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Cover - Hashim Akin

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3D optical storage could offer significant advantages over conventional 2D techniques - page 450.

3.5nm
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Depletion regions
0.4nm
Gate

The remarkable characteristic of this 2D Mesfet is that it functions with only 200 electrons in its channel, allowing ultra-low-power operation - page 444.

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June 1996 ELECTRONICS WORLD 441
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Computer Shopper Nov 1995

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ELECTRONICS WORLD June 1996
Thanks - but no thanks

Over many years, I have watched with horror as electronics companies launched products which seemed to have been designed without any thought for the paying customer. Even as they hit the market, we just knew they were going to fail.

This represents a tricky dilemma journalists covering electronics. Should we write off the product as a flop, and thereby help to make it one? Or should we try to give it a helping hand, thereby putting consumers at risk of buying something that will later become obsolete?

In the case of Sir Clive Sinclair’s C5 trike, it was an easy call. No matter that the battery went flat as quickly as anyone who has flattened a car battery with the starter could have predicted. Had Sir Clive ever actually tried taking one out on a busy road, and seen how it felt to be next to the wheels of a lorry?

Last year ICL and Fujitsu joined forces to launch the Indiana range of PCTVs. Soon after Olivetti followed suit with Envision. These three companies had one thing in common; no experience in consumer electronics. Even so it is hard to imagine how their designers could have made such a basic mistake.

Although the idea of using one tv screen for both a pc and tv is superficially seductive, there is an obvious and fatal flaw. Television screens are intended to display relatively low resolution tv transmissions. They look fine from across the room. Move close to a tv screen and you see interface jitter; horizontal lines flicker vertically as the fields change. On 50Hz PAL sets, there is the added problem of wide area flicker on bright white images. The colour shadow-mask pitch is coarse, to match the tv scanning raster. So pc text and graphics look coarse too.

A tv screen makes a rotten pc display. That is why, after the era of Sinclair Spectrums and Amstrads, the pc industry moved away from tv standards, to higher refresh rates, and to start with pcs with dedicated monitors.

Did anyone inside ICL, Fujitsu and Olivetti ever sit down in front of their own PCTVs and actually try some serious word processing, database building or spread sheet design? Apparently not.

Like blood from a stone I have now squeezed the admission that production of the Indiana and Envision stopped last winter, and neither ICL/Fujitsu or Olivetti has any firm plans for the future.

Philips developed Laservision, the 12in optical video disc, in the seventies. Rival electronics companies all round the world very quickly saw the elegance of laser read-out and gave up their research on alternative systems. Never mind that Laservision was launched later than promised, and was thus faced with competition from the vcr. The technology spawned CD, CD-ROM and recordable disc.

Stubbornly, RCA pressed ahead with CED Selectavision, a 12in grooved capacitance disc that needed a caddy. It was launched in the US and UK, but plain as pigskeewer never had a chance. Hardly a disc played without skipping and the player had to incorporate a ‘nudge control’ to knock the stylus when it stuck in the groove. RCA’s blinkered vision cost the company so much in hard cash that its name and trade infrastructure were bought by Thomson of France.

Commodore once had a healthy share of the computer and hobby market. Seeing Philips preparing for the launch of CD-interactive, Commodore dreamed up the idea of CDTV – an Amiga-based interactive CD that was incompatible with the Philips format. Commodore hoped that by rushing to the market ahead of Philips, CDTV would become the de-facto standard.

One selling point was that the CDTV player was also a cd player. So how could Commodore have made the basic mistake of using a computer CD-ROM drive that played discs only when loaded into a caddy? Commodore’s answer was that CD data discs need protection from finger marks.

With caddies costing around £5 a time, it was impractical to give one away with each CDTV disc. Audio cds come in jewel boxes, anyway. So Commodore gave away one free caddy with each player and the owner had to load each disc in the caddy before playing it.

Instead of taking a CDTV out of its jewel box and loading it into a cd player tray, the owner had to take the CDTV out of the jewel box, and load it in the caddy before loading the caddy in the player. There was no less handling of the disc, just extra inconvenience.

The caddy issue raises its ugly head again this year, but from the opposite direction. Toshiba, Thomson and Time-Warner are promising to launch high density DVD movie players, with Philips, Sony and others following next year, in less of a rush. By early May there was still no final agreement on vital standards issues, such as copy protection to prevent people dubbing high quality movies from disc to tape, and regional control, to stop movies released in the USA playing on European players.

But there is one issue on which the DVD Alliance is solid. Having learned from Commodore’s mistake on CDTV the hardware and software companies have pledged that the DVD system will not need a caddy. Discs will be packaged like ordinary CDs and loaded direct into a tray in the player.

How strange then that some prototype DVD players still use caddies. Although the more modern prototypes use caddy-less tray loading, it pays to watch any DVD display and see how the demonstrator handles the discs – with extreme care, at the extreme edges. No-one grabs a DVD disc with sticky fingers. But unfortunately this is exactly how the public treats CDs.

They wash their DVDs that way too. Will the error correction cope? We are assured it will, even though the capacity of the new disc is so much greater, thanks to pits that are so much smaller.

If the DVD Alliance launch their system without first giving players and discs to sticky fingered next door neighbours, they could pay the same price paid by Sir Clive, RCA, ICL, Fujitsu, Olivetti and Commodore. That price is horrendously expensive failure thanks to the public’s reluctance to pay money for neatly packaged Inconvenience.

Barry Fox
EMC all over again?

Last year's panic over the introduction of CE marking for the EMC Directive is set to be repeated when the net technical harmonisation standard rears its head. If industry thought EMC compliance caused problems, then enter the Low Voltage Directive, or LVD. According to industry experts, many electronics engineers and manufacturing companies have little understanding of the LVD.

Andrew Perkins, instrument sales manager for Schaffner EMC, involved in conducting seminars said: "People's knowledge of the LVD is akin to that of the EMC Directive this time last year. Less than 5% of people attending the seminars are doing the tests correctly." The lack of understanding is also resulting in few companies conforming to the Directive.

As of January 1, 1997, the LVD attains the same status as the EMC Directive, with manufacturers being required to CE mark for LVD conformity.

This means that someone will have to sign a declaration of conformity and will be held accountable in law. Safety divisional manager at TRL, EMC is Simon Barrowcliffe. "Small and medium sized companies are exhibiting good engineering practice in most cases", he said. "They are designing for safety but they are not conforming to the LVD."

What is the Low Voltage Directive?

- The LVD is the standard specifying design and testing of equipment connected to the mains. The 'Low' refers to voltages below 1 kVac and 1.5 kVdc. It differs from the EMC Directive in several significant respects:
  - The LVD dates from 1973 and is one of the first technical harmonisation standards. This has given companies ample time to prepare for LVD-day. Unlike the EMC Directive, there is no excuse for ignorance and little or no leniency will be allowed.
  - Harmonised European standards for the LVD are much more complex. Longer design and testing time will be necessary to ensure total compliance by 1997. One route to compliance is via approval schemes such as the BABT scheme for telecoms and the BEAB for consumer and white goods. These specify third party testing and formal quality systems such as ISO9001. Smaller companies with less funds and manpower may find these schemes prohibitive in cost and time.

EMC amnesty for UK manufacturers

British electronics manufacturers have been given an 'unofficial' EMC breathing space by the government.

The DTI has told local trading standards officers, who have been officially policing the EMC Directive since January, to take a lenient approach to enforcement at least for this year.

"There has been guidance to local authorities to allow a period of grace," said David Roderick, director of the safety committee for the Institute of Trading Standards. In a move, which seems to be out of step with the rest of Europe, the DTI appears to be responding "sympathetically" to those suppliers waiting to have their products qualified in test houses. "There is concern about the test houses not being able to cope, so where people have made attempts to get things done, we will look upon it sympathetically," said Roderick. No one in the UK has yet been prosecuted under the EMC Directive, and Roderick believes any prosecutions are unlikely in the near future.

Ultra-low power 2D mesfet nearer production

University of Virginia's research group that recently demonstrated an ultra-low power two dimensional mesfet has moved a step closer towards a production version. The remarkable characteristic of the original device is that it operates properly with only 200 electrons in its channel. Conventional fet structures are swamped by second order effects at such low currents.

The advance is in the construction of the device. The developmental fet was made using a AlGaAs/InGaAs/GaAs heterostructure, built up layer by layer on the substrate and patterned using an electron beam. Bill Peatman, a member of the team said: "This technique is not at all suited to mass production."

The newest version is constructed using ion implanted bulk n-doped GaAs - a conventional technique and material. Peatman said: "We made two transistors with the same geometry, one with the original construction and one by ion implanting bulk GaAs. The two behave very similarly, and were closely predicted by our theoretical models."

The channel of the 2D mesfet is a 0.4 µm wide strip connecting the drain and source. The strip is only a few nanometres thick, hence the 2D label. Gates on either side of the channel control the width of the conduction region within.

Wireless LAN technology connecting handheld pcs - including one from SAIC the UK ruggedised computer supplier - is being tested for electromagnetic compatibility in Russia's Mir space station. A 2.45GHz spread spectrum wireless LAN from US company Proxim was taken up to the space station on the shuttle Endeavor in March. According to Yuri Gwaldiak, a NASA computer engineer: "Today's astronauts are surrounded by miles and miles of cable. Wireless networks eliminate their need."

One important aspect of the EMC tests will be the system's susceptibility to the impact of cosmic particles on processors and memory devices.
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Plastic lights

Cambridge Display Technology, CDT, is to give the first public demonstration of its light emitting polymer display in San Diego. The forum is the Society for Information Display's annual conference and the device is a 60 column by 16 row photo-emissive dot matrix.

The display is said to combine characteristics of leds and lcds. Mark Gostick, a spokesman for CDT, said: "The technology has the emissivity of leds, with the patternability of lcds. Light is emitted wherever activated electrodes cross one another."

Although the demonstration is a dot matrix display, the first commercial use is likely to be as a backlight. Gostick said: "The life to 50% brightness is currently 3000 hours. This is perceived to be a bit short for a display, but easily long enough for a backlight.”

Light emitting polymer backlights offer similar advantages to electroluminescent materials. They are extremely thin, with the same emissivity and a similar lifetime.

The stated advantage of light emitting polymers is that they only require several volts, eliminating the inverter normally associated with electroluminescent lights. Gostick said: "The lack of an inverter makes a light emitting polymer backlight cost less to produce than the equivalent electroluminescent panel.”

Most of the development displays made by CDT so far have been green and on glass substrates, but other types have been made. Gostick said: "We have made red and blue versions, the blue ones are the most difficult to produce. We have also made some flexible displays on plastic film, but the lifetime is not as good as the glass ones yet.”

Steve Bush, Electronics Weekly

Satellite tracker handles big waves

Receiving satellite transmissions at sea presents difficulties, with the signal being lost when the boat turns, pitches and rolls. This can be overcome by gyroscopically directed dishes, but the comparatively low power of satellites has meant that these have had to be large.

The launch of three high power digital satellites by GM Hughes Electronics has enabled KVH Industries of Rhode Island to produce a compact system capable of operating up to 200 miles off the US coast.

The TracVision actively stabilised antenna has an 18in carbon fibre dish guided by a robotic arm to maintain tracking accuracy to within 1°. It uses a digital, Earth-referenced compass and attitude sensors to hold directional information over long periods of time, with digital gyroscopes to provide instant directional information.

Measurements are relayed to the cpu, where software calculates the rate of the movements and translates them to stable, land-based coordinates. These are then converted to commands in the motor control unit to guide the robotic arm.

Two motors power the arm and antenna in the opposite direction to the movement of the vessel. It is capable of locating the target satellite with or without GPS information. The elevation range is ±110° from vertical, azimuth range is 360° continuous, and the tracking rate is 18°/s at horizontal.

Received video signals are passed to a set-top receiver box. A patented robotics gimbal joint system prevents cables from wrapping around the antenna pedestal and causing transmission breaks.

A smaller Tracphone version has also been produced. It has an 11.5in dish for satellite phone and fax communications. With this version, internal sensors measure the heading, pitch and roll of the vessel relative to the satellite. These are then transmitted to a microprocessor that directs the motors controlling the arm and dish.

Steel Bush, Electronics Weekly

IN BRIEF

Power fet resistance lowered to 260mΩ

Motorola has announced its next generation of power fet technology – hdtmos-2 – which gives a 30% reduction in Rds on compared with the original hdtmos process.

Significant structural changes have been made to the 1.5mm±2mm chip down from 800m±2 to 260m±1. Sample devices are expected in the second quarter of this year.

Chip prices softening

Softening chip prices could be hit further as 14 new fabs come on stream before the end of June, and at least 16 more are due to start up in the second half of the year. The total for the year could be as many as 40 new fabs.

These are in addition to 60 fabs which, according to US analysts, came on-stream last year: 19 in America, 12 in Japan, 11 in Europe and 18 in Asia.

FCC considers Internet radio band

The US Federal Communications Commission is considering setting aside a band of radio frequencies for high speed Internet access over short distances. The FCC says it may set aside two bands of frequencies in the 5.15-to-5.35GHz and 5.725-to-5.875GHz bands for use by wireless computer devices that would send and receive data over a 0.25 mile range at rates as high as 25Mbits/s.

"Low-power radio technology can serve as a low-cost, high-bandwidth on-ramp to the information superhighway,” said FCC Commissioner Susan Ness. The FCC has been petitioned by Apple Computer and the Wireless Information Networks Forum to set aside the frequencies.

Apple and other supporters say these radio frequencies could be used for campus style local area networks. These would operate over a university or corporate campus linking computer users, and would be a cheap alternative to cellular phone or PCS networks.

Apple is also proposing a related technology that would allow high speed wireless data communications over a 6 mile range at 1.5Mbits/s. Using wireless technology could prove popular within corporate offices and save on the considerable expense of rewiring offices for high speed networks.
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Energy technology is truly green

Plants can exploit virtually every photon they absorb to create a working electron that is efficiently taken up into chemical processes. Our own efforts in making electricity from the Sun using solid state devices are a lot less efficient. But a team from University of California and Princeton University looks to have taken a step towards developing a technology that uses a mechanism much closer to that used in nature.

The technology is based on materials given the name 'chemophylls' because of the way they mimic the chlorophyll process. A prototype chemophyll is a stable preparation that binds a layer of electron-donor substance (papd) to a layer of an electron acceptor substance (pv). The donor papd can be grown as true film, without crystalline structure, while the acceptor pd forms extremely small crystals that can be neatly arrayed in layers one molecule deep.

Both papd and pd consist of an active core sandwiched between structures called phosphonates. Atoms of zirconium bind to the phosphonates, tying the assembled 'sandwiches' to one another. The structure resembles a stack of papd and pd sandwiches piled on an ultra-thin sheet of gold foil. Sandwiches with papd 'filling' alternate with sandwiches with pd filling, and zirconium atoms hold them together.

