demand, offering all but manuals, which are too big. Call +1-408-492-8341. Integrated Device Technology. Tel., 01372 363734; fax, 01372 378851.

Bruel & Kjaer. A 9-page brochure from B&K introduces an 'integrated approach' to the measurement of sound and vibration and describes the company's capability in the software field, which enables the production of dedicated programs for specific application. Bruel & Kjaer (UK) Ltd. Tel., 0181 954 2366; fax, 0181 954 9504.

PIC handbook. A new edition of Microchip's *Embedded Control Handbook* for the PIC 16/17 family of field-programmable, eight-bit microcontrollers and memory is available. It includes over thirty new and revised application notes and software code for specific applications. There are also schematic and timing diagrams and maths routines. Arizona Microchip Technology Ltd. Tel., 01628 851077; fax, 01628 850259.

Snap switches. Cherry has a new publication describing the *DG* range of snap switches, which measure 12.8 by 5.8 by 6.5mm and switch up to 3A

Automatic cable cutter. Mil-Tech's *Autocutter* is a programmable, self-feeding cable, wire and extrusion cutter that takes material up to 100mm wide, including ribbon cable, thin wire, plastic extrusions and rubber mouldings, special sections being handled by custom versions of the machine. Requirements are a single-phase supply and compressed air at 80psi. Mil-Tech. Tel. and fax, 01477 571864.

at 125Vac or 2A at 30Vdc. Cherry Electrical Products Ltd. Tel., 01582 763100; fax, 01582 768883.

Fluke. The 250-page 1996 instrumentation catalogue from *Fluke*, is now available free. The company's products cover a very wide range of measuring instruments, latest additions being the Multi-product calibrator and a range of digital storage/analogue oscilloscopes in the Combiscope range. Fluke UK Ltd. Tel., 01923 240511; fax, 01923 225067.

Materials

Cleaning solution. Electrolube has a new solution, *Printasolve*, to remove ink from surfaces such as printer rollers, ink jets and franking machines. It is effective for solvent and oil-based inks, is economical in use and poses no threat to the ozone layer. Available in 400ml aerosols, it is sprayed onto a cloth or bud and applied to the surface, following which an Electrolube Air Duster helps drying. Electrolube Ltd. Tel., 01734 403014/031; fax, 01734 403084.

Production equipment

Solder-wave measurement. From Alpha Metals comes the *Solder Wave Optimiser*, a portable instrument to measure both board/wave data and temperature in wave soldering. Data shown includes dwell time, immersion depth, contact length and conveyor speed, preheat, maximum minus preheat temperature, maximum preheat slope and maximum slope over the wave. It also indicates wave-to-board parallelism, all data being down-loaded to a PC. The unit is simply passed through the soldering operation and collects all the data in one pass. Alpha Metals. Tel., 0181 6656666; fax, 0181 6654734.

Fluid dispenser. *I & J Fisnar* produces a finger switch fitted, by means of a Velcro band, to barrels,



cartridges and soft tubes, allowing precise fingertip fluid dispensing from the company's DSPE 501A pressure dispenser as an alternative to the more common foot switch. Intertronics Ltd. Tel., 01865 842842; fax, 01865 842172.

Pcb coordinate measurement. *Maxtscan 100* coordinate measuring machines by Graticules are claimed to reduce the time to check hole size and position by over 80%, or from three days to half a day. Since the system has a cad interface, raw data files may be used without editing or translation. With a bed size of 1m square, the machine handles two average board of 450 by 250mm, measuring the average 3000 dimensions in two hours to within 25µm accumulative and ±5µm repeatability. Associated software is Windows-based. Graticules Ltd. Tel., 01732 359061; fax, 01732 770217.

UV light source. The *Dymax Light-Welder 3010-EC* is a high-power, high-intensity ultraviolet light source for the curing of adhesives, coatings and encapsulants. Either manual (pedal) operation by timer and foot-pedal or automatic via interfacing to dial tables, turrets etc. is possible. Intertronics Ltd. Tel., 01865 842842; fax, 01865 842172.

Power supplies

Electron beam power. A new series of electron beam power supplies is introduced by AP&T; the first being the *Carrera V*, a 5kV (less than 1% ripple) type for electron beam deposition in film coating. The switched-mode design includes arc detection and recovery, which operates within 3ms, rapidly enough to leave constant emission control and melt unaffected. An arc rate monitor protects equipment in the vacuum chamber against multiple arcs. No water supply is needed, since the unit is air-cooled. Advanced Products and Technologies Ltd. Tel., 01865 724863; fax, 01865 725831.

Tactile keyswitch. *TACT ECO* from Secme is a low-cost tactile keyswitch taking up only 6mm square of the board and standing 4.3mm or 5mm off it. A moulded insert prevents flux buildup during soldering. Operation is momentary with a Sharon click for touch feedback; ratings are 12V and 50mA, with a contact resistance of less than 100mΩ. EAO-Highland Electronics Ltd. Tel., 01444 236000; fax, 01444 236641.

50W dc-to-dc. Features of Abbott's *NB* series of 50W dc-to-dc converters are a total size of 1.5 by 3 by 0.4in and 90% efficiency. Outputs of 2-28V are available from inputs of 14-40V and interfaces for paralleling, sync., enable/disable and 'power good' are included, different pin arrangements being optional. Abbott Electronics Ltd. Tel., 01233 623404; fax, 01233 641777.

LV dc-to-dc converter. From Semtech, the *SC1630CS* low-voltage, 120kHz step-up converter to drive external power switches in higher-voltage and power application. Only six further components are needed to form a step-down arrangement to obtain an 80% efficiency and a few more for a step-down type to give 86% efficiency at 2A and 300µA quiescent. Output voltage can be set internally to 5V or externally to an arbitrary value. Input is 7V, switch-off current 105µA, shut-down mode current 7µA, stabilisation 0.6% V_{out} and regulation 2.5% V_{out} . Semtech Ltd. Tel., 01592 773520; fax, 01592 774781.

Sla chargers. In both rack-mounted and cased versions, the *SM* family of chargers for sealed lead acid batteries can also be used as power supplies for other equipment, thereby being usable for powering pcs or floating 12V and 24V batteries up to 100Ah.



NEW PRODUCTS CLASSIFIED

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They are protected and can be used in series or in parallel to provide up to 250W, under 1% ripple and with a universal input in the 95-277Vac range. Electrospeed. Tel., 01703 644555; fax, 01703 610282.

Bench supply. CPX200 from Thurlby Thandar is meant for those who need 0-35V and 10A, but not necessarily at the same time. Output power of this dual supply is 175W per output, maximum current being adjusted in a switched-mode regulator as voltage decreases; 35V at 5A to 17.5V at 10A. Outputs work in constant-current or constant-voltage mode with auto crossover and mode indication. They can be connected in series or parallel. Thurlby Thandar Instruments Ltd. Tel., 01480 412451; fax, 01480 450409.

150W, rack-mounted. Weir has a plug-in, rack-mounted unit in a 3U by 14HP panel containing a three-output, 100W convection-cooled (150W force-cooled) supply, approved to all manner of safety and emc standards and working over the 90-254V input range with no tap changes. The HSS 103 is completely enclosed, has a panel-mounted IEC socket, switch and mains fuse. Weir Electronics Ltd. Tel., 01243 865991; fax, 01243 868613.

Power to the Pentium. Semtech's MP60 series of 35W switching dc-to-dc converters have a connector conforming to Intel's voltage regulator specification for the P6 processor. They supply 10A continuously at 2.9V and additional functions can include a 'Power good' signal, output enable and upgrade present, and programmable output voltage from 1.5V to 3.6V. Semtech Ltd. Tel., 01592 773520; fax, 01592 774781.

40W/in³ dc-to-dc. UPM converters from Amplicon Liveline use a synchronous rectifier buck regulator optimised for 5V inputs. Power and control electronics are on separate boards to obtain improved thermal performance and isolation of the control circuitry from heat sources. The devices are meant to provide 2.5-3.6V, fixed or programmable, from a 5V input at 40W/cubic inch power density and 90% peak efficiency. Amplicon Liveline Ltd. Tel., 0800 525 335 (free); fax, 01273 570215.

Radio communications products

Linear amplifiers. Pacific Amplifier Corporation's range of rf linear amplifiers covers the 1MHz-2GHz range and includes power amplifier sub-systems, power modules and low-cost lab. amplifiers. The PAC205 is a remotely controlled feedforward base station for the Special Mobile Radio band of 935-940MHz, while the PAC221 is a broad-band gain block for the 746MHz-930MHz cellular and telephony bands. Intermodulation is low at -60dBc and

-30dBc respectively, for output powers of 80W and 30W. Anglia Microwaves Ltd. Tel., 01277 630000; fax, 01277 631111.

Microwave amplifiers. Wessex has a new series of PCN and PCS microwave power amplifiers in modular form, in instrument cases or in rack-mounted versions. Bands covered are 1.805-1.880GHz or 1.930-1.990GHz with a gain of 29dB minimum (gain in band ± 1 dBpk-pk) to give 43dBm output power. They are unconditionally stable and protected against open or short-circuits and can be provided with cooling fans. Harmonics are 50dB down; spurious -70dBc. Wessex Electronics Ltd. Tel., 0117 9571404; fax 0117 9573843.

Protection devices

Transient suppression. For the surge protection of low-voltage semiconductors, from 2.8V to 4.5V, Semtech's four-layer enhanced punchthrough diodes are said to possess advantages over silicon avalanche types in having a leakage current of 10 μ A and capacitance of 50pF. ESD protection is to 15kV, peak pulse current is 30A and maximum clamping voltage at 1A 4.3V, 4.9V and 6.5V. Semtech Ltd. Tel., 01592 773520; fax, 01592 774781.