When illuminated with light, electrons move from the papd to the pd, creating an electrical potential. By providing an appropriate receptor material to accept the electrons on the other side of the surface, a continuing, one-way flow is set up.

University of Southern California chemist Mark E Thompson, leader of the scientific group responsible for the prototype's creation, explains the possible application of the chemophyll: "While this material would be inefficient at powering a device, like a radio or a hair dryer, that runs on electric current, it could be an extremely effective power source for light-induced chemical reactions, such as breaking down water into oxygen and hydrogen or making natural gas out of the carbon in carbon dioxide to create clean-burning fuels."

Three characteristics make it particularly interesting. It is produced in extraordinarily uniform layers and can deliver its energy evenly throughout an entire surface of arbitrarily large size. It can also be applied as a coating to an irregular surface.

Like chlorophyll - and unlike silicon cells - its photovoltaic action is a wet process, working in an electrolyte solution conducive to chemical reactions. It might even float on top of a liquid that would form the raw material for the chemical reaction.

Finally, although the conversion efficiency of the material so far prepared is low, the team is confident that materials of similar structure with much higher efficiencies can be made.

More information from Mark Thompson, Associate Professor, Department of Chemistry, University of Southern California, Los Angeles, CA 90089-0744, USA Tel: (213) 740-6402 Fax: (213) 740-0930 email: mthompson@cheml.usc.edu

Micromachines get a little intelligence

Several research teams have recently made breakthroughs in the manufacture of micromachines. Now researchers at Sandia National Laboratories, Albuquerque have taken the next step by giving their micromachines built in intelligence.

Unlike previous machines, an intelligent micromachine can signal for more power, communicate that it is operating too fast or slow, or even perform actions on an automated basis. But using standard fabrication techniques, machines consisting of tiny motors with integrated circuit 'brains' on individual silicon chips - are now being mass produced by Sandia researchers.

Compact design, made possible by sinking the motors in tiny etched trenches, enables the fabrication of entire electromechanical systems on a chip.

In the past, the difficulty with joining a microcircuit to a micromachine on a silicon chip has been that aluminum circuit interconnectors, if formed first, melt when the micromachines are heat-treated. But if the micromachines are fabricated first, their elevation above the chip surface creates bumps that distort the delicate process of etching accurate microcircuits.

In the Sandia process tiny trenches are etched in silicon chips and the machines are fabricated within these depressions. The machines, heat-treated, are then submerged in a tiny hardening sea of silicon dioxide. The hardened silicon dioxide re-creates a level chip surface upon which circuitry is fabricated by
Optical switch polarises arguments

A light-activated optical switch under development at the Georgia Institute of Technology could be the basis for a new type of rewritable three-dimensional data storage system. By making use of a small number of 'trigger molecules' to induce a phase transition in liquid crystal materials, the system would write, read and erase information using different forms of polarised and unpolarised light.

Such an optical storage system could offer significant advantages over conventional computer floppy disks, magnetic tape and compact disks, which use two-dimensional media to store data. The optical switch materials could also be used in spatial light modulators, and in active coatings for optical fibres.

The idea is that the liquid crystal could be written to, with circularly-polarised light, read from, with linearly-polarised light, and erased, with unpolarised light.

Operation of the new optical switch is based on chiral molecules that Gary Schuster, professor of chemistry at Georgia Tech, and co-workers are using to trigger changes in the liquid crystal. Chiral molecules exist in right-handed and left-handed forms. Each form is affected differently by circularly-polarised light — which also exists in right-handed and left-handed versions.

When right-handed trigger molecules are struck by left-handed light they may be converted preferentially to left-handed molecules.

If the chiral molecules are dissolved in a liquid crystal material, this structural change can be used to prompt a phase transition in the crystal, altering the optical properties of the liquid crystal material.

Multiple phase transition 'switches' could therefore be used together to store digital information.

Returning the storage material to its original state would make the system truly rewritable — and of significant potential value as a computer data storage media, for example.

Schuster believes the system is an improvement over earlier optical switches not only because it is rewritable, but also because a small number of photons can trigger the phase transition. This makes the liquid crystals amplifiers for the photonic signal.

Before the switches become useful for optical data storage, they must be converted from two-dimensional layers to a true three-dimensional system, perhaps written by three-dimensional holograms.

Development of practical optical switches has long been frustrated by the lack of suitable materials. Optical materials studied earlier used irreversible photochemical changes to store information. That meant data could be written to them only once, limiting their practical value for computer information storage.

Though the Georgia Tech system shows promise, Schuster emphasises that much work remains to be done before it could reach practical application.

"From a scientific point of view, looking off into the future, this is what we hope to do with these materials," he said. "Our grandchildren might see the first computers based on this system."
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Researchers pack more power into electric motor

Two researchers in the Department of Electrical and Electronic Engineering, University of Hong Kong, claim to have invented a new design of permanent magnet motor that promises much higher power densities and efficiencies than other similar units. The permanent magnet synchronous motor drive has been developed for electric vehicle applications, with the greater power density resulting in savings in energy and space.

Improvements in the motor’s performance have been bought about by developing a special rotor geometry and using new neodymium-iron-boron magnets that have only recently become available.

In the past, efforts to boost the power density of pmsms have relied on increasing the flux density in the air gap – resulting in a smaller effective gap and leading to a strong armature reaction which limits the force density of the motor.

But the Hong Kong researchers say they have reduced this armature reaction considerably by making air slots along the D-axis within the motor (‘An advanced permanent magnet motor drive system for battery-powered electric vehicles’, CC Chan and KT Chau, IEEE Transactions on Vehicular Technology, Vol 45, No 1). According to Chan and Chau the improvement is possible because in the new design, the air slot at each pole centre has to be bridged by the armature field while the flux path for the excitation field experiences only an insignificant change. Power density of the experimental motor is 209kW/m³.

So far a prototype of the 3.2kW battery powered drive system has been designed and built for an experimental vehicle. Specifications for the vehicle are a top speed of 30km/h, an acceleration of 0-20km/h in 8s and a payload rating of 200kg, able to be carried up a 15° gradient.

In principle, the proposed pmsm could be scaled up to fit the needs of a regular on-road electric vehicle with a 50kW motor. But the researchers warn that any economic advantages would be diminished because of the need for bigger quantities of the relatively expensive magnetic material.

More information from the authors at University of Hong Kong, Pokfulam Road, Hong Kong.
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This article introduces a radio communication system which allows two or more computers to communicate with each other over a wireless link via their RS232 interfaces. Reliable communication can be achieved over a distance of 30 metres in buildings or 120 metres on open ground.

The radio link uses Radiometrix low power 418MHz uhf fm data transceiver modules which are type-approved to the Radio-communication Authority Specification MPT1340 in the U.K. This avoids the need to submit the final product for further approval.

By implementing suitable software, this system can form a local computer network. It can also be used for remote sensing or remote control applications.

**How the system works**

The system consists of a number of identical radio transmitter/receiver units, each connected to the RS232 interface of the various computers, Fig. 1. Each unit can be configured as a transmitter or a receiver. When it is configured as a transmitter, serial data from the RS232 interface is fm modulated by a uhf radio-frequency carrier signal and is transmitted to the surroundings. If it is configured as a receiver, radio signals picked up by the antenna are demodulated. Demodulated data – the serial data put into the transmitter – is fed into the RS232 interface.

A system may include one master unit and several slave units. By using appropriate software, half-duplex data transmission can be achieved between the master unit and the other units, Fig. 1.

**RS232 interfacing**

The RS232 is an industrial standard bi-directional asynchronous serial data communication interface for data communication between two devices. With computers, it is frequently used for connecting peripherals such as printers, modems and mice.

A pc’s parallel port comprises eight data lines and information is conveyed one byte at a time. Unlike the parallel i/o port, the serial COM port has only one line for data transfer in either direction. Each eight-bit byte is transmitted or received serially, one bit at a time.

With asynchronous data transmission, the transmitted signal contains both data and synchronisation information. This synchronisation information enables the receiver to reassemble, or ‘frame’, the serially transmitted words correctly back into parallel form.

The format of the transmitted serial data includes a start bit, 7 or 8 serial data bits, a parity check bit and 1 or 1.5 stop bits. Figure 2 shows a typical serial data format. The receiver device, which runs at the same clock frequency as
the transmitter, detects the start bit and receives the data bits. It checks the parity bit and upon receiving the stop bit it terminates the data receiving cycle and waits for the next transmission. The rate at which data bits are sent is measured by baud rate.

In practice, the asynchronous communication is facilitated by a family of industrial standard computer peripheral ICs known as uarts, or universal asynchronous receiver and transmitters. Most computers use a 8250 or 16450 uart. These devices operate at ttl voltage levels but RS232 operates at higher, bipolar voltages. For this reason, ttl-to-RS232 and RS232-to-ttl converters are needed.

These devices operate at ttl voltage levels but RS232 operates at higher, bipolar voltages. For this reason, ttl-to-RS232 and RS232-to-ttl converters are needed.

The pc's RS232 interface
A standard RS232 interface is a 25-pin interface, which is housed in a 25-pin D-type male connector. A 9-pin version is also used on pcs. Figure 3 gives the pin layout and functions of the RS232 connectors viewed from the back of the computer. The functions of the pins are briefly described as follows:

Prot – protective ground line. This pin connects the metal screening of the cable to the chassis of the equipment.

GND – Ground line provides a common voltage reference for all signals.

TD – Transmitting data output. Serial data is transmitted on this line.

RD – Receiving data input receives serial data.

RTS – Request to send is a handshake output indicating that a transmitting device is ready to send data. If handshake is not required, it can be used as an output.

CTS – Clear to send is a handshake input from which a receiving device is informed to receive data. If handshake is not used, it can be used as an input.

DTR – Data terminal ready is a handshake output and indicates that a transmitting device is ready. If handshake is not used, it can be used as an output.

DSR – Data set ready is a handshake input from which a receiving device is informed that the data set is ready. If handshake is not used, it can be used as an input.

An pc-compatible computer can have up to four RS232 interfaces. They are labelled COM1 to COM4. Each COM port is associated with a uart chip inside the computer. Operations of the COM port are controlled by internal registers of the uart. For register functions, see Table 1.

The i/o addresses of each internal register is calculated by adding the offset to the base address of a COM port. Table 1 shows the i/o address for COM port 1 which has a base address of 3F816.

Some registers are used for the wireless link described here. Register 0016 (data buffer register) stores the received data and the data to be transmitted. Register 0416, the modem control register, controls the status of RTS and DTR of the port. When handshake is not required, the two lines can be used as outputs. Register 0616, the modem status register, stores the status of CTS and DSR, which, when handshake is not used, can be used as two inputs.

Base i/o addresses for COM1 to COM4 are summarised in Table 2.

Table 1. Functions of the pc uart's internal registers.

<table>
<thead>
<tr>
<th>Offset</th>
<th>Address</th>
<th>Function</th>
<th>Description</th>
</tr>
</thead>
<tbody>
<tr>
<td>0016</td>
<td>3F816</td>
<td>Receiver buffer register</td>
<td>Acts as one-byte memory</td>
</tr>
<tr>
<td>0116</td>
<td>3F916</td>
<td>Interrupt enable register</td>
<td>Sets the mode of interrupt request</td>
</tr>
<tr>
<td>0216</td>
<td>3FA16</td>
<td>Interrupt identification register</td>
<td>Checks mode of interrupt request</td>
</tr>
<tr>
<td>0316</td>
<td>3FB16</td>
<td>Data format register</td>
<td>Sets the format of serial data transmission (useful register)</td>
</tr>
<tr>
<td>0416</td>
<td>3FC16</td>
<td>Modern control register</td>
<td>Sets modem controls (RTS, DTR, etc) (useful register)</td>
</tr>
<tr>
<td>0516</td>
<td>3FD16</td>
<td>Serialisation status register</td>
<td>Contains information on status of receiver and transmitter</td>
</tr>
<tr>
<td>0616</td>
<td>3FE16</td>
<td>Modern status register</td>
<td>Contains the current status of DCD, RI, DSR and CTS. (useful register)</td>
</tr>
<tr>
<td>0716</td>
<td>3FF16</td>
<td>Scratch-pad register</td>
<td>Acts as one-byte memory</td>
</tr>
</tbody>
</table>

Table 2. Base i/o addresses for the pc's COM ports.

<table>
<thead>
<tr>
<th>Interface</th>
<th>Base address</th>
<th>Addresses for internal registers</th>
</tr>
</thead>
<tbody>
<tr>
<td>COM1</td>
<td>3F816</td>
<td>3FB-3FF16</td>
</tr>
<tr>
<td>COM2</td>
<td>2F816</td>
<td>2FB-2FF16</td>
</tr>
<tr>
<td>COM3</td>
<td>3EB16</td>
<td>3EB-3FF16</td>
</tr>
<tr>
<td>COM4</td>
<td>2EB16</td>
<td>2EB-2EF16</td>
</tr>
</tbody>
</table>

Fig. 2. Format of serial data generated by uarts. This example has eight data bits (data bits 0 to 7), one parity check bit and one stop bit. The start bit is for synchronising the receiving devices.

Fig. 3. Pin layout and functions of the pc's RS232 connector.
Connecting devices using RS232

Two types of RS232 link are shown in Figure 4. The arrows show the direction of data flow. Figure 4a) is known as a null modem. Some lines of the two RS232 interfaces are used as handshakes between the two devices.

Figure 4b) shows a connection using only three lines. One line is for transmitting data and the other for receiving data. The connection is arranged so that the transmitting line of the first device, TD, is connected to the receiving line of the second device, RD.

Software control

Before a COM port can be used, it must be configured. This configuration includes setting of the following: the bit rate, length of data bits, number of stop bits and the parity check bit. Two computers communicating with each other must have the same configuration. There are three methods of carrying out the configuration.

- The first method is to use DOS command ‘MODE’ under DOS prompt. The syntax of the command is:

  ```plaintext
  MODE COMn: baud=b, parity=p, data=d, stop=s, retry=r
  ```

- The second method is to use a basic-input-output-system, or BIOS, interrupt. This allows the interface to be configured from within the user’s program. It requires that register AH is loaded with 0, DX is loaded with a number 0 to 3 representing COM1 to COM4 to be configured, and AL is loaded with an 8-bit initialisation code. The bit function of this code is shown in Table 3.

- The third method configures the COM port by writing data directly into the data format register, at offset=0316, of the uart. This method is slightly involved and outside the scope of this article. Details can be found however in other pc hardware books.

Table 3. BIOS initialisation bit functions.

<table>
<thead>
<tr>
<th>Bit 7</th>
<th>Bit 6</th>
<th>Bit 5</th>
<th>Bit 4</th>
<th>Bit 3</th>
<th>Bit 2</th>
<th>Bit 1</th>
<th>Bit 0</th>
</tr>
</thead>
<tbody>
<tr>
<td>Define bit rate bits</td>
<td>Define parity check</td>
<td>Stop Bit</td>
<td>Data length</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>111= 9600 011= 600</td>
<td>00= No parity</td>
<td>0= 1</td>
<td>010= 7 bit</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>110= 4800 010= 300</td>
<td>10= No parity</td>
<td>1= 2</td>
<td>11= 8 bit</td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>101= 2400 001= 150</td>
<td>01= Odd</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>100= 1200 000= 60</td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

Fig. 4. RS232 pin connections between pcs and external devices.

Sending and reading serial data

There are several ways to read and send serial data via the RS232 interface. The following method is the most flexible one as far as pc interfacing is concerned. The method is known as direct port access.

To send data out of the COM1 interface, you can write data directly into the data buffer register, at offset=3F8H. The following instructions can be used:

```plaintext
OUT 3F8h, X IN BASIC, and PORT[$3F8]:X IN Turbo Pascal
```

X is the data in decimal. To read data from the COM1 port, we can read data from the data buffer register, 3F816 - the base address of COM1. The following instructions can be used:

```plaintext
Y=INP[$3F8] in BASIC, and
Y:=PORT[$3F8] in Turbo Pascal
```

where Y is the input byte in decimal.

Reading and writing data via handshake lines

To output data from the RTS and DTR lines, you should write to the modem control register at offset=0416. Data bits DB0 and DB1 correspond to RTS and DTR. The following Turbo Pascal command makes the RTS and DTR lines of port COM1 go low. Note that both lines are inverted by ttl-RS232 transceivers inside the pc.

```plaintext
dx:=0; {COM1 is to be initialised}
```

Functions of specifiers in the command can be found in any DOS handbook1. For example ‘MODE COM1: 96,n,8,1’ configures COM port 1 to have a baud rate of 9600, no parity check, 8 bit data length and 1 bit stop bit. This command can be included in the autoexec.bat file.

Procedure initialise:

```plaintext
Procedure initialise;
  begin
  with register do begin
    VAR register:registers;
    var register:registers;
    begin
      AH:=0; {load interrupt function number}
      AL:=128+64+32+0+0+0+2+1; {load initialisation code, 1100011B}
      dx:=0; {COM1 is to be initialised}
      intr($14, register); {Call the BIOS interrupt}
    end;
  end;
end;
```

- The second method is to use a basic-input-output-system, or BIOS, interrupt. This allows the interface to be configured from within the user’s program. It requires that register AH is loaded with 0, DX is loaded with a number 0 to 3 representing COM1 to COM4 to be configured, and AL is loaded with an 8-bit initialisation code. The bit function of this code is shown in Table 3.

The following Turbo Pascal program shows how to achieve the same function as the DOS command ‘MODE COM1: 96,n,8,1’. The initialisation code is 11100011B.

```plaintext
Procedure initialise;
  begin
    with register do begin
      begin
        AH:=0; {load interrupt function number}
        AL:=128+64+32+0+0+0+2+1; {load initialisation code, 1100011B}
      end;
    end;
  end;
```

- The third method configures the COM port by writing data directly into the data format register, at offset=0316, of the uart. This method is slightly involved and outside the scope of this article. Details can be found however in other pc hardware books2.
To read data from DSR and CTS lines, you should read the modem status register at offset 0616. Data bits DB5 and DB4 correspond to the DSR and CTS lines of the port. Again, these two lines are inverted by ttl-RS232 transceivers.

COM port base addresses
The base addresses for different COM ports have been shown earlier. A convenient way to find the addresses automatically is to use software commands. When the computer is switched on or reset, the BIOS checks all possible RS232 addresses. If it finds an installed one, it writes the addresses of the port, in a two-byte word, to specific memory locations. For COM1, the locations are 0000:040016 and 0000:040116. By peeking these locations, the base address can be obtained. Memory locations for COM1 to COM4 are listed below.

<table>
<thead>
<tr>
<th>Port</th>
<th>Memory address</th>
</tr>
</thead>
<tbody>
<tr>
<td>COM1</td>
<td>0000:0400 - 0000:040116</td>
</tr>
<tr>
<td>COM2</td>
<td>0000:0402 - 0000:040316</td>
</tr>
<tr>
<td>COM3</td>
<td>0000:0404 - 0000:040516</td>
</tr>
<tr>
<td>COM4</td>
<td>0000:0406 - 0000:040716</td>
</tr>
</tbody>
</table>

Another useful one-byte memory location is 000:401116. It stores the total number of installed COM ports. The information is contained in bits 3, 2 and 1 of the byte. Data bits 7 to 4 and 0 are used for other purposes.

DB3  DB2  DB1 Number of RS232 ports installed
0    0    0            0
0    0    1            1
0    1    0            2
0    1    1            3
1    0    0            4

The following example in Turbo Pascal 6 first detects the number of RS232 interfaces installed on the pc and assigns the number to a variable, namely 'Number_of_COM'. It then reads from the memory locations holding the base address of COM1 and assigns the address to the variable 'COM1_address'.

```
Procedure detect_COM1;
var
  COM1_address, number_of_COM:integer;
begin
  number_of_COM:=mem[0000:0401];  (*read number of parallel ports*)
  number_of_COM:=number_of_COM and (8+4+2)) shr 1; (*DB3, DB2 and DB1 extracted from the byte*)
  COM1_address:=memw[0000:0400];  (*Memory read procedure*)
  writeln('Number of COM installed: ', number_of_COM);
  writeln('Addresses for COM1: ', COM1_address);
end;
```

Data transceiver module details
This wireless link uses wireless data transceiver modules supplied by Radiometrix. I recommend that you obtain a full descriptions of the modules before you try to implement them.

The modules are available in two types, namely BiM-418-F and BiM-433-F. The first operates at 418MHz and is type-approved to MPT1340 of the Radio Communication Authority in the UK. The latter is for European use in the 433.92MHz band.

Both modules provide a low-cost solution for a bi-directional half duplex data transmission at speeds up to 40kbit/s over a distance of 30 metres inside buildings and 120 metres on open ground. The block diagram of the module, its mechanical dimensions and pin functions are shown in Figs 5&6.

The module consists of a uhf fm transmitter and a matching superheterodyne receiver. For the transmitter, digital data is fed into a modulation lineariser via an R/C low-pass filter. This filter restricts the bandwidth of the modulation signal.

Output of the lineariser drives a varicap diode, the changing capacitance of which modifies the frequency of a uhf...
RF DESIGN

oscillator. The centre frequency of the oscillator is derived from a surface acoustic wave (SAW) resonator. The modulated uhf signal is amplified by a tuned buffer amplifier and fed to a Tx/Rx switch.

After passing through the switch, the signal is emitted to the surroundings via the antenna. For the receiver, the uhf signal picked up by the antenna is switched to the receiver section. It is amplified and fed into the first mixer. After that the signal is amplified and fed into the discriminator, where an audio signal is produced. The signal then passes through a data slicer, converting the analogue signal to digital form.

Pin-functions of the Tx/Rx module
Pin functions of the module are briefly described as follows. Pins 9, 10 and 18 are the ground pins, i.e. 0V, which are connected to the negative rail of the power supply. Pin 17 is the positive supply pin, $V_{CC}$.

A dc supply voltage between 4.5V to 5.5V should be connected. Applying a voltage above 5.5V or reversing the polarity of the power supply will cause permanent damage to the module. When the module is in transmit and receive modes, current assumption is about 12mA. In the stand-by mode, current reduces to about 1μA.

Pin 14 is the transmit data input pin (TXD). It can be driven directly by cmos logic running on the same supply voltage as the module. Analogue signals generated by modems or dtmf encoders can also be fed into this pin.

Pin 12 is the output of the received data. It can be connected directly to cmos logic. Pin 13 is the output of the analogue signals. It can be used with modems or dtmf decoders. Carrier detect is output on pin 11, -CD. When the module is in receive mode, a low state on -CD indicates a signal above the detection threshold is being received. This output can only drive one COM ports logic input.

Pins 15 (-TX) and 16 (-RX) are used for selecting operation modes of the modules. They could select one of the four modes listed below.

<table>
<thead>
<tr>
<th>Pin 15 (-TX)</th>
<th>Pin 16 (-RX)</th>
<th>Modes</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>1</td>
<td>Stand-by</td>
</tr>
<tr>
<td>1</td>
<td>0</td>
<td>Receive</td>
</tr>
<tr>
<td>0</td>
<td>1</td>
<td>Transmit</td>
</tr>
<tr>
<td>0</td>
<td>0</td>
<td>Self test loop</td>
</tr>
</tbody>
</table>

Pins 1 and 3 are the rf ground. They are internally connected to Pins 9, 10 and 18 and should be connected to the ground plane of user's pcb board against which the antenna radiates. Pin 2 connects to the antenna.

Antenna
Three types of integral antenna are recommended and approved for use with these modules. The configuration of the antenna and selection chart are given in Fig. 7. The present system utilises a helical type antenna.
Type approval
The BiM-418-F is type-proved to the RA MPT1340 for license exempt use within the UK for telemetry, telecommand and in-building security — but only provided that the following requirements are met:

- The transmitting antenna must be one of the three variants given above.
- The transmitter module must be directly and permanently connected to the transmitting antenna without the use of an external feed. Increasing the rf power level by any means is not permitted.
- The module must not be modified nor used outside its specification limits.
- The module may only be used to send digital data. Speech or music is not permitted.
- Equipment in which the module is used must carry an inspection mark located on the outside of the equipment and clearly visible, the minimum dimensions of the inspection mark must be 10 by 15mm and the letter and figure height must be not less than 2mm. The wording must read: "MPT 1340 W.T. LICENSE EXEMPT".
- The trimmer control on the module must be inaccessible to the end user. This control is factory set and must never be adjusted.

Failure to meet the above conditions invalidates the modules' type approval. Further information on MPT1340 specification issued by the RA (DTI) may be obtained from the RA’s library service on +44-(0)171-211-0211.

Requirement for digital data transfer
The data path through a pair of BiMs is ac coupled and there are several requirements for successful data transfer. Pulse width time — the time between two consecutive transitions in the serial code — must be between 25μs and 2ms. Receiver BiMs require at least 3ms of 10101010 bit-sequence preamble to be transmitted before the actual data is transferred. The receiver is optimised for data waveforms with 50:50 mark-space averaged over any 4ms period. It will work reliably for sustained asymmetry up to 30/70 either way. But this will result in pulse-width distortion and decreased noise tolerance.

RS232 radio link hardware
The system consists of a number of identical units which are connected to computers via the RS232 interface. The transmission rate of the system is 9600 baud.

Figure 8 is the complete circuit diagram. It consists of three blocks, namely, the RS232/COM driver unit, the BiM-418-F module and the power supply system. The first unit is built around the TC232 RS232 driver, IC1. This IC requires a single 5V power supply and converts the right voltage levels for the RS232 standard and COM port logic.

One of the RS232 output lines, DTR, is converted to the COM-port logic level using a simple voltage clamp circuit using a 5.1V zener diode. All the signal lines between the TC232 and the BiM-418-F are buffered by the 4503BE, IC2. The pins of the BiM and RS232 are connected as shown in Table 4.

The 78L05 5V voltage regulator requires between 8 and 15V dc input. Suggested component layout and assembly of the unit is shown in Fig. 9.

Software requirements
RS232 serial data can be transmitted at 4.8 to 38.3kbit/s between a pair of BiMs. In order to send RS232 serial data through the BiMs, the data needs to be packetised to meet the Tx/Rx module requirements. The packetised data includes the following parts:

- 3ms of preamble data of 5516 or AA16 to allow the receiver BiM to settle.
- one or two bytes of FF16.
- one byte of 0116 to show the start of data.
- data bytes and
- check bits.

In practice, the format of the packetised data may vary according to the users’ needs.

There are three methods of improving the mark-space ratio of the serial data to be transferred.

Method 1. Each byte is divided in half. The first half is the bit to be sent and the second half is its complement. Each byte has a guaranteed mark-to-space ratio of 50:50.

Method 2. Amongst the 256 possible eight-bit codes, 70 codes contain four zeros and four ones, each with a 50:50 mark-to-space ratio. Examples of these are, 1716, 1B16, 2716 and E816. They can be transferred between two RS232 ports using a data format of one start bit and one stop bit with no parity check bit. The actual data to be sent will be translated using these codes and then transferred. This also allows byte checking on receipt as all received codes must contain exactly four ones and four zeros.

Method 3. Each byte is sent twice. The first one is the true data and the other is its complement. This will again give a 50:50 mark-to-space ratio.

Control software in Turbo Pascal has been written to demonstrate data communication between two computers via the system.

Technical support
Designers’ kits and assembled units together with demonstration software are available from the author. Please direct your enquiry to Dr Pei An, 58 Lamport Court, Lamport Close, Manchester M1 7EG. Tel: Ans/Fax: +44-(0)161-272-8279. If you would like to obtain more information on the BiMs, contact Radiometrix Ltd, Tel: +44-(0)181-810 8647, Fax: +44-(0)181-810 8648.

References
1. Microsoft DOS 6.0 manual.
3. Radiometrix Ltd. Low power UHF data transceiver module.
4. Radiocommunication Agency (RA), MPT1340, obtained from RA library service, Tel 0171-211 0211.
Fast wireless data transceiver pair for just £60.80

Normally, a pair of BiM418F uhf transceiver modules, as used in Pei An's wireless RS232 link, costs £105.84. But Radiometrix is offering one pair only per EW reader at the special discount price of just £60.80 - fully inclusive of VAT, postage and packing and data.

Operating at 418MHz, these modules are capable of half-duplex data communication at speeds to 40kbit/s over distances of 30m inside buildings or 120m over open ground.

The modules integrate a low-power uhf transmitter and matching dual-superheterodyne receiver together with data recovery and Tx/Rx changeover switching. Requiring a single 4.5 to 5.5V supply, the modules interface directly to 5V c-mos logic.

High data rates and fast Tx/Rx changeover of less than 1ms make the BiM transceiver ideal for high integrity one to one links/multi-node pocket switch networks. Rapid Rx power up - also less than 1m - allows effective duty cycle power saving of the receiver for battery powered applications. For example, current flow is on average 15pA for 1ms on and 1s off cycling.