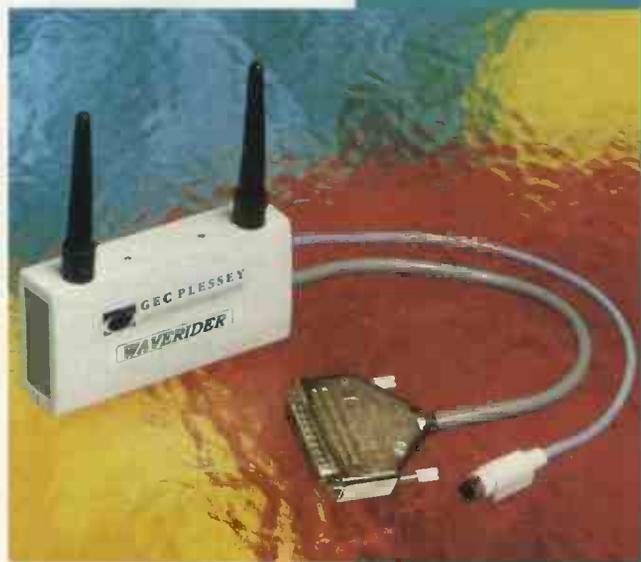
Snubber diodes. Fast-recovery, high-power diodes by IR are intended as snubbers for gate turn-off thyristors and possess a soft-recovery characteristic to avoid voltage spikes and ringing, which cause high power dissipation and misfiring of the switch. Devices in the SD103 N-RXX - SD453 N-RXX set are packaged as studs and there are others in disc form and as isolated modules. International Rectifier. Tel., 01883 713215; fax, 01883 714234.

Switches and relays

Hf relays. Matsushita's RK/RG high-frequency relays have either one or two changeover contacts and imposes an insertion loss of less than 0.3dB at 900MHz. Power consumption is 200mW - less when pulse driven - and can be made to latch with one or two coils. Size is 20.2/11.2mm, standing off 9.7mm. Matsushita Automation Controls Ltd. Tel., 01908 231555; fax, 01908 231599.

5GHz reed relays. Coto Wabash (US) offers a range of small, surface-mounted reeds handling 5GHz signals with less than 60ps rise times with less than 0.2dB insertion loss at 1GHz. Size is 9.3 by 4.6 by 4mm. Coil is for 5Vdc, is of 150 Ω resistance and 10¹⁰ Ω insulation resistance, giving a switching speed of under 0.5ms. Contact rating is 3W. Coto Europe. Tel., (Netherlands) 0031 45 320838; fax, 0031 45 320838.

Secure keyboards. Banging repeatedly with a hammer, while unusual as a method of testing



electronic equipment, was used to prove Lucas's new membrane switch panels, claimed to be 'vandal-proof'. Exposed areas such as bezel and key tops, are of stainless steel; below is an elastomeric pad to seal the membrane switch against cleaning fluids, moisture and dust. The sub-panel is of injection-moulded plastic or a laminate of pcb, aluminium and/or stainless steel. The type of hammer is unspecified. Lucas Control Systems Products. Tel., 01535 661144; fax, 01535 661174.

Photovoltaic relay. PVG612 by IR is a new member of its microelectronic photovoltaic relay family, switching up to 60Vac/dc at 1A (2A on dc) with only 5mA drive. It is a single-pole, normally open switch, the output Hexfet being optically isolated from the control circuit. International Rectifier. Tel., 01883 713215; fax, 01883 714234.

S-s relays. Solid-state relays in Crydom's PF range handle 10Arms at 40°C ambient. They are controlled by a logic-level signal and operate from 280-660V mains, being optically isolated to 4kV, VDE certified to EN 60606-1 with UL approval. Inrush current can be as high as 250A at a dv/dt of 500V/ μ s. Crydom Europe. Tel., 0181 763 0550; fax, 0181 763 0499.

Transducers and sensors

Absolute shaft sensor. Control Transducers has a range of absolute measuring sensors, which are effectively industrial rotary potentiometers using the *Mystar* plastic material for long life and reliability. A three-wire connection to the controller is needed. Design is such that axial and radial shaft play is taken up without effect on accuracy or life. Control Transducers. Tel., 01234 217704; fax, 01234 217083.

Waverider evaluation. GEC Plessey's DE6003 digital radio transceiver is now supported by *Waverider* a DE6003 evaluation kit, which consists of two transceivers, a user guide and an installation disk. It connects to a laptop or desktop computer via the parallel port and includes device drivers to allow use with Windows for Workgroups, Novell Netware or any TCP/IP networking software. The result is a 2.4GHz frequency-hopping transceiver with a range of 200 feet in a typical office, offering a 625kb/s data rate. Gothic Crellon Ltd. Tel., 01734 788878; fax, 01734 776095.

COMPUTER

Computer board-level products

VGA module. Datasound Laboratories has a mezzanine module for the *Chameleon* credit-card-size embedded PC The CLIO-200 uses stack-through connection to the host processor and provides 1024/768/16 or 800/600/256 pixel resolution for CRTs and also support for LCDs, EI plasma panels and colour STN, TFT, LCD, EI or plasma flat panels. Support for special types of display can be provided. Datasound Laboratories Ltd. Tel., 01462 675530; fax, 01462 482461.

Fast analogue input. From Pentland Systems, the MPX203 40Msamples analogue input board, giving 8-bit resolution on a ± 1 V analogue input. The board is configurable in software and supports both scanner and continuous acquisition modes.

Outputs from the on-board fifo and converter are accessible at the VME interface. Pentland Systems Ltd. Tel., 01506 464666; fax, 01506 463030.

Computers

Rack-mount computer. AMC offers the *AMC614* computer, which is intended to be part of ISA, PCI or EISA systems in telecomms or industrial use, taking a choice of PC/AT compatible backplanes including a 14-slot ISA, a 14-slot EISA or combination ISA/PCI or EISA/PCI types. Advanced Modular Computers Ltd. Tel., 01753 580660; fax, 01753 580653.

Industrial workstation. *AWS-822* from IMS is an industrial computer with a 14in colour crt display and sealed front-panel membrane keyboard. It has eight 16-bit ISA slots and comes with 286, 386, 486 or Pentium processor, is adequately fan-cooled and has a disk mounting assembly for floppy and 1in high hard drives. Integrated Measurement Systems Ltd. Tel., 01703 771143; fax, 01703 704301.

Portable workstations. Logic Instrument S.A. introduces the *Tetra* range of portable workstations for industrial and scientific use. Using

processors from 486DX2-66 to the P90, the computers have five full-length slots for ISA, EISA and PCI cards, three drive bays and either colour or mono screens. Chassis is plated aluminium for rfi and emi protection, in an ABS outer case. Power comes from either mains or a car battery. Logic Instrument S.A. Tel., 0033 1 39 89 96 22; fax, 0033 1 34 28 00 50.

Data communications

Infrared receiver. Unirode's *UCC5340* BICMOS, micropower ir receiver is the first of the Air Light series, incorporating a limiter to provide a wide dynamic range in which it complies with the Infrared Data Association requirements of 1cm-1m. Bandpass filtering improves s:n margins for 2.4kb/s to 114b/s operation and reduces device recovery time, allowing direct interfacing with IrDA-compatible detector diodes. Output drives a 40pF load at cmos/ttl level. Unirode (UK) Ltd. Tel., 0181 318 1431; fax, 0181 318 2549.

LocalTalk tx/rx. *LTC1324* from Linear Technology is a 5V micropower LocalTalk transceiver, taking 1mA in use and 1µA when shut down. No emi filters are needed,

since the device possesses a 400ns slew-rate mode of operation for data rates of less than 250kb/s. Receiver and driver, both differential, have a choice of active-high or active-low enable pins to three-state the outputs. ESD protection is to ±10kV. Linear Technology (UK) Ltd. Tel., 01276 677676; fax, 01276 64851.

Development and evaluation

PIC development. Microchip offers the *PICMASTER-17B* universal in-circuit system, a development tool for the PIC17C4X microcontroller family. It runs on PCs under Windows and includes an emulator control pod, a target-specific emulator, the *PRO-MATE* programmer, a PC host-interface card, emulation control software, demo hardware and software and documentation. Arizona Microchip Technology Ltd. Tel., 01628 851077; fax, 01628 850259.

Programmers

Universal modules. Stag has a range of removable modules for the *Eclipse* universal programmer to take fpga, pga, quad flat pack and shrink dip devices. The universal pga socket handles all fpga and pga devices up to 208 pins as standard, 256 pin

drivers accessing each pin of the pga socket for true universality. Stag Programmers Ltd. Tel., 01707 332148; fax, 01707 371503.

Software

Statistics for Windows. *Genstat*, a general statistics software package, is now available for PCs running Windows, using the familiar Windows facilities and providing a number of additions to the standard analyses. It provides pre-configured analyses in the normal Windows manner and also via a command line and takes data from Excel and other spreadsheets. Nag Ltd. Tel., 01865 511245; fax, 01865 310139.

OS for embedded PowerPC. The Swedish company Enea Data is about to introduce a version of its *OSE Delta* real-time operating system for Motorola PowerPC MPC821/860 processors, to include dynamic memory management, integrated error handling and automatic process supervision. It is a distributed system meant for use in embedded non-stop and fault-tolerant systems such as telecomms; applications in OSE for 68xxx processors will be portable to the PowerPC. Reflex Technology Ltd. Tel., 01494 465907; fax, 01494 465418.