Typical applications of the transceivers are:
- Medium speed computer networks
- Laptop to pc to printer links
- High integrity wirefree fire/security alarms
- Building environment control/monitoring
- Vehicle alarm systems
- Remote meter reading
- Authorization/access control

UK Version - BiM-418-F
Euro Version - BiM-433-F

Features of the BiM transceivers
- Miniature pcb mounting module
- Licence exempt operation in UK on 418MHz, MPT 1340 (BiM-418-F)
- ETS 300-220 tested for European use on 433.92MHz (BiM-433-F)
- SAW controlled fm transmission at -6dBm erp.
- Double conversion Superhet receiver
- -107dBm receive sensitivity
- Single 4.5 to 5.5V supply <15mA - Tx or Rx
- Half duplex data at upto 40kbit/s
- Reliable 30 metre in-building range
- Direct interface to 5V c-mos logic
- Fast 1ms power up enable for duty cycle power saving
- On board data slicer, supply switches and antenna change over

Order form
To receive your BiM418F transceiver pair with full data sheet, fill in this coupon and send it to Radiometrix Ltd at Clausen House, Perivale Industrial Park, Horsenden Lane South, Greenford, Middlesex UB6 7QE, together with a cheque or postal order. Please note that this offer closes on Friday 30 August. For further information, write to or call Radiometrix on 0181 810 8647, or fax on 810 8648.

Please send one pair, BiM4128F or one pair, BiM433F (433.92MHz)

to
Name
Company (if any)
Address
Post Code
Day time telephone
For which I enclose a cheque/PO for £60.80 payable to Radiometrix.
When Guglielmo Marconi applied for a patent on his wireless telegraphy apparatus exactly a hundred years ago he was in effect announcing the start of an era. Electromagnetic radiation outside the visible spectrum could now be utilised as an aid to human communication. It was the beginning of a new technology and commercial enterprise.

British Patent No. 12039, 'Improvements in transmitting electrical impulses and signals and in apparatus therefor', application date 2 June 1896, was the world’s first radio patent. Queen Victoria was still on the throne.

It wasn’t a discovery and Marconi was more inventor than scientist. Physicists and others had already been experimenting with this form of radiated energy for a decade or more, though without fully understanding what they were dealing with. They certainly had no practical applications in mind. Branly in France, Hughes and Lodge in England, Popov in Russia and Thomson in America, among others, had all detected apparent radiation from electrical discharges—some natural, some artificially generated—but it was not explored much further. The electron was as yet unknown.

Certainly Maxwell had defined the conditions for the propagation of electromagnetic waves—calculating their speed and inferring that light might be one form of them—as early as the years 1861-1865. His Royal Society paper ‘A dynamical theory of the electro-magnetic field’ was to become a scientific classic. But the experimental physicists of the time were either unaware of it or sceptical of what they considered a dubious concept.

Helmholtz in Germany, however, eventually understood the importance of this mathematically-based theory. It was under him, in Berlin, that Heinrich Hertz was then working as an assistant professor. Helmholtz realised that the theory needed experimental verification and invited Hertz to attempt it.

After several years this challenge resulted in the famous series of experiments of 1886 at Karlsruhe technical high school—described in Annalen der Physik of 1887 and 1888.

Hertz’s experiments

Hertz generated electromagnetic radiation with an induction coil, a spark gap and what we now call a dipole aerial. He detected it with a tuneable loop resonator containing another, very small, spark gap. He demonstrated that the radiation had the characteristics of waves—like light, could be reflected, refracted, diffracted, polarised and made to produce interference patterns. His method was to set up standing waves with a sheet metal reflector and locate their nodes with the loop resonator. Thus he validated Maxwell’s theory.

It was Hertz, incidentally, who pointed out the irrelevance of Maxwell’s original notion, derived from analogies, that there had to be some kind of physical medium which the electric and magnetic forces acted upon—an aether. The essential relationships, he said, were in the equations themselves.

Hertz’s laboratory experiments were recognised everywhere. Branly, Lodge and Popov among others immediately repeated them and gave demonstrations to scientific and non-technical audiences. Yet despite the fact that telegraphy existed and some of its pioneers like Morse and Preece had long been seeking a wire-less version, few seemed to realise that the Hertzian waves, as they became known, were offering them a potent new means of electrical signalling. The exceptions were Crookes, who predicted this possibility in 1892, and Lodge, who demonstrated in 1894 that it could be done.

Marconi’s interest piqued

One of the scientists who followed up Hertz’s experiments, however, was Righi at Bologna University. Marconi, then a teenager living near Bologna, had attended his lectures and studied his writings on e-m radiation. This was the start of Marconi’s interest in Hertzian
waves. But it was later, in 1894, after reading more about Hertz's work, that he became fired with the idea of using these waves for communication. Then began the famous series of radio signalling experiments at his home in 1895.

These experiments at Bologna used apparatus broadly similar to that of the physicists who had followed Hertz's work. The transmitter was a keyed induction coil producing high-voltage pulses, a spark gap and a short Hertzian, or half-wave, dipole aerial. The receiver was little more than a dipole aerial with a non-rectifying detector in the form of a Bransly type of coherer - though the term coherer was actually coined by Lodge.

Initially, the ranges achieved were very short - up to about 100 metres. But then Marconi had the idea, suggested by Popov's experiments in detecting distant thunderstorms, of replacing the Hertzian dipole in both transmitter and receiver with an elevated electrode, or antenna, on one side and an earthed metal plate on the other.

**Marconi's important idea**

These modifications gave a tremendous increase in range - eventually, with an 8m high antenna, to about 2.4km. Thus Marconi, or Marconigraph, which later became Wireless World and the present Electronics World. When radio valves and components began to be utilised in other, non-communication, applications, electronics technology was born - eventually to subsume radio - and the journal started on the road to its present form.

**The link with Electronics World**

This journal, which has just completed 85 years of continuous publication, owes its existence to the wireless telegraphy company that was established with the commercial protection of the 1896 patent. Founded as The Marconigraph in April 1911 (see the reproduction of the front cover), it was then mainly a house magazine, but aimed to "acquaint the lay reader with the latest possibilities in connection with this most marvellous invention" - at a price of 2 (old) pence per month.

In 1913 the magazine was sold to an independent publisher and became The Wireless World. When radio valves and components began to be utilised in other, non-communication, applications, electronics technology was born - eventually to subsume radio - and the journal started on the road to its present form.
HISTORY

sions of the oscillating and radiating elements. Though in practice the radiated spectrum is very wide. There are three kinds of spark transmitter, all using induction coils with vibrating interrupters pulsing the primary windings; and one kind of receiver, based on a coherer with resonating aerial elements connected to it.

Morse code signals are sent by a key in the pulsed primary circuit of the induction coil. At the receiver these are read by an ordinary telegraph instrument actuated from the coherer through a relay.

In one of the transmitters the radiator is essentially a Hertzian dipole formed by a horizontal array of metal balls providing spark gaps, as shown in Fig. 1. Other texts describe this as a Righi oscillator. The balls are solid brass, 100mm in diameter. The outer two gaps are about 25mm while the central gap is only about 1mm.

Obviously these spherical electrodes provide much of the oscillator's capacitance as well as being radiating parts of the dipole. Capacitance and dielectric strength between the two middle balls are increased by an oil dielectric held in a compartment. This 'increases the power of the radiation'. Oscillations of 25cm wavelength (uhf) are produced.

To improve the range and directivity of this radiator, the assembly is mounted at the focus of a cylindrical parabolic reflector, made by

Later spark gaps, like this 1905 example, had motors rotating the spheres to avoid the pitting caused by discharges all at one place.
bending a sheet of brass or copper. Fed from an induction coil normally giving a 25cm spark, the system achieves a range of over 3km.

Tuning the receiver with tin foil

Corresponding to this transmitter is a receiver with a resonating half-wave dipole, Fig. 2, placed at the focus of a similar reflector. Here the dipole elements are 12mm wide strips of copper, cut to the required length to tune with the transmission. The appropriate lengths are found with a separate tuning aid – essentially a resonator containing a tiny spark gap, formed by strips of tinfoil which are experimentally cut to various trial lengths.

Two chances at the outer ends of the dipole provide high impedance to rf and isolate the aerial from the electrical wiring. This was done to concentrate the received energy in the resonator and coherer.

In the middle of this receiving dipole is a coherer, or what patent calls a 'sensitive tube'. This comprises a 38mm long sealed glass tube of about 2.5mm internal diameter, containing two silver contact plugs with a gap of about 1mm between them. In this gap lies a quantity of loosely packed metal powder or filings: a mixture of 96% nickel filings and 4% silver filings is recommended. The sealed tube is preferably evacuated.

The coherer is connected in a circuit containing a 1.5V battery and a relay. Initially the loose powder has low conductivity and passes very little current. When the dipole picks up a signal, however, the rf energy causes the filings to cohere and conduct strongly. The resulting current through the coherer then operates the relay and the telegraph instrument.

Mark tap space

Unfortunately the filings continue to cohere after the Morse dot or dash has ceased, so a small vibrating hammer or tapper, similar to an electric bell mechanism, has to be used to unstick the grains ready for the next Morse symbol.

A different kind of transmitter is illustrated by a diagram similar to Fig. 3. Two metal plates are suspended by insulators from a rope and connected through spark gaps to the induction coil to form a Hertzian oscillator. No reflector is used. From the description, the mode of operation is not clear, though obviously the two plates not only form a dipole aerial but also provide a lot of capacitance and hence radiated power. The larger and higher they are, according to the text, the greater the signalling range.

Finally the patent reveals the use of the earthed antenna – the Marconi aerial discussed above – in a transmitter, Fig. 4, and a corresponding coherer receiver. At both stations the single elevated metal plate is suspended from a pole, though kites covered with tinfoil are also successful. This system is advantageous where "obstacles such as many houses or a hill or mountains intervene between the transmitter and receiver."

Again Marconi reports that the higher and larger the antenna plates the greater the range achieved.

Start of a company and industry

Acceptance of this first patent in July 1897 provided the commercial basis for setting up a company that same month to develop and promote the Marconi apparatus. This was The Wireless Telegraph and Signal Company Limited, which eventually became The Marconi Company and finally part of GEC.

Some of the early demonstrations and communication achievements are described and illustrated in the EW&WW article ‘Marconi Beginnings’ in the January 1992 issue, on pages 74-76.

Further reading

The article Demodulation - a new approach - on the following pages is winner of Electronics World's 'Best rf article '95' award - a Hewlett Packard programmable signal generator priced at £4000. Some 30 entries were considered, representing articles both published and unpublished articles offered to EW during 1995 - which is why the judging took longer than planned. All were interesting and there was a healthy shortlist of contenders for first place. But Archie's work stood out in that it represents a technique that truly breaks new ground in the important area of demodulation.

Since the article's publication in the December 1995 issue, Archie has further developed his amplitude-locked-loop and is presenting a paper describing the technique's mathematical model at BSI 96 International at Chertsey on 4 June.

The US Army has been looking at the demodulation system for incorporation into their new pan-forces radio system called 'Speakeasy'. From their initial evaluation, they agree that the technology works "as advertised".

Additionally, DRA Malvern is using the analogue-locked-loop technique in advanced laser dopplerimetry to minimise co-channel interference.

Since publication of the article, eight advances have been made to the high-performance demodulator. Among these, the system's speed has been increased by a factor of five and overall dc coupling is being used to reduce cross-modulation products.

For his rf design article, 'Demodulation - a new approach', Archie Pettigrew wins an HP 8647A synthesised rf signal generator. Fully programmable and operating from 250kHz to 1GHz, this generator is designed for a variety of general purpose applications and semi-automated receiver test. With 300 storage registers and ten user-definable sequences - accessible from a remote keypad - the HP 8647A easily adapts to any test procedure. The generator features HPIB interface, a solid-state programmable attenuator and built-in am/fm modulation capability.
Archie Pettigrew’s new demodulator concept uses amplitude-locked loop techniques to produce significant improvements in the quality of fm and am reception.

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

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

Both these breakdown processes are caused by interference in the form of multi-path Doppler or quasi-synchronous reception. All these forms become worse as frequency of the carrier is increased i.e. as wavelength is shortened and/or as transmission becomes mobile.

By using an amplitude-locked loop and associated circuitry, many of these interruptions can be avoided, and more reliable communications achieved.

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

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

A phased-lock loop is similar in that it consists of a voltage controlled oscillator contained in a high gain, high bandwidth servo loop. By using an amplitude-locked loop and associated circuitry, many of these interruptions can be avoided, and more reliable communications achieved.

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

Fig. 1. Amplitude-locked loop consists of a linear multiplier, modulus detector and a high gain integrator. When the loop is closed, envelope variations of the carrier are reduced to insignificant proportions due to servo action and an error signal called the inverse modulus is produced.
where \( m(t) \) represents the modulating function of time and \( \text{sinar} \) is the normalised carrier amplitude of \( \omega \) radians per second.

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

\[
v_{sc}(t) = \frac{1}{1+e(t)} \text{sinar}
\]

where \( e(t) \) is the loop error voltage which becomes insignificant due to the high open loop gain. That is, \( v_{sc}(t) = \frac{1}{10} \text{sinar} \)

Voltage \( v_{sc}(t) \) represents a stabilised carrier with no envelope variations. Voltage at the second input to the multiplier must therefore be the reciprocal of the input modulation. As a result, \( v_{2}(t) \) is \( \frac{1}{1+e(t)} \) and \( v_{3}(t) \) is \( \frac{-m(t)}{1+e(t)} \) by subtracting unity.

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

Much the same arguments were used for the PLL when it was first suggested. For example, the PLL could easily have been replaced by a piece of wire and at a much lower cost etc. Perhaps for the above reasons the concept of the ALL has never been investigated, even in the valve or tube era of electronics. If the ALL is not directly suitable for demodulation of am, can it be used to replace the limiter-filter in the demodulation process? Indeed it can as will be explained.

**Application to fm demodulation**

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

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

**Operating limits**

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

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

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

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

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

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

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

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

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

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

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

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

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

**Fig. 5. Comparison between a limiter-filter and an amplitude-locked loop demodulator showing that the amplitude-locked loop is capable of extracting a phase coherent carrier well down into noise levels.**
made constant. The PLL must have a wide capture and track range for fast lock-in and frequency tracking, yet it must have an extremely narrow noise bandwidth for stability during every speech silence.

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

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

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

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

**Carrier generation**

A pure squaring device follows the analogue locked loop to generate a coherent carrier at 20. When the ALL is out of lock, the PLL is being driven by a zero level carrier. Since the bandwidth of any linear PLL is a direct function of the input amplitude, its closed loop bandwidth drops to zero.

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

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

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

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

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

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

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

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

**Fig. 8.** Complete FM201 demodulator using an amplitude-locked loop, a phase-locked loop and an analogue signal processor. Inverse modulus from the amplitude-locked signal is multiplied by the phase-locked loop output to produce the impulse alone signal which is scaled and subtracted from the original phase-locked loop signal.
Stability of the voltage controlled oscillator is then determined solely by the time constant of the filter following the phase detector. This can be made large ie of the order of one second. Carrier stability is thus maintained due to this very long time constant. By use of the ALL, phase-capture transients are very short when signal is present and phase loss transients are long when the signal is absent. By this technique, the coherent carrier can be recovered reliably even during periods of poor carrier-to-noise ratio. The circuit diagram of this system is shown in Fig. 4.

The ALL is contained in the hybrid block U₁. The circuits which follow the ALL represent the normal synchronous AM demodulation technique, namely, a pure squaring device (U₂) followed by a narrow track range PLL (U₃), a divide-by-two, (U₄) and finally a synchronous multiplier (U₅).

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

It is interesting to note that there is no threshold effect present as would be the case in fm or angle demodulation. The output signal-to-noise ratio tracks the input carrier-to-noise ratio in a linear manner. Results obtained from the above demodulator exceeded the performance of the normal synchronous demodulator in that carrier recovery could be achieved down to and below unity carrier-to-noise ratios. Figure 6 shows an oscilloscope trace of a 300Hz sine wave signal which is being gated on and off at 6 and 4ms intervals. Carrier-to-noise ratio with signal present was 0dB. Noise and carrier amplitudes were equal.

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

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

**FM demodulation**

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

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

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

The catastrophic fm threshold effect begins and rapid deterioration of the output signal-to-noise ratio then follows. This same effect causes fm reception to be rendered unintelligible if two fm transmissions arrive at the antenna at equal or near equal strength to each other - assuming co-channel frequencies.

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

**Capture effect**

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

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

Frequency modulation works well when;

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

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

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

The Ampsys FM201 demodulator

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

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

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

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

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

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

A new baseband signal is created which is free of harsh spikes and is now perfectly intelligible even when the carrier and interference are identical in magnitude. A simplified version of the relevant waveforms is shown in Fig. 9a-d. Voltages $v_1(t)$ to $v_6(t)$ correspond to those marked on Fig. 8.

Worst-case fm reception

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

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

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

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

- When the interfering frequency is offset from the centre of the passband, similar subtraction can be achieved by inserting an offset voltage at the input to the final multiplier. The interfering frequency however must be fixed.
- When quasi-synchronous reception occurs, baseband signals combine and an improvement in the signal-to-noise ratio of approximately 26dB is measured at the equal amplitude reception point. Multipath distortion causes a small reduction in signal-to-noise ratio.
- When two modulated carriers are present, the result is similar to that of a crossed line in a telephone. Although this is not ideal, it is preferable to complete loss of intelligence.
- With a very weak carrier, harsh spikes are removed and noise subjectively more acceptable. White noise can never be removed since there is never enough unique information.
- When the carrier-to-interference ratio is high, the 'impulse-alone' product diminishes rapidly since there are no envelope variations. The weaker transmission is suppressed as in normal demodulation and all beneficial characteristics of fm are retained, for example, quiescing and co-channel suppression.

To summarise

A new circuit concept has been proposed called the amplitude-locked loop which can be used in conjunction with a phase-locked loop to improve the quality of fm demodulation. By using the fundamental property that fm is transmitted at fixed amplitude, and that a unique relationship exists between the reciprocal of the modulus and the fm phase perturbations, a new signal has been derived called the 'impulse-alone' signal.

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

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

Acknowledgments

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References

Ignoring cost, nearly all power switching applications can benefit from replacing a bipolar output device with a power mosfet. But there is one very common application where this is still not the case. David Sharples explains why.

**DESIGNING CRT deflection**

The number of power applications that still have a bipolar transistor monopoly reduces by the day. The ingress of new technologies, such as high performance mosfets, igbts and power ICs, brings the humble bipolar transistor ever closer to the end of the road.

But there is one very common application where a bipolar high voltage transistor is the only viable power switch: tv and monitor horizontal deflection. As this situation is likely to continue for some time, while crt remains the dominant display technology, it is worth taking an extended look at this application.

**Horizontal deflection circuitry**

To understand the operation of the circuit shown in the panel consider the transistor turned on at time $t_0$, as shown in the waveforms in Fig. 1. DC voltage across the coil will result in a linear ramp in coil current:

$$\frac{dl}{dt} = \frac{V}{L_c}$$

This current, $I_C$, flows through the transistor to ground. At time $t_1$ the transistor is turned off. Turn-off is not instantaneous for a bipolar transistor and the current continues to rise for a couple of microseconds after the forward base current has stopped. This phase of operation is known as the transistor storage time: conduction is maintained by charge 'stored' in the device during the previous on time.

To turn a bipolar transistor off efficiently charge has to be extracted from the base, ie a reverse base current. When the base is depleted of charge to a level that restricts the flow of electrons from emitter to collector, the load will start to see turn-off. In this example, the load is coil $L_e$.

At the end of storage time, $t_2$, the voltage across the device, $V_{CE}$, starts to rise. This rise in voltage reduces the voltage across $L_e$ which in turn reduces the current ramp. With optimum charge extraction during the storage time the transistor collector current, $I_C$, will fall rapidly to zero in around 300ns.

As current in $L_e$ cannot change so quickly, it is diverted to the flyback capacitor, $C_{fb}$. This current causes the voltage on $C_{fb}$ to rise; $L_e$ and $C_{fb}$ operate as an LC circuit: sinusoidal rising voltage, cosinusoidal falling current.

At time $t_3$, current will be zero and a peak voltage will be reached. This is the peak $V_{CE}$ on the horizontal deflection transistor.

Peak voltage now drops as the current is reversed and $C_{fb}$ discharges. If diode $D_2$ was removed the system would become an LC resonant circuit. However, the diode has an important role.

At time $t_4$, the voltage reaches zero the current is a maximum, a negative going voltage forward biases the diode and the capacitor current is re-routed to the diode. The sinusoidal rising and falling voltage coincides with a polarity change in the coil current, this has then facilitated the flyback function.

With a negative peak current the beam is ready to start a scan at the left hand side of the screen. With a steady forward voltage drop across the diode, $V_p$, the voltage across the coil is constant yielding a linear ramp in coil current. As the diode/coil current approaches zero the transistor is turned on again. This early turn-on ensures a smooth zero-crossing as the coil current now becomes transistor current at the left hand side of the screen. And so the cycle continues.

Typical waveforms for the deflection transistor $I_C$, $I_B$ and $V_{CE}$ are shown in Fig. 2.

**Typical device types**

Two of the semiconductor devices shown in the circuit in the panel are standard commodity parts and two are unique to this application.

**Primary switch transistor $Q_1$.** In the circuit in the panel this is shown as a mosfet but in most television circuits a cheaper bipolar is used. For a low primary supply voltage an industry standard bipolar BC337 or 2N7000 mosfet could be used. For high primary supply voltages industry standard types can still be used: BF422 (bipolar) or BS108/BSN254 (MOSFETs).

**Turn-off diode $D_1$.** Most small diodes will fulfil the requirements of this application. A well-used type is the Philips BYD33D.

**Deflection transistor $Q_2$.** Most deflection circuits have a peak flyback voltage of 1100-1200V; to allow for faulty conditions the standard peak voltage of deflection transistors is 1500V. A small 14in television will only require a peak coil current of 2A. For the large screen, home cinema type tvs – up to 36in – peak coil current can be as much as 7A. Over specifying the current is not efficient in this application: a bigger piece of silicon may not be better than a correctly specified device.
Raster scanning basics

In the UK and mainland Europe, conventional TVs have a complete picture change 25 times every second, i.e. a change every 40ms. Each picture is made up of a series of lines, a 625 line system being the current standard. A change of 625 lines in 40ms implies 64ps per line. From this comes the 16kHz - actually 15.625kHz - horizontal, or line, frequency. Each line is produced by the horizontal deflection of the picture tube's electron beam as it scans across the screen, left to right. The time the beam takes to scan from left to right is the scan time, T_{scan}. When the beam reaches the right hand side of the screen it has to fly back to the left hand side before the start of the next scan. Time taken for the beam to fly back is the flyback time, T_{fb}. Typically, the 64µs per line is composed of a scan time, T_{scan}, of 52µs and a flyback time, T_{fb}, of 12µs.

Originating at the cathode gun at the back of the picture tube, the electron beam is accelerated to the screen by a high potential anode, typically 25kV. The beam is deflected horizontally during the scan time by a magnetic field produced by a deflection coil. This field is produced by a current ramp through the deflection coil, as shown above.

This current waveform can be produced in a variety of ways but there is one dominant circuit which is used in over 90% of television and monitor designs, this is shown in its simplest form in the diagram below.

Only a few of the world's major semiconductor suppliers offer a complete range of deflection transistors. Over 95% of the televisions produced will have a deflection transistor from one of these suppliers. Philips types range from the industry standard, and much copied, BU508 to the new BU2530AL.

**Fig. 1. Typical deflection circuit waveforms for the main circuit elements of the circuit in the panel.**

Damper diode D_2. As the damper diode is in parallel with the deflection transistor (see the circuit in the panel) it has the same peak voltage requirements. Also the coil current is symmetrical about 0A, therefore, the current requirements of the damper diode are the same as the deflection tran-
PRODUCTS

Fig. 2. Horizontal deflection transistor current and voltage waveforms found in a typical large-screen television.

Products for this application are offered by most of the major semiconductor rectifier suppliers and some of the deflection transistor suppliers; including Philips Semiconductors. Philips types range from the glass bead BY228 to the BY459 in TO220 style packages.

East/west and S correction

To compensate for the curved surface of the picture tube, and obtain 'true' image reproduction, a number of corrections have to be carried out to the deflection waveforms; a linear ramp in coil current does not produce a watchable image. For horizontal deflection the two important correction functions to be aware of are S-correction and east-west correction.

Given a typical display surface and a linear coil current ramp a pattern of equidistant lines would actually show as close packed lines in the centre of the screen and widely-spaced lines at the sides. The electron beam has to travel further to the edges of the screen than the centre.

To obtain an equidistant display, a ramp that is fast at the centre and slower at the edges is needed. This changes the shape of the coil current from a linear ramp to an S-shaped waveform, Fig. 3, hence the name S-correction.

S-correction is usually achieved by adding a capacitor in series with the deflection coil. In addition, compensation has to be made for the resistive component of the coil impedance. Coil resistance tends to shift the distortion to the right, and to compensate this a series negative resistance is required. This effect is achieved by a saturable inductance in series with the deflection coil.

The need for east-west correction arises from the fact that given the curved surface of the CRT, the electron beam has to travel further per scan at the centre of the screen than at the top and bottom. And the amount of deflection is proportional to current. Adding this correction feature has the effect of applying an envelope to the deflection current with peaks for the centre.

Although the complete picture changes every 40ms, this is done by changing alternate lines so a frame changes every 20ms. Two frame changes produce a complete picture change. The frame frequency, or vertical deflection frequency, is then 50Hz. This 50Hz waveform is superimposed onto the deflection current to produce the east-west correction.

Modulated current is shown in Fig. 4. This is achieved by applying a modulated voltage across the deflection coil.

Generating eht

As well as being the power switch in the horizontal deflection circuit the transistor performs a secondary function – generating the extra-high tension, or eht. The eht circuit is a basic flyback converter power supply with a special high voltage transformer producing the 25kV required at the secondary windings. This transformer is commonly called the line output transformer, or LOT.

The horizontal deflection transistor provides the switching function to the primary of the LOT. For this reason the horizontal deflection transistor is also referred to as the line output transistor. Fundamentally, the eht function adds an offset of up to 1A to the total deflection transistor current, \( I_C \).

Coil current now becomes symmetrical about the high-tension offset, therefore, the damper diode current reduces by this amount.

Adding the eht and correction functions to the basic circuit produces a modified circuit; a version of which is shown in Fig. 5.

IDTV, HDTV and monitors

So far, the discussion has been limited to European stan-
The 50Hz frame frequency of European TV follows the 50Hz mains frequency. For 60Hz mains countries, including the US and Far East, 60Hz frame frequency is used. For a 625 line, 50Hz system the line frequency is 15.625kHz, often referred to as 16kHz. For 60Hz systems fewer lines are employed — frequently 525 lines — yielding a line frequency of 15.75kHz. This difference in line frequency is not significant to the deflection transistor.

New idtv systems offer ‘flicker-free’ viewing with a 100Hz frame frequency, which is less detectable to the eye than 50Hz. In turn the line frequency doubles to 31.25kHz. This change is very significant to the deflection transistor. Current proposals for hdtv systems indicate a 64kHz line frequency that is not significant to the deflection transistor.

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Current proposals for hdtv systems indicate a 64kHz line frequency that is not significant to the deflection transistor.
Fault finding in the deflection drive

Consider a 16kHz tv found to have a failed deflection transistor. Replacing it with an identical type restores full set functionality. If the picture shows some visible distortion then the chances are that there is a fault in the 'load' side of the deflection transistor, ie something causing a change to the IC.

If, however, everything appears ok there may still be a problem with the drive that does not immediately result in device failure. For 16kHz television, a typical deflection transistor should dissipate less than 2W which means that the device can run in free-air with a 25°C ambient. This can be tried in the workshop. If free-air operation leads to device failure or a visible distortion the drive circuit needs to be re-optimised.

This technique will not, of course, resolve the failure if it was caused by a single overstress event. These failure modes could be the result of heating induced picture tube flash, or interference from some other source of electromagnetic radiation. This type of failure mode cannot be diagnosed easily.

a visible distortion on the screen or require some additional correction.

During the on-state the device should be operated in the 'saturation' region of the transistor operating area.

Off-state: The device must be able to withstand any voltage applied across the collector-emitter terminals when there is no positive drive.

Switching: The transition between on-state and off-state must be achieved quickly and without unnecessary dissipation. This is the most critical feature of operation with regard to the device requirements. Unfortunately, the switching characteristics of power devices are very circuit dependent. An appreciation of the interacting variables is essential if an optimised design is to be produced. The 'Base drive optimisation' section offers an empirical path through the maze of information surrounding bipolar horizontal deflection switching.

Using device data

Semiconductor companies usually adopt formats for data which have evolved from in-house rules laid down many years ago. It can be difficult to relate much of the published data to the requirements of the application. But, there are vital pieces of information that can be extracted.

The first parameter to check is peak voltage – the largest voltage value in data. This can be expressed as $V_{cE}$, $V_{CE}$, $V_{CB}$ or $V_{CEO}$. For a solid-state physicist there are interesting differences between these parameters, but for a circuit designer all these can be viewed as the same. Most televisions of 21in and above, have a peak flyback voltage in normal running of 1100-1300V, in which case a 1500V peak voltage device is required.

There are no significant differences in the peak voltage characteristics of devices from different suppliers. This is the simplest data to comprehend.

A lower voltage value, $V_{CEO}$, is also often given in data. The subscript 'O' means the base is open circuit. Because the base is never open circuit in a deflection circuit, $V_{CEO}$ is not of prime importance.