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TEKTRONIX 465 Dual Trace 100MHz Delay Sweep	£400
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HITACHI V423 Dual Trace 40MHz Delay Sweep	£350
GOLDB OS1100 Dual Trace 30MHz Delay Sweep	£240
HITACHI V223 Dual Trace 20MHz Delay Sweep	£250
BECKMAN 9020 Dual Trace 20MHz Delay Sweep	£240
KIRUSUI 5530A Dual Trace 35MHz	£220
IVATSU SS5702 Dual Trace 20MHz	£200
GOLDB OS300 Dual Trace 20MHz (No handles)	£180
TEKTRONIX 2220 Dual Trace 100MHz Digital Storage Cursors	£1700
TEKTRONIX 468 Dual Trace 100MHz Delay Sweep Dig Storage	£750
PHILIPS PM1305 4 Channel 35MHz Digital Storage	£650
GOLDB OS400 Dual Trace 10MHz Digital Storage	£240
H.P. 1741A Dual Trace 100MHz Analogue Storage	£400

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EATON/ALTECH 380K1 Syn Sig Gen 1-200MHz with PM3602	P.O.A.
MARCONI 2019 Synthesized AM/FM Sig Gen 80kHz-1040MHz	£1750
H.P. 8640A AM/FM Sig Gen 500kHz-1024MHz	£750
H.P. 8620C Sweep Oscillator Multi-Freq only	£200
RACAL 9081 Syn AM/FM Sig Gen S-520MHz	£450
ADRETE 7100A AM/FM Sig Gen 100kHz-1.3GHz	£1000
ADRETE 7100A AM/FM Sig Gen 300kHz-650MHz	£650
MARCONI T2321A Dispersive Freq Meter 0.005% Un-used	£225
MARCONI T2933B Audio Power Meter Sine/d	£250
MARCONI T2163 Attenuator DC-1GHz	£100
R&S Video Noise Meter UP52Z with PSFZE 40-Hz-10MHz	£1500
WILTRON 6429 RF Analyser 10-2000MHz	EPDA
WILTRON 6510A Programmable Sweep Gen 1-2GHz	EPDA
W&S Selective Level Meter SPM11 with Harmonic Distortion Analyser	PSF11
H.P. 4277A LC2 Meter with Test Fixture	EPDA
R&S VHF Field Strength Meter Type HFV 25-300MHz	£250
H.P. 8150A Programmable Pre-Pulse Gen 50MHz	EPDA
PHILIPS PM5136 Sweep Fun Gen 0.001Hz-20MHz Sine/Sq/Tri etc.	£500
PHILIPS PM5130 Syn Fun Gen 0.001Hz-2MHz Sine/Sq/Tri	£750
H.P. 5006 Signature Analyser	£150
H.P. 8504A Signature Analyser	£100
H.P. 8495B Attenuator DC-180Hz 0.70dB in 10dB steps	£350
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SCHLUMBERGER STABLOCK 4021 Test Set	£800

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LETTERS

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

Valve amplifiers made clearer

In the editing of the second extract from my book 'Valve Amplifiers', the precise need for the cathode build-out resistor in the concertina phase splitter was not made clear. When originally conceived in 1982, the purpose of the build-out resistor was to equalise the output resistances, reduce imbalance, and thereby reduce distortion.

However, when tested with square waves, I found that removing the build-out resistor caused a coggle on the leading edge of the waveform, which could not otherwise be eradicated. Each output of the phase splitter forms a low pass filter with any shunt capacitance, and since these capacitances are likely to be similar, the output resistances should also be similar to preserve high-frequency balance; hence the coggle when the build-out resistor was omitted. So although the build-out resistor was originally included to improve 1kHz distortion performance, it has been retained to improve transient response.

Various observers have commented that the output impedances are not equal, and that when the stage drives a real load, its balance will therefore be compromised, although none of them have offered any analysis to show what these impedances might be, or what remedial action should be taken.

Looking into the cathode, we see R_k down to ground, in parallel with r_k the anode path to ground;

$$r_k = \frac{R_a + r_a}{\mu + 1}$$

Substituting;

$$Z_{out} = \frac{R_k \frac{R_a + r_a}{\mu + 1}}{R_k + \frac{R_a + r_a}{\mu + 1}}$$

Simplifying, and noting that

$$R_a = R_k = R_L;$$

$$Z_{out(cathode)} = \frac{R_L(R_L + r_a)}{R_L(\mu + 2) + r_a}$$

Looking into the anode, you will see R_a to the ht rail; which is ac ground, in parallel with the cathode path to ground;

$$r'_a = (\mu + 1)R_L + r_a$$

Substituting;

$$Z_{out} = \frac{R_L((\mu + 1)R_L + r_a)}{(\mu + 1)R_L + R_L + r_a}$$

Simplifying;

$$Z_{out(anode)} = \frac{R_L^2(\mu + 1)R_L r_a}{R_L(\mu + 2) + r_a}$$

If we inspect this equation, we see that the terms involving μ are the only significant terms, and that if is reasonably large, then $(\mu + 1) = (\mu + 2)$, so that the output impedance reduces to R_L .

The cathode output impedance has to be calculated fully, but generally gives a value of about 1k Ω . It is because the ratio of output impedances is so large, that it is essential to equalise them, in order to maintain high-frequency balance. To a first approximation, a build-out resistor in the cathode equal to the load resistance would be required.

Gain to anode or cathode can be calculated;

$$A = \frac{\mu R_L}{R_L + r'_a}$$

Where

$$r'_a = R_L(\mu + 1) + r_a$$

giving;

$$A = \frac{\mu R_L}{\mu R_L + R_L + 2r_a}$$

But very little error results from;

$$A \approx \frac{\mu}{\mu + 1}$$

Because of this low value of gain to the anode, Miller capacitance is very low, and the stage has wide bandwidth.

It is usual to direct couple to the anode of the input stage, and let that determine the dc conditions of the stage, resulting in the saving of a coupling capacitor and a low frequency time constant. By doing so, a pair of triodes achieves about 6dB greater gain than any phase splitter solution based on the differential pair.

It is true that the concertina has a higher output impedance than a differential pair solution, but this was used to advantage in my design

by making this the dominant pole, which meant that further slugging (shunt capacitance) actually improved high-frequency balance, as it reduced the effect of variations in input capacitance in the output valves.

On a historical note, the term 'build-out' resistor derives from BBC practice in equaliser design for analogue music lines. Individual half-section equalisers would often be used to reduce the complexity of the complete equaliser, but their output resistance was no longer 600 Ω , so a 'build-out' resistor was added to avoid disturbing the next section.

Regrettably, there is a typing error (my fault) in Fig. 4, where the output power is specified as 10W into 4 Ω ; it should be 8 Ω .

• I always enjoyed Frank Ogden's editorial when he contributed to the 'Comments' column, but I found it rather sad that he resorted to abuse in his letter in the February 1996 issue – after failing to spot that my design already accounted for the very object of his criticism.

As to who would seriously consider valves for anything, I would ask that he visit a specialist hi-fi shop, or a show, and see how many (British) valve amplifiers are for sale. The demand exists, so the product is made.

Morgan Jones
Southampton

Valve prejudice?

Frank Ogden's flippant assertion that only "a complete loony suffering from terminal nostalgia would seriously consider valves for anything" (Letters, February) betrays him as a devout Philistine. The absurd notion that all past wisdom is rendered obsolescent in the face of technological advance is simply juvenile. The 'Soviet Union' presented a shining example of this folly and look where that got us.

Is Mr. Ogden seriously suggesting that musicians the world over are all sick, demented, or just plain stupid?

The idea that 'new' is inherently better than 'old' is more in tune with adolescent truculence than with any considered intelligence. Please, Frank, don't use science as an excuse for your prejudices – it does neither you, nor us, nor science, any favours.

Simon Yorke
County Durham

Cable communication

Ben Duncan's article about loudspeaker cables in the February's issue was illuminating and highlighted very well the problems of the design and modelling of transmission lines.

The lumped model has its limitations, but Ben Duncan has applied it well. The model he has used describes a quarter wave section of 66 Ω cable operating at 10MHz, but at high audio frequencies it is a good approximation of two metres of 300 Ω cable. This is a good basis for the analysis that followed but is tricky to modify to see the effects of different cable types.

A better approach for this task is to use the distributed model, available in Spice simulators. For simple scenarios such as that considered here, the difference is small until alternative cable impedances are considered. With a 2m, 300 Ω cable modelled the high frequency roll-off is indeed still present. It is a natural consequence of the end-to-end impedance transformation and consequent mismatch which takes place when a transmission line reaches an appreciable fraction of a quarter wavelength. The greater the intrinsic mismatch, the worse the loss.

A better cable. To prevent this, and make the whole system flat in the frequency domain, the solution is to make the cable the same impedance as the loudspeaker – 8 Ω . The impedance of the loudspeaker is then presented unchanged to the amplifier output at all frequencies and there is no roll-off. The response is flat up to the frequency at which the line starts to support non-TEM modes – several GHz. Perhaps counter-intuitively, the value of the shunt capacitance increases considerably to achieve this match. Interestingly the extensive speaker cables which place the two conductors far apart to reduce capacitance, are doing the worst thing possible for frequency response.

Attempts have been made in the past to produce cables of low impedance by weaving enamelled wire together, but they have vanished, presumably because chafing caused short circuits. It is actually quite easy to make reliable low impedance cables for yourself. The impedance does not have to be particularly accurate 8 Ω , merely

somewhere near.

Adhesive copper tape about 18mm wide is available from many stockists and is mainly used for electromagnetic shielding. A good sized reel can be had for about £10.

The copper tape is stuck on each side of a soft plastic strip, about 1mm thick, to form a sandwich. The hole is then insulated with some

tough plastic. The brown PVC tape used for packaging is suitable. The diagram shows a cross section of the cable. Connections to each end can be soldered with short pieces of thick wire. This assembly will lie flat under a carpet. In terms of the corrosion difficulties for stranded cables by the article, this has none of these, and being flat the skin essentially negligible. The cable will contribute only its resistive term to the circuit.

Make the cable disappear. If you really want to remove your loudspeaker cables from the equation, the answer lies within the power amplifier. Most designs use negative feedback, taken from as close to the loudspeaker terminals as possible. This guarantees the most uncompromised voltage source possible at that point. The loudspeaker cables then degrade this to an extent dependant in their length and quality.

Power supply designers have known the answer to this problem for years. Even the simplest designs usually offer remote sensing terminals to guarantee the accuracy of delivery at the load rather than the supply terminals. They generally have a configuration that allows the

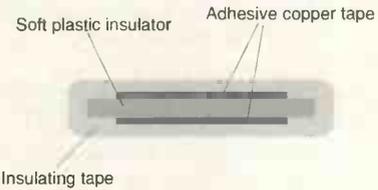
user the choice of high accuracy with a bi-wired remote sensing set-up, or by omitting the second pair of wires, the standard accuracy with the cable impedance degrading performance.

With an audio amplifier – which is just a fast ac power supply – this can be achieved with couple of simple modifications.

The feedback sensing point, which is used to define the location of the net voltage source, is coupled back to the comparator circuit with a resistor, usually several kilo-ohms. This sensing point can be moved by the addition of extra resistors and bi-wiring to the loudspeaker. The precise value of the feedback resistors is not critical as long as it is the same in both channels.

The effect of a change is merely on gain, pro rata to the change so the addition of a few ohms – 10Ω is about right – will not change performance. Provided the bi-wiring is used, the new resistors can be shorted out and have no effect at all. The junction of these two resistors is taken to the far end of the speaker cable with a second wire, effectively moving the sensing point for the negative feedback all the way up to the speaker.

As far as the resulting driving



impedance is concerned there is no cable. The same can be done with the ground side of the cable into the comparator circuit. The topology of this section is more likely to change from design to design and it may be that such a point is not accessible. In that case, you will have to settle for just half a solution.

Feedback wires can be omitted without compromise of the original quality, but have a care. Either disconnection or reversal of the main cable with the external feedback in place will result the smell of burning 10Ω resistors.

Donald Pearce
Hampstead

Internet integrity

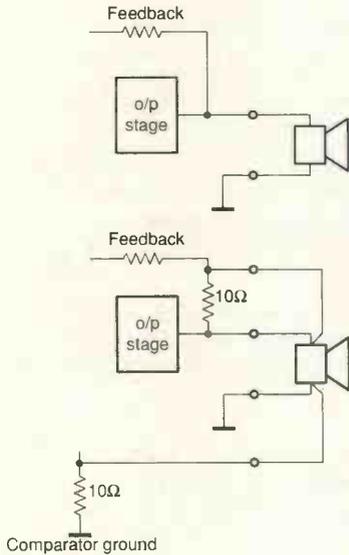
The number of Internet users continues to grow at a significant pace. Users have the facility to call down virtually any page from this massive database and also have the opportunity to create their own Web pages. How much thought has been given to the integrity of the data which can be accessed from the net? Can we really believe that all we read has not been corrupted by a determined hacker first? With all the high grade of technical data available, is it right that we should accept its authenticity without question? The very openness of the net should be questioned.

Interesting enough, I have recently read that a new encryption technique using a one pad key is awaiting the grant of its patent. It was promised that this would then be passed into the public domain. If business users are intent on passing sensitive information over the net, then a secure form of encryption is an absolute necessity.

Geoff Lewis
Canterbury
Kent

More current drive

I noted the letter by J.R. Allison (Bradford) in the December edition. He comments on the use of current feedback in audio amplifiers, or 'current-drive' as I prefer to call it in a system context. I should point out that this subject has been investigated in the 1980s at Essex University by Dr Paul Mills, a then research student under my supervision but now with Tannoy.



Creative fiction

Circuit simulators are extremely powerful tools-for investigating electronic scenarios. I have always thought it too obvious to need saying that relying wholly on simulation can lead you off into deep error unless you do regular reality checks with real circuitry to ensure your model is not fallacious. I was wrong; it does need saying.

While I must decline to give Duncan's article 'Modelling Cable' in the February 1996 issue the full dissection it deserves, I feel it only right to reassure people who are appalled at the prospect of loudspeaker cables made up from a Sargasso Sea of 10mV threshold diodes. I believe the simple experiment shown below clears up the point.

A Blameless power amplifier delivers 50W into an 8Ω load, at 0.001% thd at 1kHz. The load return path is the cable under test A-B. This was initially 4m of multi-strand copper – not tinned-copper – mains-cable, and the voltage drop across it was 140mV rms at 1kHz. I think it is clear that if the multi-strand cable was a cats-cradle of 10mV 'mystery diodes', then under these conditions some at least of them would be turning on and off and would severely degrade the

linearity. This would be particularly noticeable at the strand ends C, D etc, which are not formally connected to B, and according to Duncan make "occasional and random contact" along the length of cable. In actual fact they make intimate and continuous contact in any normal cable.

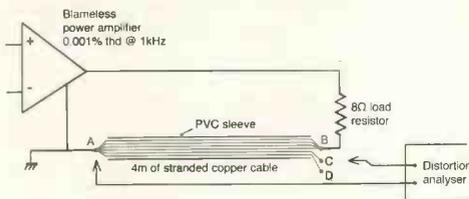
None of this happens. It is not possible to measure thd to 0.001% on a 140mV signal due to noise limitations, but the thd residual at points at C, D etc, is identical to that at B, at the measurement noise floor of 0.007%. Clearly this is just a noisier version of the Blameless amplifier distortion. No extra diode distortion of any kind is detectable; this holds across the 20Hz-20kHz band.

Above 2kHz the voltage drop increases due to cable inductance; by 10kHz the level is 10dB up, rising at 6dB/octave. This confirms my view that in most circumstances the series resistance and inductance of a loudspeaker cable are the only significant parameters. If you are trying the experiment, then don't leave the cable wound on its reel; not only will this increase its inductance, but the steel reel-centre will produce spurious third-harmonic distortion.

This is good evidence that Mystery Diodes do not exist. There is also the point that a copper-oxide rectifier has a forward voltage drop of 300-600mV, and I can find no evidence in solid-state physics that cupreous oxide could ever give a diode effect at such a low voltage as 10mV.

It remains true that an ounce of experiment can dispose of several long tons of speculation.

Douglas Self
London



Aspects of this subject were also investigated by David Birt, then with the BBC Designs Department. Rather than discuss the merits – and problems – of current drive, I would like to draw your readers' attention to the following publications:

'Distortion reduction in moving-coil loudspeaker systems using current-drive technology', Mills, P.G.L., Hawksford, M.O.J., *JAES*, vol. 37, no. 3, pp129-148, Mar '89.

'Transconductance power amplifier systems for current-driven loudspeakers', Mills, P.G.L., Hawksford, M.O.J., *JAES*, vol. 37, no. 10, pp809-822, Oct '89.

'Transient analysis: A design tool in loudspeaker systems engineering', 80th Convention of the AES, Montreux, paper 2338 (E8), Mar '86.

'Loudspeaker power amplifiers with load-adaptive source', Birt, D., *JAES*, vol.36, no.7,8, pp552-61, July/Aug '88.

I should point out that a two-way

active system was designed and built as part of the research programme. The high-frequency channel used pure current drive while the bass-midrange used a combination of current drive and servo technology using velocity sensing to control damping. Paul Mills has also more recently designed a sub-woofer loudspeaker for Tannoy using current-drive techniques.

At Essex since the mid 1980s, we have also investigated digital techniques for crossover design where precision crossover and drive unit correction functions can be implemented in the digital domain. This can result in a loudspeaker with a near-perfect impulse response, where both amplitude and phase responses are corrected.

Of course, the actual loudspeaker must be appropriately designed to benefit from this technology – see Audio Physic 'Tempo' or 'Virgo' for examples of good design practice.

Also – and of equal significance – much of the analogue processing can be eliminated with high-grade d-to-a converters being interfaced directly to the power amplifiers. The

elimination of much of the analogue circuitry including passive crossover networks and pre-amplifiers, together with band splitting across multiple d-to-a converters and power amplifiers, is in my view a revelation. It represents one of the most important contributions of digital technology to sound reproduction.

Redundant analogue amplifier stages only downgrade audio performance. On the other hand, products such as the Meridian 565 offer digital gain control, digital tone controls and surround-sound modes. They also operate efficiently in both analogue and digital worlds. As a bonus, the dual phase-locked loop of the 565 lowers jitter from digital sources.

In the near future we can also look forward to the power d-to-a converter which could eliminate the need for analogue power amplifiers as we currently know them.

In my view, the future of audio technology is about to change dramatically, and with the coming of high-capacity cd with multi-channel sound*.