The second parameter to check is recommended operating current, which may not be too easy to find. Some suppliers give an $I_{CS}$ value – ie the $I_C$ in saturation – in the summary at the top of the data sheet. For other suppliers the value may have to be extracted from another characteristic.

All suppliers quote either $V_{CESat}$, $h_{FE}$ or some switching values for a given $I_C$: this can be taken as the $I_{CS}$ value. The $I_{CS}$ value is determined by the chip size and hence has a large influence on the device cost.

There are two possible pitfalls when comparing $I_{CS}$ data: the operating frequency and 'specmanship'. A device specified at 4.5A for 64kHz operation in a pc monitor will have a bigger chip than a device specified at 4.5A for 16kHz operation in a tv. Also, some devices are over-specified on $I_{CS}$. If some devices were operated at the $I_{CS}$ recommended in data they would have to be bolted to a very large, water-cooled heatsink. This is not practical for tv applications.

The only other relevant specifications are the maximum junction temperature, $T_{JMax}$, which is usually 150°C for bipolar deflection transistors, and the thermal resistance values. These specifications are required for base drive optimisation. Initially, at least, all other data is irrelevant.

Base drive design

Base drive design can only be optimised empirically. Advances in circuit simulation models fall short in this application: the available device (HVT and diode) and component (LOT) models are insufficient. This situation is unlikely to improve much in the next five years. Industry standard bipolar HVT models are not possible as the different construction of a 1500V, 5A, 64kHz device from supplier A will require a different drive from a 1500V, 5A, 6kHz device from supplier B.

The bipolar HVT is a current driven device so base current is the important waveform to analyse. A successful design will have matched this $I_B$ waveform to the load ($I_C$ waveform). The method, with a few 'rules-of-thumb', is outlined in the panel 'Optimising base current'.

For small screen tv, with an optimised drive, the power should be 2W or less. For the largest tvs the power could be as high as 3W in the extreme load case.

From the power, the device junction temperature, $T_J$, can be calculated. Thermal resistance between the junction and the heatsink, $R_{jhs}$, will be between 2K/W for a large, non-isolated device to 5K/W for a small, isolated device. So,

$$T_J=T_{JMax}+(Power x R_{jhs})$$

A good rule-of-thumb is that the $T_J$ is 10°C higher than the heatsink temperature.

The standard $T_{JMax}$ for bipolar deflection transistors is 150°C, however, it is not good practice to use this as the limit in the above experiment. The thermal analysis has to be good for a complete production program so it is usual to reset the desired $T_{JMax}$ to a lower value, typically 110°C. This should result in the elimination of any thermal failures during the production run of a tv model and enhanced reliability for the horizontal deflection stage.

Using a small heatsink compensates for not carrying the tests out at an elevated ambient. In the laboratory, the tv may have the back off and the main pcb may be exposed to freely circulating air. In domestic use the set will be placed usually with its back to the wall in the corner of a room. Heat flow in these two situations is significantly different, so some allowance must be made for this in the laboratory measurements.

Glossary of terms

<table>
<thead>
<tr>
<th>Terms</th>
<th>Description</th>
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<tr>
<td>I_C</td>
<td>collector current</td>
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<tr>
<td>V_CE</td>
<td>collector-emitter voltage</td>
</tr>
<tr>
<td>h_B</td>
<td>base current</td>
</tr>
<tr>
<td>h_{Tend}</td>
<td>base current, end of on period</td>
</tr>
<tr>
<td>h_{off}</td>
<td>peak reverse base current</td>
</tr>
<tr>
<td>V_BE</td>
<td>base-emitter voltage</td>
</tr>
<tr>
<td>h_E</td>
<td>current gain: $I_C/I_B$</td>
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Other abbreviations

<table>
<thead>
<tr>
<th>Abbreviations</th>
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<tr>
<td>EHT</td>
<td>extra high tension</td>
</tr>
<tr>
<td>HVT</td>
<td>high voltage transistor</td>
</tr>
<tr>
<td>LOT</td>
<td>line output transformer</td>
</tr>
<tr>
<td>R_{th}</td>
<td>thermal resistance</td>
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Cable shunted

Although Ben Duncan's analysis of loudspeaker cables is highly interesting, it is evident that a shunt capacitance of 228pF (impedance greater than 10kΩ) cannot have any effect on a low impedance load. Indeed, if the amplifier were current driving, the cable modelled would have no effect at all.

A T-network is an excellent approximation of a transmission line at low frequencies (see figure) by linear approximation of the trigonometric functions. The total inductance and capacitance per metre in these are the same as the intrinsic inductance and capacitance per metre of the transmission line.

The inductance parameter seems to be measured at cut-off to a 5.62 load, i.e. at about 1MHz. The corresponding capacitance at that frequency must be about 60pF (or else the dielectric constant is as high as 10.2).

If, due to increasing skin depth, the characteristic impedance becomes much smaller at lower frequencies, the inductance must correspondingly decrease so as to make LC=1/c² constant. Am I right?

Otherwise there would be dispersion, i.e. lower-frequency waves propagating slower, but dispersion has to do with the dielectric which is relatively constant at low frequency. I couldn't find any loudspeaker cables hanging around, so I refrained from making the measurements myself.

Michael Williams
Jerusalem

Delayed reflections

I was intrigued to read Mr Russell's article on Transmission lines in Electronics World March issue. I designed and manufactured the MDN284 48S/SK delay line that you describe, in about 1965. Lexor Electronics of Coventry supplied these things mainly to the English Electric Company, who were Anglicising an RCA Computer and acquiring an interest in the LEO computer set up by J. Lyons. We designed a whole range of delay lines for them. Later the company became English Electric.

The article analyses the basic parameters correctly. However Mr Russell makes no mention of the factor which could possibly foul up the results if not taken into account.

Simply stringing together a cascade of Ls & Cs gives you a 'Constant K' line. This has a group delay characteristic from hell. Delays at low frequencies will be much less than those at high frequencies. Therefore when you double or otherwise multiply frequencies, the apparent wavelengths and thus nodal relationships will not be as expected.

Lexor delay lines were mainly designed for time-domain - pulse-operation. They were used in computers for equalising the delays through various diverging and converging paths in the computer, and a major third parameter following delay and impedance is overshoot and its opposite, undershoot. These are evident in response to a pulse input.

Briefly, when undershoot and overshoot are equal, then all frequencies within the pass-band of the delay line (it can be considered as a type of low-pass filter with a Gaussian response) are delayed by the same amount.

In order to achieve this within the design, you have to provide for magnetic coupling between adjacent sections of the line. Moreover the sign of the coupling should be positive between adjacent section and negative between alternate sections to avoid reflections (poor swr) at the highest frequencies.

In the 2484 series, this could not be contrived due to physical constraints, and all couplings were positive. Since the units were employed solely in the time-domain however this was not a problem.

The optimum coupling in a classic 'm'-derived design requires a 'K' of approximately 27%. This is managed in the 2484 series by windings the coils on ferrite 'cotton-reel' bobbins, and arranging them axially with a closely controlled spacing. Too close results in group delay negative with frequency, and vice-versa.

Values up to 500ns/560Ω are available in 14 pin, 0.3in DIL form these days. However, we still supply them all to order.

J.C. Pledger
Active Electronics Labs (tel. 01926 484050)
Warwickshire

Is oxygen to blame?

I read with great interest Anthony Hopwood's account of Prof. Henshaw's ideas on the causes of some cancers and leukemias in humans, in the April issue. I am sure some of these diseases can result from chemical exposure to ionising radiation. Others from Madam Curie onwards died of leukemia. All these people had one thing in common with people living under power lines and who worked for 26 years on the rust cyclotron in the world to be built in a hospital for medical research. There, alpha particles and deuterons were accelerated up to 15MeV. Even though several of my colleagues died of cancer, I do not believe the alphas caused direct damage to body tissue.

I think the causes of large number of victims of radiation combined with high voltages is best shown by going back to my first employers at Lexor delay lines. Although Ben Duncan's analysis of loudspeaker cables is highly interesting, it is evident that a shunt capacitance of 228pF (impedance greater than 10kΩ) cannot have any...
Ouch!
I would like to make it clear that Bill Russell's April 1996 article on transmission lines has used the ttl devices output for the top trace in Fig. 6. The actual input of the transmission line shows a pulse after 16.8us. The amplitude of this pulse is 2.6V since a third of the returning 4V initial pulse is inverted and reflected back along the line.
The other nit I would like to pick at is: has Douglas Self actually measured his distortion levels with - to use an objective term - music? Music frequently contains peaks that may exceed five times the rms level. Many low distortion amplifiers use high voltage power supplies to supply these short duration peaks.
For uncompromising audio purists I suggest they bridge two of Self's amplifiers together and suffer a SBF loss in the wallet!
Andrew Pate
Surrey

No contest
In response to the 'challenge' in Letters in the May issue, Allen Wright appears to have misunderstood the nature of scientific enquiry. In the famous example, the hypothesis 'All swans are white' is disposed of forever by seeing one black swan. It is not necessary to look at more black swans in different ways to confirm the finding.
Mr. Wright makes no mention of the rather less plausible hypothesis 'All copper conductors are composed of 10mV diodes' it is only necessary to show once that this is not true. I think that my experiment reported in the March '96 issue of EW is proof positive that magical 10mV diodes do not and cannot exist in solid copper. Further evidence is given by my JAES paper 'Granularity Distortion' which tested the matter down to the -150dBu level.
As far as I can gather, Mr. Wright's own hypothesis seems to be 'Thinner speaker cables sound different, and I think they sound better, so this must be because there are fewer magic diodes, as there is less total weight of copper.' Mr. Wright assumes - apparently for no other reason than because he wants to - that any change in subjective quality must be due to magic diodes, and that the highly unscientific procedure he outlines will 'prove' it. It is, I hope, unnecessary to point out the gaping holes in the logic here.
If an audible thin-cable effect exists, then I suggest the reason is as follows: single-strand cable will have a much higher resistance than normal speaker cables. The extra resistance means that the speaker impedance curve will be partially superimposed on the basic speaker/room frequency response. These extra response irregularities - which will not have been anticipated by the speaker designer, who will have assumed that his brain-children are to be driven by a low-Z source - may or may not make the overall sound better according to unguessable subjective criteria.

I may well be that a listener could convince himself that the result is greater clarity and lesser distortion; but he has no right to then state that the difference is due to mysterious effects when a very obvious explanation is right under his nose.
Bringing Occam's Razor.
Mr. Wright makes no mention of double-blind A/B testing in connection with his assertions, so I assume they were conceived under informal listening conditions.
I therefore decline to accept his statements about enhanced clarity and reduced distortion.
While I can only judge by the title, I believe that any book entitled 'SuperCable Cookbook' would upset a lot of people, especially those who feel that technical books should preferably contain facts.
1. 'Ultra-low noise amplifiers and granularity distortion' JAES Nov 1987, pp 907-915.
Douglas Self
Idnston, Herts

Have you any queries?
If you have any electronics-related questions that you have been unable to find an answer to, why not see if other readers can answer them? Simply write to me, the editor, at the address on page 267, fax 0181 652 8956, or e-mail martin.eccles@rbp.co.uk.

I would be grateful for your assistance. In the diagrams is a small transmitter and receiver which include encoding and decoding facilities.

Receiver: When the receiver is switched on, once a signal is received, the signal is processed by the components to the left of the solid black vertical line. The output of this processing is the same code sent by the transmitter which enters pin 14 of IC1. This chip has also been set to the same code as the transmitter. When the correct matching code is received pin 17 goes high to turn on the chime section of the circuit. The section to the right of the solid line is fully understood.

Question 4: How is the signal processed by the receiver in order for the 4069 hex-inverter to work since the frequency of around 340MHz is too high a frequency for the chip to operate?
Question 5: Could it be that the inverters are working as filters? If so how?

Transmitter: The main integrated circuit provides a number of combinations which can be sent by connecting pins 1 to 9 to ground. When power is
offered by Call Waiting. However, the number of such a call is not transmitted. From the data in Mr Segaran's article it appears that such an enhancement to the service is probably not possible; in any case BT has informed me that they do not intend to try.

This is a great shame. What is the 'metering and message waiting status' information that BT would like to transmit without alerting the phone user? I have asked BT to explain the 'message waiting' flag on the display of my CD50 but they cannot be bothered to answer.

I note that the software provided with the CID-PCI is for Windows only. For the benefit of pc owners who, like me, have thrown away their copy of Windows, I would like to have DOS tsr software.

I apologise if this letter sounds negative. I have great enthusiasm for the concept of Caller ID and I am saddened by the reality.

Chris Bulman
Bedford

Recalling the facts
The letter from Chris Bulman certainly had a negative spin on Caller ID as a useful service. Although some of the points he makes are relevant, others are factually incorrect.

The 'number unavailable' message is mostly from older exchanges, although calls from subscribers on the catv network will also be delivered as 'unavailable'. However, when the interface that between BT and the catv network is upgraded, this information should be available. Regarding mobile telephones, subscribers to the 'Orange' network both receive and transmit caller ID information.

Cerisitly, if other network operators want to offer a competitive service, they too will have to offer this.

Chris is correct in that for 'number withheld' to be set permanently on a line or group of lines, a special request has to be made to BT. As for companies requesting suppression of their number, this is mainly for administrative convenience. Companies that operate DDI (direct dial in) exchanges, have individual numbers for each staff member, but the number delivered is that of the main 'switchboard'. If calls are returned to this number, and not to the individual, then the switchboard will be swamped. I assume that other companies withhold their numbers for other reasons. In practice, this problem is not as extensive as Chris suggests. In our experience, about 70% of all calls to our office have number attached, the remaining being split into number withheld, number unavailable, international and payphone calls.

In the area of Call Waiting, BT is well advanced in trials of such a system and should be introducing it within a matter of months. Its protocol has been defined for some time, and is already operational in some states in the US. Supplier's Information Note 242, as referred to in my article, gives technical details.

Introduction of name on a national basis is much more of a problem. Maintenance of a national database is much more of a problem. BT seems to be shying away from.

By the way, existing customers can obtain details on the protocol for exchanging Caller-ID data via the serial port.

Seggy Segaran
York

Coaxial cable tester
Design brief in the March 1996 issue discussed working with avalanche transistors and Nick Wheeler's article covered coaxial cable testing. I have combined elements of the two, constructing within a few minutes this avalanche pulse generator for testing and experimenting with coaxial cables.

PW Fry G45BF
Southampton

Do you have a vtr?
David Markie has designed a comprehensive remote control interface with pc-based control software for vtr. It uses the industry standard, or 'Sony', protocol. Write to the editorial offices (Quadrant House address) for more details.

Non-slewng audio power
In this article, from the March issue, the CSA amplifier of Fig. 1 there is a 200Ω resistor between the 50Ω resistor pair junctions.