Finally, I hope that the

technological changes that digital processing offers will put to rest the rather protracted and non-convergent debates on analogue circuitry especially power amplifiers, that continue to appear in these pages, much of which is soon to be of only historic interest, at least at the leading edge of audio. The fact is all analogue circuits are flawed for numerous reasons and each must be judged as a complete entity, circuit diagrams are only part of the story. A technology that eradicates these areas of uncertainty is far more likely to succeed when properly engineered.

Although I am fascinated by analogue circuit design and I have more than my fair share of controversial views, the more I learn of it the less I want to use it. Of course, treating all circuits including digital as an analogue process has its merits, but that is a different story.

Professor M Hawksford
Centre for Audio Research and Engineering

*Acoustic Renaissance for Audio (ARA) proposal web pages:
<http://www.ibmpcug.co.uk/~meridian/ara/>

Electromagnetic clarification?

Requirements of standards vary, but in particular, EN55011 specifies a voltage maximum, a distance and an antenna. This appears to be fine when the current measurement from the specified loop antenna, with its factors, is converted to a voltage. The resultant voltage can then be referred to the standard for a pass or fail.

Fine so far. It is when one makes a measurement at 30m distant from the source, with a loop antenna, at a frequency of less than 2MHz that I see a problem.

This measurement will be in the near field, with a current measuring device. How, then, is to be converted to a voltage when the voltage and current relationship is not established, as it is in the far field?

I consider the far field to be around $1/5$ -wave length distant of the source and the relationship in the far field to be, in simple terms, 1V/metre to 0.00265A/m or a correction factor of log 20 377, or 51.5dB.

M J Nicholas
*Bournemouth
Dorset*

Electro-magnetic confusion

The recent correspondence on electromagnetic compatibility in your February issue, barely touched on the growing electromagnetic pollution of the environment.

There seems to be a tendency to ignore the effects of broadband 'hash' on unfashionable parts of the electromagnetic spectrum. The fact that most pcs blot out radio reception on nearby long and medium-wave (150kHz - 1.5MHz) radio receivers is dismissed as unfortunate. But then the preferred band for 'serious' radio listening is fm.

The truth is that we are drowning in a sea of electromagnetic pulses from electronic light ballasts, domestic appliance controls and television and pc displays and systems – and nobody cares.

This off-hand attitude to radio interference now extends far up the broadcast spectrum. We are shortly to be exposed to channel 5, which will occupy uhf channels 34-39. According to my frequency allocation list, the band 606-614MHz in that sector was allocated for radio astronomy. The jamming of research radio reception does not end there. *Electronics World* readers may recall that the search for intelligent life elsewhere in the galaxy was recently stalled by a microwave oven in the observatory kitchen. Its a moot point whether intelligent life had anything to do with the signals.

I fear it will take some spectacular computer system crashes due to rogue electromagnetic pulses to convince the slew-rate junkies, terabyte nerds, surfers and vhf anoraks currently in the ascendant that limiting electromagnetic pollution by improving the design, construction and screening of radiating equipment is in everyone's interest.

Anthony Hopwood
*Upton-upon-Severn
Worcestershire*

Early EMC

Electromagnetic compatibility is widely regarded as something invented in Brussels a few years ago, that became a legal requirement at the beginning of this year. In fact, controls on emission have been around for some forty years, one of the first being the requirement for 'suppression' of the ignition systems of petrol engines to minimise interference with Band I 405-line television transmissions.

Control of immunity in volume-produced products, however, is often considered to have been introduced only a decade or so ago, as a result of interference from amplitude modulated citizen's band transmissions.

It is interesting therefore, to note that in the Dec 1971 issue, *WW* reproduced an item from *The Marconiograph* of sixty years previous concerning a patent issued to Capt. H. J. Round. This is not only the earliest reference to a 'valve' in the *The Marconiograph*, but also perhaps the earliest example of improvement of immunity.

For readers who do not have either of those issues to hand, the invention consists of a reversed-polarity diode connected in parallel with the diode detector of a simple detector-only receiver. Both diodes are provided with variable reverse bias and the reversed-polarity diode is biased off for signal reception. The amount of bias, and its polarity, are not well defined. However, a 'powerful atmospheric discharge' forces both diodes to conduct. This results in a great reduction in detection efficiency, due to the low 'back resistance', as well as heavier damping of the parallel tuned circuit. As a result, no sound, or probably only a weak click, is heard in the earphone.

As might be expected from the ingenious Captain, this two-pronged attack is clearly likely to be more effective than just connecting the 'crash diode' across the tuned circuit.

John Woodgate
*Rayleigh
Essex*

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Non-slewing audio power amplifier

Advancing his 'non-slewing' technique, which enabled an instrumentation amplifier capable of $\pm 1000V/\mu s$, Giovanni Stochino now applies the principles to audio power amplification.

In a recent article¹ some high slew rate voltage feedback amplifier architectures were presented and discussed. A virtually non-slewing 50Ω power amplifier configuration was finally proposed, where high speed is accompanied with low input offset voltage and noise, as well as with low total harmonic distortion in the audio frequency range.

In this article I show how the same basic principles, embodied in Fig. 9¹, with appropriate adaptations, can be applied to the design of high performance audio power amplifiers.

Non-slewing amplifier performance

In non-slewing architectures, as explained in the previous article¹, all main slewing mechanisms are virtually eliminated. Therefore, when the input signal is within the common mode input voltage range, the current at the output of both input and intermediate gain stages, is always controlled by the differential input voltage $V_d = V_1 - V_2$.

Figure 1 allows us to compare the performance of a non-slewing amplifier, nsa, with a similar yet conventional and slewing architecture, or csa. For clarity, only the input and intermediate stages are detailed, while the output stage is shown in the form of an ideal voltage follower.

Biasing currents are the same in both nsa and csa configurations. The results of Spice simulation for an input stimulus consisting of a 10kHz square wave with a superposed high frequency sinusoidal voltage of 500kHz are shown in Fig. 2.

It is apparent that the non-slewing architecture has no visible transient intermodulation distortion, while the conventional configuration shows clear signs of the presence of inter-

modulation distortion due to slewing times T_f and T_r – both of about 7μs.

In Fig. 3, transfer curve I_b , which is the intermediate stage output current, versus V_d is shown for both $I_B=0$ and $I_B \neq 0$.

Three regions of operation are identified:

- I - input and intermediate stage in class-A operation;
- II - input stage in class-AB, intermediate stage in class-A operation;
- III - input and intermediate stage in class-AB operation.

Transconductance $g_m = I_b / (V_1 - V_2)$ in these regions is approximately as follows:

$$g_m(I) = 2 / (R_c + 2V_T / I_A) \text{ (small signal)} \quad (1)$$

where $V_T = kT/q$ is the thermal voltage of around 25mV at ambient temperature, and,

$$g_m(II) = 2 / (2R_c + R) \quad (2)$$

$$g_m(III) = 1 / (2R_c + R) \quad (3)$$

Maximum positive and negative current $I_{b(max)}$ at node B is defined by the maximum input voltage $V_{max} = V_{EBO} + V_{be(on)}$ that should never be exceeded, according to the following relationship:

$$I_{b(max)} = (V_{max} - 2V_{be(on)}) / (2R_c + R) = (V_{EBO} - V_{be(on)}) / (2R_c + R) \quad (4)$$

Here, voltage V_{EBO} is the rated base-emitter reverse voltage of input transistors. If, for instance, $V_{EBO} = 6V$, $R_c = 50\Omega$ and $R = 200\Omega$, equation (4) yields 18mA. This value, added to I_B , is enough to sustain a rate of change – still linear – of $\pm 160V/\mu s$ across a capacitance

Performance of the 'non-slewing' amplifier.

Test conditions: ambient temperature = 20°C; $V_{cc} = V_{ee} = 55V$ regulated.

- Gain = 30dB
- Output power 20Hz-20kHz, 110W/8Ω; 180W/4Ω
- Small signal -3dB bandwidth, 800kHz (at node F before output inductor)
- Input offset voltage, 4mV
- Maximum output voltage rate of change, $\pm 170V/\mu s$ (at node F)
- Overload recovery time (up to 300% input overload) $\leq 120ns$
- Distortion, see Table 1.

Table 1. THD+noise of circuit in Fig. 4, bandwidth 80kHz.

$V_{out} (V_{pp})$	thd+noise(%) 8Ω load		thd+noise(%) 4Ω load	
	1kHz	20kHz	1kHz	20kHz
	10	0.004	0.016	0.004
20	0.003	0.020	0.004	0.040
40	0.003	0.030	0.008	0.040
60	0.003	0.040	0.009	0.050
80	0.004	0.045	0.010	0.060

thd+noise instrumentation = 0.002%

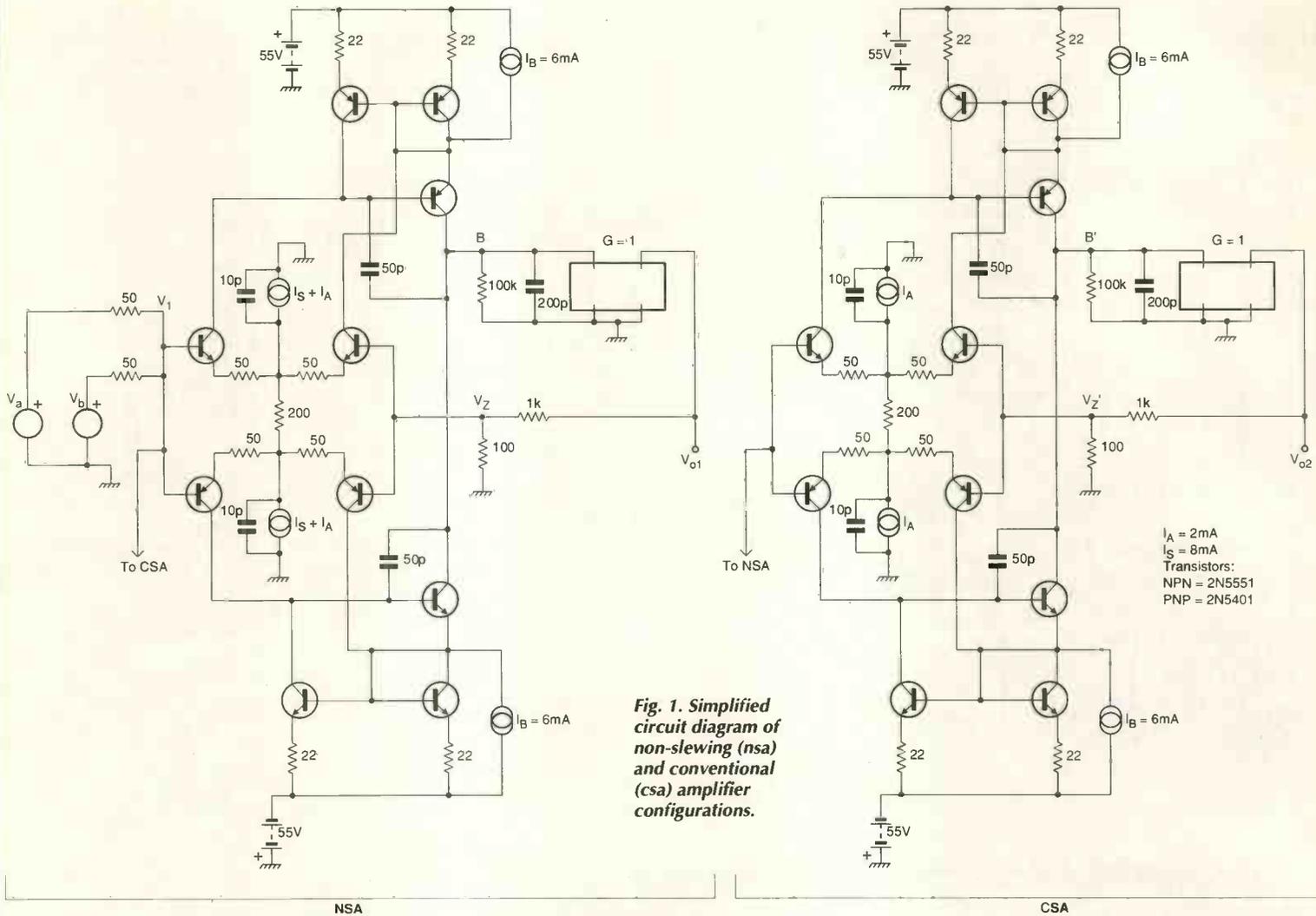


Fig. 1. Simplified circuit diagram of non-slewing (nsa) and conventional (csa) amplifier configurations.

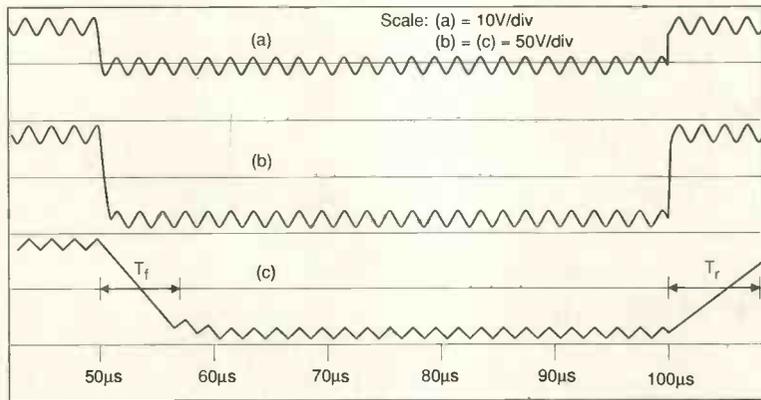


Fig. 2. Spice simulation results demonstrate the absence of visible TIM in non-slewing architectures. (a) is $(V_a + V_b)/2$, (b) is output of non-slewing amplifier, V_{o1} , and (c) is output of conventional slewing amplifier, V_{o1} .

of 150pF at node B, which is a realistic value in an audio power amplifier.

It is worth noting again that this maximum rate of change is not to be confused with the non-linear slew rate limitation due to input and/or intermediate stage overdrive phenomenon in csa architectures.

The remaining problem in this type of nsa configurations, as well as in all other high slew rate class-AB architectures, is associated with the large full scale non-linearity due to class-AB operation, as is apparent from Fig. 3. This

can leave a residue – although rather low – of transient intermodulation products, even when overall feedback is applied to the amplifier. Additionally this limits large signal linearity at high frequency where loop gain is reduced for reasons of loop stability.

An effective way to further limit the already low transient intermodulation distortion in non-slewing audio power amplifier architectures, is to set the bias current of input and intermediate stages so that class-A operation is retained with actual music programs at the

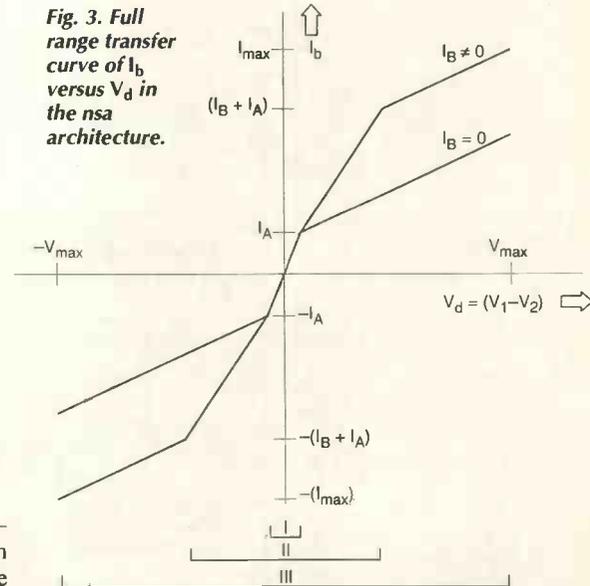


Fig. 3. Full range transfer curve of I_b versus V_d in the nsa architecture.

maximum expected power and frequency. Considering that the peak current needed to sustain full swing at node B is equal to $2\pi f V_{cc} C_0$, where C_0 includes also the base-collector capacitances of transistors connected to node B, the previous reasoning leads us to set $I_A \geq 1\text{mA}$ for $f = 20\text{kHz}$, when $V_{cc} = 50\text{V}$ and $C_0 = 150\text{pF}$. As suggested by Self², a safety

AUDIO

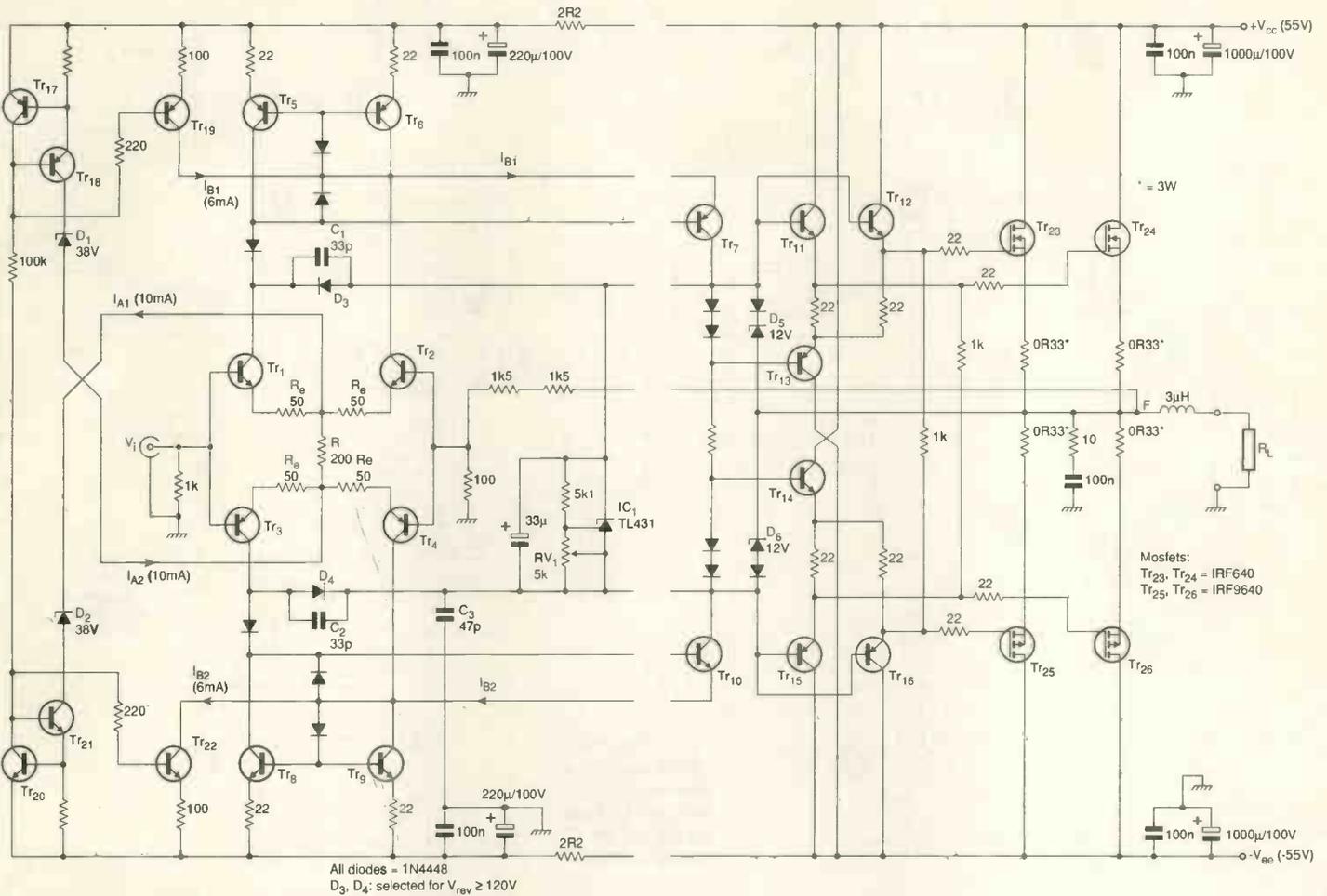


Fig. 4. Detailed circuit diagram of a viable non-slewing audio power amplifier.