Regarding mobile telephones, subscribers to the catv network will be swamped. I assume that other companies withhold their numbers for other reasons. In practice, this problem is not as extensive as Chris suggests. In our experience, about 70% of all calls to our office have number attached, the remaining being split into number withheld, number unavailable, international and payphone calls.

The circuit is sold by Rapid Electronics in kit form, but the company cannot help.

Grahame Collins
Reading College

Schumann Resonance
I wonder if any of your readers could help me with information on Schumann Resonances - particularly in relation to the instrumentation required to detect it - including antenna. The topic is mentioned in EW, Research Notes, May 1993.

Ted Crowley
Inovtron
Co. Wicklow
Ireland

Would Mr Callegari, who wanted to contact people measuring Schumann Resonances, please let us have his address? - Ed

Headphone bass
Why do small 'personal' earphones give such a good bass response? I do not think they are infinite baffle designs, so why is the low frequency cut-off not higher than it apparently seems to be?

David Gibson
Leeds

June 1996 ELECTRONICS WORLD 479
For overseas PAL versions state 5.5 or 6 mHz sound specification.

TV operators. A composite video output is located on the rear panel for reception of TV channels not normally receivable on most television frequencies. TELEBOX MB covers virtually all television frequencies and are fully tested, aligned and shipped to you with a 90 day guarantee and operate from standard voltages and are of standard size and are fully functional. All are IBM-compatible (6317) supported on your PC.

Massive purchases of standard 5.5" and 3.5" drives enable us to present prime product at industry beating low prices. All units (unless stated otherwise) are supplied direct from brand new equipment and are fully tested, aligned and shipped to you with a 90 day guarantee and are of standard voltages and are of standard size. All are IBM-compatible (6317) supported on your PC.

3.5" Panasonic JU4537 20MB IDE I/F £25.95
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3.5" SEAGATE 2E2-331 80 MB FDD £29.95
3.5" Shugart 810 8" SS HH £29.95
3.5" IBM 1480 10 MB FDD £59.95
3.5" Fujitsu M2032R 28MB FDD £29.95
3.5" IBM 1470 12 MB FDD £49.95

The TELEBOX is an attractive fully cased mains powered unit, construction receivers* (TELEBOX MB). Push button controls on the front panel. The TELEBOX can be mounted in a standard 19" rack. TELEBOX MB covers virtually all television frequencies and are fully tested, aligned and shipped to you with a 90 day guarantee and operate from standard voltages and are of standard size.

Lowest price quality. Integrated 120 Mb IDE drive with single 1.44 Mb 3.5" floppy disk drive. With keyboard, 4 Mb of RAM, SVGA monitor output, 256k cache and FULLY TESTED, full 90 day guarantee and operate from standard voltages and are of standard size. All are IBM-compatible (6317) supported on your PC. TELEBOX MB covers virtually all television frequencies. TELEBOX MB covers virtually all television frequencies and are fully tested, aligned and shipped to you with a 90 day guarantee and operate from standard voltages and are of standard size.

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A Class B amplifier requires thermal compensation to avoid thermal runaway, and because bias setting is critical for the minimisation of crossover distortion, which is generally regarded as the most pernicious of non-linearities. The biasing of a Class B output stage requires the establishment of an accurate voltage drop $V_\text{q}$ across emitter resistors $R_e$ of tiny value, by means of hot transistors with varying $V_\text{be}$ drops. It's surprising it works as well as it does.

This thermal compensation has two main problems, attenuation and delay. Since the thermal sensor is more or less remote from the junction whose gyrations in temperature will hopefully be cancelled out, heat losses and thermal resistances cause the temperature change reaching the sensor to be generally too little and too late for complete compensation.

As in Part 1, all the voltages and errors here are for one half of an output stage, using symmetry to reduce the work involved. These 'half amplifiers' are used throughout this piece for consistency, and the error voltages only doubled to represent reality (a complete output stage) when they are compared against the tolerance bands previously quoted.

In this study, we are faced with errors that vary not only in magnitude, but also in their persistence over time; judgement is required as to whether a prolonged small error is better than a large error which quickly fades away.

The **absolute error criterion**
The same issue faces most servomechanisms, and I borrow from Control Theory the concept of an 'Error Criterion' which combines magnitude and time into one number\(^1\text{,}^2\). The most popular criterion is the Integrated Absolute Error (IAE), which is computed by integrating the absolute value of the error over a specified period after giving the system a suitably provocative stimulus; the absolute value prevents positive and negative errors cancelling over time. Another common criterion is the Integrated Square Error (ISE) which solves the polarity problem by squaring the error before integration – this also penalises large errors much more than small ones. It is not immediately obvious which of these is most applicable to bias control and the psychoacoustics of crossover distortion that changes with time, so I have chosen the popular IAE.

One difficulty is that the IAE error criterion for bias voltage tends to accumulate over time, due to the integration process, so any constant bias error quickly comes to dominate the IAE result. In this case, the IAE is little more than a counter intuitive way of stating the constant error, and must be quoted over a specified integration time to mean anything at all. This is why the IAE concept was not introduced in the first part of this article.

Much more useful results are obtained when the IAE is applied to a situation where the error decays to a very small value after the initial transient, and stays there. This is why the IAE concept is not introduced in the first part of this article.
can sometimes be arranged in amplifiers, as I
hope to show. In an ideal system where the
error decayed to zero without overshoot, the
IAE would asymptote to a constant value after
the initial transient. In real life, residual errors
make the IAE vary slightly with time, so for
consistency all the IAE values given here are
for 30s after the step input.

The emitter-follower stage
In the first part of this article, it was shown
that the basic emitter-follower (ef) stage with
the sensor on the main heatsink has significant
thermal attenuation error and therefore under
compensation temperature changes. The $V_q$
error is $+44mV$, the positive sign showing it is
too high. If the sensor is on the TO3 can top
over compensates instead, $V_q$ error equals
$-30mV$.

If an intermediate configuration is contrived
by putting a layer of controlled thermal resis-
tance $80^\circ C/W$, between the TO3 top and the
sensor, then the $50s$ timescale component of
the error can be reduced to near zero. This is
the top error trace in bottom half of Fig. 1.
The lower trace shows the wholly misleading
result if sensor heat losses are neglected in this
configuration.

Despite this medium term accuracy, if the
heat input stimulus remains constant over the
very long term (several kilo-seconds) there
still remains a very slow drift towards over
compensation due to the slow heating of the
main heatsink, Fig. 2. This extra complication
over very long periods was glossed over in
Part I due to sheer lack of space.

This long term drift is a result of the large
thermal inertia of the main heatsink and since
it takes $1500s$ ($25$ minutes) to go from zero
to $-32mV$ is of doubtful relevance to the
timescales of music and signal level changes.
On doubling to $-64mV$, it remains within the
$V_q$ tolerance of $\pm 100mV$. On the shorter $50s$
timescale, the half amplifier error remains
within a $\pm 1mV$ window from $5s$ to $60s$ after
the step input.

For the ef stage, a very long term drift com-
ponent will always exist so long as the output
device junction temperature is kept down by
means of a main heatsink that is essentially a
weighty chunk of finned metal.

The ef system stimulus is a $20W$ step as be-
fore, being roughly worst case for a $100W$
amplifier. Using the $80^\circ C/W$ thermal semi-
insulator described above gives the upper error
trace in Fig. 3, and an IAE of $254mV/s$ after
30s. This is relatively large because of the
extra time delay caused by the combination of
an increased $R_{22}$ with the unchanged sensor
thermal capacity $C_6$. Once more, this figure is
for a half amplifier, as are all IAEs in this arti-
cle.

Up to now I have assumed that the temper-
atures coefficient of a $V_{bias}$-multiplier bias gen-
erator is rigidly fixed at $-2mV/^\circ C$ times the
$V_{bias}$-multiplication factor, which is about $4.5x$
for ef and $2x$ for cfp. The above figures are for
both halves of the output stage, so the half
amplifier value for ef is $-4.5mV/^\circ C$, and for
cfp $-2mV/^\circ C$.

Speeding the response

Staying with ef, if we boldly assume that the
$V_{bias}$ generator can have its thermal coefficient
varied at will, the insulator and its aggravated
time lag can be eliminated. If the $80^\circ C/W$
thermal pad is replaced with standard materi-
al, between the sensor and the TO3 top, the
optimal $V_{bias}$ coefficient for minimum error
over the first $40s$ proves to be $-2.8mV/^\circ C$, which
is notably less than $-4.5$. The resulting
$30s$ IAE is $102mV/s$, more than a two times
improvement. See the lower trace in Fig. 3, for
comparison with the semi-insulator method
described above. In view of the fixed time
constants, dependant upon a certain weight of
metal being required for heat dissipation, it
appears that the only way this performance
could be significantly improved upon might
be to introduce a new kind of output transist-
or with an integral diode that would sense the
actual junction temperature, being built into
the main transistor junction structure.
Although it would be of immense help to
amplifier makers, no one seems to be keen to
do this.

From here on I am going to assume that a
variable temperature coefficient (tempco) bias
generator can be made when required. The
details of how to do it must wait for a later

---

**Fig. 1.** EF behaviour with semi-insulating pad under sensor on TO3 can top. The sensor in the upper temperature plot rises more slowly than the flange, but much faster than the main heatsink or coupler. In lower $V_q$-error section, upper trace is for a $80^\circ C/W$ thermal resistance under the sensor, giving near zero error. Bottom trace shows serious effect of ignoring sensor cooling in TO3-top version.

**Fig. 2.** Over a long timescale, the lower plot shows that the $V_q$ error, although almost zero in Fig. 1, slowly drifts into over compensation as the heatsink temperature (upper plot) reaches asymptote.
article. It should be an extremely useful device, as thermal attenuation can then be countered by increasing the 'thermal gain'. It does not however help with the problem of thermal delay.

In the second ef example above the desired tempco is -2.8mV/°C, while an ef output stage plus V\textsubscript{be}-multiplier has an actual tempco of -4.5mV/°C. In this case we need a bias generator that has a smaller tempco than the standard circuit. The conventional ef with its temp sensor on the relatively cool main heatsink would require a larger tempco than standard.

A potential complication is that amplifiers should also be reasonably immune to changes in ambient temperature, quite apart from changes due to dissipation in the power devices. The standard tempco gives a close approach to this automatically, as the V\textsubscript{be}-multiplication factor is naturally almost the same as the number of junctions being biased. However, this will no longer be true if the tempco is significantly different from standard, so it is necessary to think about a bias generator that has one tempco for power device temperature changes and another for ambient changes. This sounds rather daunting but actually proved fairly simple.

**Complementary feedback-pair output**

As revealed in Part 1, the complementary feedback pair (cfp) output stage has a much smaller bias tolerance of ±10mV for a whole amplifier, and surprisingly long time constants. A standard cfp stage therefore has larger relative errors than the conventional emitter follower (ef) stage with thermal sensor on the main heatsink. This is the opposite of conventional wisdom. Moving the sensor to the top of the TO3 can was shown to improve the ef performance markedly, so we shall attempt an analogous improvement with driver compensation.

The standard cfp thermal compensation arrangements have the sensor mounted on the driver heatsink, so that it senses the heatsink temperature rather than that of the driver itself. See Fig. 4a for mechanical arrangement, and Fig. 5 for thermal model. As in the ef, this gives a constant long term error due to the sustained temperature difference between the driver junction and heatsink mass. See the upper traces in Fig. 7, plotted for different bias tempcos. The cfp stimulus is a 0.5W step, as before. This constant error cannot be properly dealt with by choosing a tempco that gives a bias error passing through zero in the first fifty
Fig. 7. The Vq errors for normal and improved sensor mounting, with various tempcos. The improved method can have its tempco adjusted to give near zero error over this timescale. Not so for the usual method.

Fig. 8. The Vq error and IAE for the improved sensor mounting method on driver back. Error is much smaller, due both to lower thermal attenuation and less delay. Best IAE is 52mV/s; (with gain=0.0038) twice as good as the best EF version.

Fig. 9. Conceptual diagram of the junction estimator. Controlled voltage source E1 acts as an 'analogue computer' performing the scaling and subtraction of the two sensor temperatures V4 and V5, to derive the bias voltage.
then things get interesting. Looking at Fig. 6, it can be seen that the difference between the driver junction temperature and the heatsink is due to $R_1$ and $R_2$. The value of $R_1$ is known, but not the heat flow through it. Neglecting small incidental losses, the temperature drop through $R_1$ is proportional to the drop through $R_2$. Since $C_2$ is much smaller than $C_3$, this should remain reasonably true even if there are large thermal transients. Thus, measuring the difference between $V(2)$ and $V(3)$ gives a reasonable estimate of the difference between $V(1)$ and $V(2)$. When this difference is added to the known $V(2)$, we get a rather good estimation of the inaccessible $V(1)$. This system is shown conceptually in Fig. 9, which gives only the basic method of operation. The details of the real circuitry must wait until we have decided exactly what we want it to do.

We can only measure $V(2)$ and $V(3)$ by applying thermal sensors to them, as in Fig. 4c, so we actually have as data the sensor temperatures $V(4)$ and $V(5)$. These are converted to bias voltage, scaled and subtracted, thus estimating the temperature drop across $R_3$ this is added to $V(5)$ to give the estimate of $V(1)$. The computation is done by Voltage-Controlled-Voltage-Source $E_1$, which in PSpice can have any equation assigned to define its behaviour. Such definable VCVS's are very handy as little 'analogue computers' that do calculations as part of the simulation model. The $V(1)$ result is then multiplied by a scaling factor called estgain which is incor-

porated into the defining equation for $E_1$, and is adjusted to give the minimum error. Note the variable tempco bias approach is used to allow for the difference in resistance between $R_1$ and $R_2$.

The results are shown in Fig. 10, where an estgain of 1.10 gives the minimum IAE of 25mV/s. The transient error falls within a ±1mV window after about 5s. This is a major improvement, at what promises to be little cost.

A junction estimator with dynamics

The remaining problem with the junction estimator scheme is still its relatively slow initial response. Nothing can happen before heat flows through $R_6$ into $C_5$, in Fig. 9. It will take even longer for $C_4$ to respond, due to the inertia of $C_3$, so we must find a way to speed up the dynamics of the junction estimator.

The first obvious possibility is the addition of phase advance to the forward bias com-

Fig. 10. Simulation results for the junction estimator, for various values of estgain. The optimal IAE is halved to 25mV/s; compare with Fig. 8.

Fig. 11. The conceptual circuit of a junction estimator with dynamics. $C$ gives higher gain for fast thermal transients and greatly reduces the effects of delay.