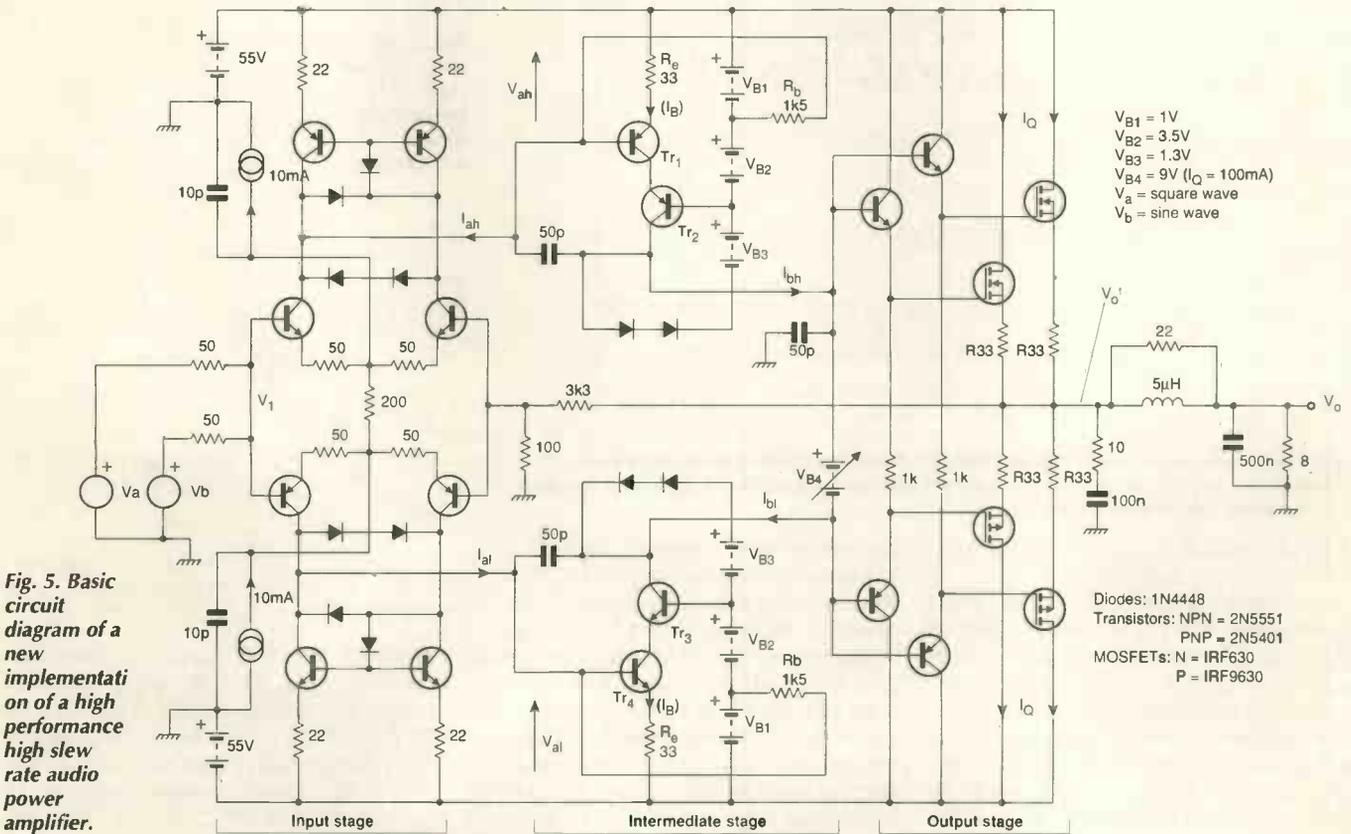


Fig. 5. Basic circuit diagram of a new implementation of a high performance high slew rate audio power amplifier.

margin of about two is recommended. This also takes into account the current contribution needed to drive the output stage and to make up for component tolerances.

In principle, this solution promises very good thd and transient intermodulation distortion performance with normal audio programs. At the same time, it is capable of assuring the very low – if any – transient intermodulation distortion offered by non-slewing architectures, should unexpectedly fast and/or large input transients occur, as maintained by Duncan³.

Non-slewing, high power audio amplifier architecture

Figure 4 shows the complete circuit diagram of a possible implementation of a non-slewing audio power amplifier, designed bearing in mind the above considerations. If compared with the 50Ω power amplifier of Fig. 9¹ discussed in my previous article, it contains some minor changes.

Firstly, cross coupling resistance R is reduced to 200Ω to improve full range linearity of the input stage and increase its maximum available output current. Current shifting component I_S has been increased accordingly to 8mA, while I_A has been kept at 2mA.

To reduce power consumption of current source transistors Tr_{18} and Tr_{21} , zener diodes D_1 and D_2 have been added. Furthermore, the output stage is built around a double pair of high-power complementary mosfets. These are the IRF 640 and IRF 9640, from International Rectifier.

Each power mosfet can dissipate 125W of power and provide more than 20A peak current. Therefore, the amplifier can safely drive very low impedance loads, provided the output mosfets are adequately heat-sinked.

Transistors Tr_{11-16} , which serve as low output resistance push-pull drivers, are capable of providing the high peak currents needed to drive the high and non linear input capacitances of power mosfets. This can amount to about 800pF worst case each. Peak output current limitation is set to about 40A by zener diodes D_5 and D_6 , but a 3-4A fuse has to be inserted at the amplifier output for safe continuous operation.

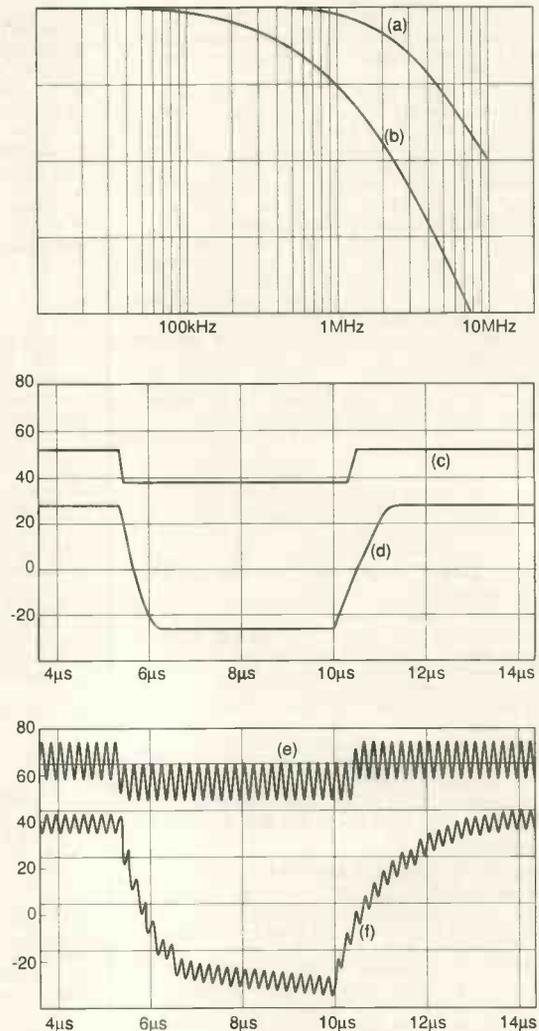
Due to the high gate-source voltage needed to drive mosfets into full conduction, maximum output voltage swing is limited to about 12V from the supply rails. This results in reduced amplifier efficiency. If needed, better efficiency can be obtained by operating power mosfets with separate supply rails of ±45V.

Output stage bias current is set by adjustable shunt regulator, IC_1 , and by trimmer RV_1 . This component has to be set to its maximum before applying power to the amplifier and adjusting bias current of Tr_{23-26} . A suitable value for total mosfet bias current was found to be about 200mA. Unless otherwise stated, resistors are 300mW, 1% metal film types.

Measured performance of the amplifier prototype, which is in good agreement with simulation data, is shown on the first page of this article.

Fig. 6. Some Spice simulation results of the amplifier configuration of Fig. 5:

- (a) Frequency response, Magnitude of V_o'/N_1 (10 dB/div)
- (b) Frequency response, Phase of V_o'/N_1 (50°/div)
- (c) Square wave response, V_1 (2V/div)
- (d) Square wave response, V_o' (20V/div)
- (e) TIM test, V_1 (5V/div)
- (f) TIM test, V_o' (20V/div).



Further developments.

In the last few months I have investigated whether better results can be attained in terms of both maximum rate of change and thd in high slew rate audio power amplifiers configurations.

Figure 5 illustrates the further evolution of the basic power amplifier circuit structure depicted in Fig. 4, which I am currently working on. In comparison with Fig. 4, where the intermediate stage current gain is low (1 to 2), Fig. 5's topology features an intermediate stage with high low frequency current gain, equalling R_b/R_e for $R_b/R_e < \beta$.

With component values shown in the figure, this gain amounts to 45. The expected advantages are increased low frequency open loop gain – and hence lower thd for the same closed loop gain – and potentially higher speed, since the current available at the intermediate stage output is larger than the corresponding current in Fig. 4's architecture.

From Fig. 5 you will find that,

$$I_{ah}(V_d=0) = I_{al}(V_d=0) = 0,$$

therefore bias current I_B is equal to,

$$[V_{B1} - V_{be(on)}] / R_e = 9\text{mA}.$$

Available peak current is defined by V_{ah} and

V_{al} peak value, ie $V_{a(pk)}$, via the relationship $I_{b(pk)} = [V_{a(pk)} - V_{be(on)}] / R_e$.

In the interests of reliability, this peak current has been limited to about 60mA, by means of the diode clamping networks at the collector of input transistors. These limit $V_{a(pk)}$ to about 2.8V. With this high peak current value it is now possible to sustain slew-rates of ±400V/μs across a 150pF total capacitance at the input node of the output stage.

I cannot provide measurement results yet, however, Spice simulations confirm the above theoretical predictions.

For the moment, I can report the simulation results shown in Fig. 6, which demonstrates the frequency response – magnitude and phase – the square-wave response and the square+sine wave response. They prove the stability and clean response of the amplifier – even with a severe 8Ω/0.5μF load impedance. In addition, they show the absence of any visible transient intermodulation distortion. ■

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1. Stochino, G, 'Ultra-fast Amplifier', *EW+WW*, Oct 1995, pp.835-841.
2. Self, D, 'High Speed Audio Power', *EW+WW*, Sep 1994, pp.760-764.
3. Duncan, B, 'Simulated Attack on Slew Rates', *EW+WW*, Apr 1995, pp.303-309.

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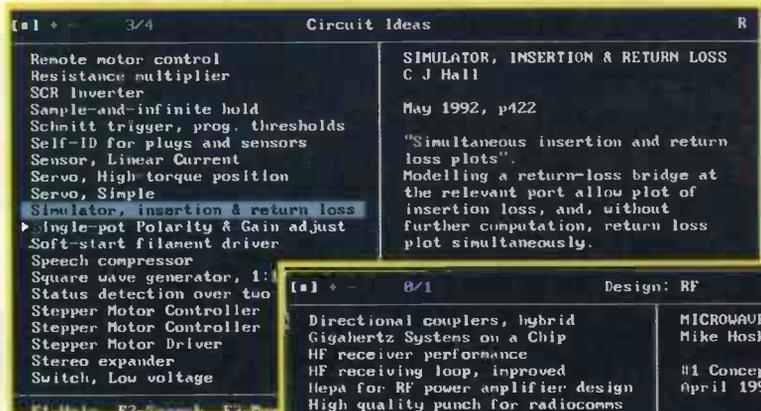


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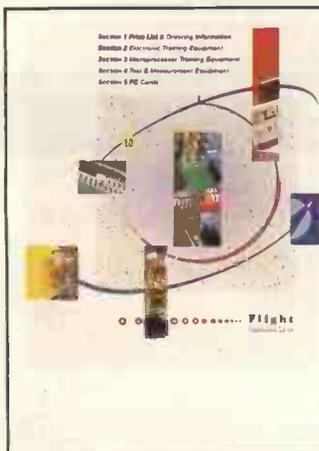
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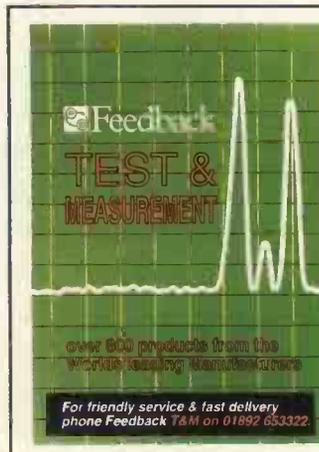
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SPECTRUM ANALYSERS



- HP141T 1.25GHz system (8552B, 8554B) £1250
- HP8557A 350MHz (fitted in 182C mainframe) £1500
- HP3580A 5Hz-50kHz audio frequency spectrum analyser £750 to £1250
- HP3582A audio frequency fft analyser dual-channel £2000
- HP8568A high-specification 1.5GHz spectrum analyser £7500
- MARCONI 2386 100Hz-26.5GHz (in 1Hz steps!) £15000
- TEKTRONIX 492 21GHz portable spectrum analyser, with options 1 2 & 3, and complete with waveguide mixer set (and diplexer) to cover 18-40GHz £7000

MARCONI INSTRUMENTS



- 2019A synthesized AM/FM signal gen 80kHz-1040MHz £2000
- 2305 modulation analyser 50kHz-2.3GHz £2500
- 2828A/2829 digital simulator/analyser £500
- 2382 spectrum analyser hi-spec 100Hz-400MHz £8500
- 2355B radio communications test sets - LATEST 'B' MODELS £4000
- 2326 TV generator & inserter (NTSC variant) £500
- 6460/6421 power meter & sensor 10MHz-12.4GHz £350
- 6500 amplitude analyser c/w 2c 6514 waveguide detectors
- 6960 microwave power meter with 6910 power sensor 10MHz-20GHz
- QA2805A pcm regenerator test set
- TF2910/4 non-linear distortion (video) test set
- TF2910 TV interval timer

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DISTRIBUZIONE E ASSISTENZA, ITALY: TLC RADIO, ROMA (06) 871 90254

HEWLETT PACKARD



- DATRON 1065 digital multimeter all ranges plus IEEE £500
- FARNELL 2081/100 100W RF power meter £200
- DC-500MHz (1GHz)
- IDACOM PT502 protocol analyser (Hewlett Packard E3910B with E3936A) £2000
- JJ INSTRUMENTS CR600 2-channel pen recorder £250
- KIKUSUI 8520 frequency response analyser with sweep generator 4600 £500
- PHILIPS PM5167 1mHz-10MHz function generator £275
- PHILIPS PM8272 X-Y & Y-t dual-channel pen recorder £850
- RACAL-DANA 1992 frequency counter 1.3 GHz IEEE option £700
- RACAL 9008 automatic modulation meter £350
- RACAL-DANA 9300 milli-voltmeter £400
- RACAL-DANA 9301A true RMS RF milli-voltmeter £350
- SYSTRON DONNER 1300 synthesized signal generator 100Hz-1GHz GPIB £1350
- TEKTRONIX AA501/SG505 distortion analyser (complete with TM503) £1250
- TEKTRONIX P6303 o'scope probes NEW 250MHz X1/X10 with readout pin £50 ea
- TEKTRONIX 1503 tdr metallic-cable tester with opts 03/04 £1000
- TEKTRONIX 2465 option CTS oscilloscope £2000
- TEKTRONIX P6201 FET PROBE £350
- WANDEL & GOLTERMANN WM30 level tracer £500
- WAVETEK 23 synthesized function generator 0.01Hz-12MHz £1250
- WAVETEK 1067 opt 522 1-500MHz sweep generator £500
- WAYNE KERR 3220 20A bias unit (for 3245 inductance analyser) £1250

TEST EQUIPMENT

- ANRITSU MF76A 18GHz microwave frequency counter £1500
- BRUEL & KJAER 1023 sine generator £1250
- BRUEL & KJAER 1027 sine random generator £1750
- BRUEL & KJAER 2619 preamplifier complete with 1/2" mic £500
- BRUEL & KJAER 2619 preamplifier complete with 1/2" mic £500
- BRUEL & KJAER 2511 vibration meter (field set with 1621 filter) £1000
- BRUEL & KJAER 2307 level recorder £1000
- BRUEL & KJAER 2317 portable level recorder £1500
- BRUEL & KJAER 2607 measuring amplifier £600

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- BRUEL & KJAER 2609 measuring amplifier £750
- BRUEL & KJAER 2308 analogue X-Y pen recorder £750
- BRUEL & KJAER 2639 preamplifier £325
- CHASE LFR1000 interference measuring receiver 9kHz-150kHz £900
- £750
- DATRON 1061 & 1061A - various, digital multimeter & 1065 - call £1000
- £250 from £500

SPECIAL OFFER THIS MONTH ONLY ON HEWLETT PACKARD DIGITAL TRANSMISSION TESTERS, PROTOCOL ANALYSERS

Examples: HP37724A SDH/PDH portable test sets-
 Condition as new, price £5000 (were listing at ca £20K!) – also
 37772A optical interfaces available for STM-1, £1500 (list over 6K).

SPECIAL OFFER THIS MONTH ONLY
**** HEWLETT PACKARD 8673M ****
synthesized signal generators 2-18GHz
 (Similar model to 8673B but coverage as above)
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(Will be listing at £7,500)

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- 64700A, MAINFRAMES WITH:
- 64747B, 64771G, 64703A, 64704A, 64751A, 64172B, 64748A, 64748C,
- 64787B, 64604A, 64783A, 64172B, 64746J, 64744A/C/E, 64780A.

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