Fig. 12. The initial transient errors for different values of C. Too high a value causes undershoot.
the circuit is thus unchanged, but transients phase advance or lead. The slow behaviour of junction with $R_{50}$ and $R_{51}$ sets the degree of attenuation 100 times by VCVS $E_2$, which is defined to give a gain of 110 times, incorporating an original gain value that gave near zero error decays back over a carefully set time to the gain initially, to get things moving, which is more than twenty times faster than the first improved cfp version, (sensor put on driver) and gives a nicely reduced IAE of 7.3mV/s at 50s. The real life circuitry to do this has not been designed in detail, but presents no obvious difficulties. The result should be the most accurately bias compensated Class B amplifier ever conceived.

In summary

It is, I hope, clear that most of this work is purely theoretical and would be much the better for confirmation by practical tests. This would be a mountain of work and I have no intention of undertaking it. Hopefully it is equally clear that it is no longer necessary to accept $V_{bi} \times \text{multiplier on the heatsink}$ as the only option for the crucial task of $V_{bi}$ compensation. The alternatives presented promise greatly superior compensation accuracy.

The third part of this series will deal with variable tempo bias generators.

References


---

**Telford Electroics**

- **Audio**
  - Variable tempco bias generators.

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**ERNIE**

- **Audio**
  - Variable tempco bias generators.

---

**AUGUE X23E**

- **Audio**
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CIRCLE NO. 117 ON REPLY CARD

CIRCLE NO. 118 ON REPLY CARD
Designing for spectrum analysis

Felice Labbrozzi outlines modules of his high-performance, yet economical, spectrum analyser.

A typical spectrum analyser, one not based on fast Fourier transforms, is essentially a superheterodyne receiver. Within this receiver the local oscillator, a vco, is swept using a sawtooth signal. The oscillator starting frequency equals the intermediate frequency.

With this technique, the received frequency theoretically starts from zero, or dc. In practice, the lower limit is affected by the IF filter response and the input dynamic range. The higher frequency is related to the first IF value.

Design criteria
In order to develop a useful set of specifications for the spectrum analyser, I examined ten commercial spectrum analysers. The specifications I chose are shown in the panel.

Measurements displayed on a spectrum analyser’s vertical axis usually need to be logarithmic. As a result, the input voltage is passed through a logarithmic amplifier.

As with any heterodyne receiver the ability to select the signals received is determined by the intermediate-frequency filter. When tuning rapidly to different signals with a bandwidth near to or lower than the band of the detected signals, there are no simple effects. This is because the time needed by a signal to pass through a filter is inversely proportional to its bandwidth.

As a result, to analyse signals with different characteristics, it is necessary to change three parameters. These are:

- bandwidth – selectivity of the intermediate-frequency filters
- span – the frequency range of the measurement in question
- sweep time – the time needed to analyse the same frequency range.

### Spectrum analyser design goals

- Frequency range dc to 1GHz
- Dynamic display of more than 80dB with linearity of around ±1dB
- Input amplitude range +30, -120dBm minimum
- Frequency accuracy dependent on only one reference oscillator
- Resolution lower than 1kHz
- Ability to print reports, graphs and comparisons
- Capability for storing measurements on dos disks

### Analyser system elements

To accommodate a range of signals from very large to those close to the thermal noise level, a very sensitive front end circuit is needed, preceded by an attenuator. This attenuator must have precise and repeatable characteristics over the whole frequency and amplitude range of the instrument. For this reason I have made it a separate module, Fig. 1.

Input frequencies over a gigahertz could be converted by harmonics from the vco. Displaying all the vco harmonic responses at the same time is likely to cause confusion. In order to separate the spurious responses from the required signals it is better to use a low-pass input filter. This will attenuate those signals higher than 1GHz. If it is necessary to extend the input frequency band, the low pass filter can be replaced by band pass preselectors of 1GHz each. The rf attenuator and the first mixer should be adequate enough to cope with this frequency range if the required amplitude response is to be achieved.

All rf circuits are mounted on one pcb and all control lines are dc. When this module receives input
RF ENGINEERING

Fig. 1. Elements of the 1GHz spectrum analyser, based on the superheterodyne technique. Control and display are handled via a PC, greatly simplifying the system.

Fig. 2. RF-IF module block diagram. Three intermediate frequencies bring the signal down to 10.7MHz.

Fig. 3. This input attenuator is adequate for applications involving frequencies up to 1GHz.

RF and intermediate-frequency

Measurement receiver Figure 2 uses three frequency conversions to bring the input signal to 10.7MHz. This frequency is processed by quartz filters with variable bandwidth and then applied to a logarithmic amplifier such as a TDA1576. After the filters, a low cost logarithmic amplifier such as TDA1576s works well.

When two signals are mixed, the product of the mixed signal has all the characteristics of the original signal, together with those of the local oscillator. To avoid additional noise as each local oscillator increases the original signal, all local oscillators are controlled via phase lock loops. In this way, the local oscillators can be locked to just one stable and low noise oscillator reference derived from the internal frequency reference.

An exception to this is the third local oscillator, which can be continuously tuned over 1MHz when the first VCO is phase locked.

Here, phase noise near the carrier remains low enough to be ignored.

The VCO has to be tuneable over a very large band – typically 1.3 to 2.3GHz. Because of this, the stability and noise performance required to detect signals hundreds of hertz apart cannot be guaranteed.

All RF amplifiers operating under 1GHz are dual-gates MOSFETS – BF 961s, or 40673s for 10.7MHz. These are straightforwardly configured, with gain being controlled via the second gate. The input amplifier and 1.3GHz amplifier comprise inexpensive minicircuits MAR 6s and BFR91s.

System gain is software controlled. Software also compensates for small ripples due to components or their coupling. This makes selection of the first mixer less critical and cheaper.

An auto calibration routine imposes a flat response to the whole system when a signal derived from a harmonic generator is connected to the input.

RF attenuation

If you are not interested in extending the frequency range over 1GHz, you can make the RF attenuator following Fig. 3.

To form a 40dB attenuator, it is better to connect two 20dB sections together. If we use...
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**Fig. 4.** Performance figures for various stages of the spectrum analyser signal processing chain.

**Fig. 5.** Front end synthesis. Theoretically, given a noise figure, NF, of 4.65, a signal just 3dB above noise can be detected.
only one pi network attenuator, stray capacitances limit the frequency response to under 1GHz. Selecting or paralleling standard carbon resistors a 70dB attenuator can be made, increasing the dynamic range of the system. It is necessary to repeat the 20dB section three times using low capacitance reed relays. Excitation coils of the two dual-ganged relays are connected together, as shown.

You must cut the connections as short as possible. The best method is to compute a 50Ω strip line, and insert the relays in the middle of it cutting the strip between the relay’s pins. Relays should have an appropriate pin-out; I used FEME ZFH 002 12 types.

Input capacitors are unpolarised tantalum and ceramic types. If you cannot use surface-mount components, cut the resistor wires down to an absolute minimum and solder the remaining tag directly onto the relay pins.

To compensate for stray capacitances, two short teflon insulating wires can be connected in parallel with the series resistors. Winding the wires together acts as a small capacitor, helping to maintain a flat frequency response.

The following circuit descriptions are not unique solutions. I have tried several different types of circuit, each with good results.

However, if you decide to change the schematics, ensure that the local oscillator rf power is higher than +7dBm. This guarantees acceptable values, imposed in the system design, for losses in the balanced mixers. If the oscillators have a high harmonic content, it is better not to increase the rf power more than 3dB otherwise you could have difficulties reducing spurious responses.

The second conversion, 1300 to 276MHz, was made in a separate silvered-brass box measuring 10 by 50 by 85mm. Inside this box I made resonant coupled circuits using small 6mm brass tubes, teflon standard insulators, small plastic tubes and 3M bolts.

The second mixer uses an OA90 or AA119 germanium diode but I also tested HP5082 series schottky diodes with good dynamic range results.

Designing the rf system
After selecting the conversion frequencies so that the harmonics or mixing products do not block the intermediate-frequency or generate strong spurious frequencies, it is possible to design the amplifier and establish the noise figure of the receiver system.

From the local oscillator and intermediate frequencies you can compute the intermediate -frequency filters so that the attenuation of the images frequencies and the 6/60dB frequency band ratio reaches the desired level. Divide insertion losses so that no filter exceeds 10dB. With a spreadsheet such as Excel you can assign the gain needed for each amplifier so that you arrive at the input level needed for the log amplifier input, Fig. 4.

You can consider the noise figure of a passive mixer as being roughly the same value as its loss. The passive stages can be grouped together in series, making it possible to draw an equivalent block diagram.

Remember that the noise figure of a circuit with a loss of A decibels, connected to the input of an amplifier with noise figure NF decibels, is A+NF decibels. A cascade of two amplifiers with amplifiers N1, G1 and N2, G2 will show looking from the input,

\[ F = F1 + (F2 - 1)G1 \]

or in decibels, a noise figure,

\[ NF = 10 \log F. \]

Now you can draw an equivalent block diagram with three or four stages in cascade. Within this diagram, you can integrate the losses before each amplifier.

To find the system’s noise figure, apply the formula to these equivalent amplifiers for two cascade modules, starting from the last amplifier. Make sure that formulas consider decimals, ie do not mix it with decimals.

The lowest detectable signal depends on the noise present, which is proportional to the bandwidth of the testing system. The minimum noise that cannot be cut, with a bandwidth B of 1kHz, has a power,

\[ N = 10 \log \frac{KTB}{10^3} = -144dBm \]

where K is the Boltzmann’s constant, 1.38x10^-23, and T= 293K, ie room temperature.

Thus theoretically, given a noise figure, NF, of 4.65, as per Fig. 5, a signal just 3dB above noise can be detected. This corresponds to a level of,

\[-144+4.65+3 = -136.65dBm.\]

When the vco generates 1300MHz it is important that the first amplifier, which has to operate on the lowest amplitude signals, does not saturate. To decrease the signal passing through the mixer, a double balanced type is insufficient.

I have made a phase network to produce a 1300MHz signal with opposite phase, and the same amplitude as the mixer output. The re-
onant lines, a quarter wavelength, shown in Fig. 6, were made with teflon RG174 coaxial cable and the trimmers are miniature carbon types, as are the resistors.

**Third local oscillator**

The third local oscillator, Fig. 7, is simple, especially if it is phase locked loop with a very high comparator frequency.

The high comparator frequency guarantees short lock time and spurious signals are far from the carrier minimum of 1MHz. As a result, they are outside the range of the intermediate-frequency filters.

With a direct synthesis phase-locked loop, it is possible to further decrease phase noise and increase stability. This circuit does not limit the noise characteristics of the spectrum analyser provided that you take care with the 9MHz voltage-controlled oscillator.

To make a stable, low noise and non-microphonic oscillator, use silvered-mica NPO capacitors, inductors with mechanically stable supports and a stable, well filtered power supply.

Remember to solder oscillator components to the board, or fix them using wax or adhesive. This will prevent microphonic effects and allow you to remove them if necessary.

**Variable quartz filter**

The quartz filter that determines bandwidth relies on the resistance of PIN diodes in parallel with the tuned circuit. Increasing current into the diodes decreases their equivalent resistance. It also decreases the ratio $Q=F/B$ between the centre frequency and the 3dB bandwidth.

If the variation of diode capacitance is negligible, they do not change $F$ but increase $B$. 
In Fig. 8, trimmer capacitors compensate for asymmetry in the frequency response. The junction fet guarantees a high impedance to help obtain a high Q.

When implementing this circuit, it is necessary to select crystals that work in series fundamental mode, with identical frequencies and with minimum equivalent resistance.

Digital interface
Control voltages are produced by the digital interface. This comprises two MAX500 quad 8bit serially-controlled d-to-a converters and two MAX543 12bit d-to-a converters. These generate voltages for tuning the vco and for sweeping and calibrating it.

Together with low noise op-amps, these d-to-a converters generate voltages to calibrate the vco and the third local oscillator. They also produce signals for tuning the third local oscillator, controlling bandwidth, calibrating gain and changing intermediate-frequency gain.

On the same board ULN2004 drivers control the rf attenuator relays, the cut-off relay and close the vco phase-locked loop in the first local oscillator. A peak detector integrates the amplitude of four successive samples of the log amplifier prior to the a-to-d conversion. Sweeping is produced by a 12bit d-to-a converter, but on screen it is not possible to detect more than 500 horizontal steps. Moreover this integration can be used to increase signal to noise ratio and sensitivity.

Interfacing to a pc
I designed most of the control ICs on to a pc AT BUS expansion card. It holds an Analog Devices' AD7821 fast a-to-d converter, a 12bit d-to-a converter, a triple programmable timer and an 8255 i/o interface.

One flat cable and two miniature coaxial cables transfer every digital control, analogue input to the a-to-d converter and sweep volt-
RF ENGINEERING

triggering a constant pulse generator.

If pulses are shorter than the triggering period, their number per second is proportional to the output frequency. Integrating these pulses, whose amplitude and duration is constant, gives a dc signal proportional to frequency. I use this voltage as feedback in an analogue adder, where tune and sweep voltages and the error voltage from the phase-locked loop comparator are connected.

To obtain low phase noise near the carrier it is better not to divide the output frequency too far. If you do, it is possible to transfer the stability and phase noise of the reference oscillator to the output signal. In my spectrum analyser, this only happens when the vco is locked and the sweep voltage is connected to the local oscillator. Also if the low frequency vco in the third local oscillator is free running, it is relatively clean, and its phase noise is not detectable on the screen with a span of several megahertz.

Phase comparator

The phase comparator is made with a very fast pulse generator and a mixer that samples output frequency, divided by four. Using this solution, pulses are separated from the oscillator output with a cleaner spectrum spread, and amplitude on the phase-comparator can be kept constant.

By connecting a signal derived from the frequency-locked-loop divider to a 16bit divider it is possible to produce a frequency meter. It is possible to count how many pulses are required to bring the counter to zero using the pc. Since the precise counting period is known, derived from the reference oscillator, it is easy to calculate the input frequency with a good degree of accuracy.

The same frequency counting system is used to measure the variable 9MHz third local oscillator. Mixing the measured frequencies in a simple formula it is possible to display the receiving frequency.

Software

The pc software controlling the spectrum analyser comprises these functions:

- draw a graticule and reference numbers for the set parameters on screen,
- draw lines between the y axis samples, repaint colours and clear graticule elements from previous lines,
- read keyboard and optical encoder to determine functions requested,
- save measurement samples in memory for later disk storage,
- configure the pc so that 'Print Screen' key can start a print out.

Some of these functions are time critical because the 'time per division' must be independent from program cycle time. To execute the graphics functions mentioned fast enough for a real-time display, using serial and parallel ports on a standard pc can cause problems.

I tried small 386 assembly language routines and specific professional software, for example National's 'LabWindows' and Hewlett Packard's 'APPCAD'. Both produced good results in terms of speed and graphics.

My main program was written in Quick Basic. It is unpretentious, and is continuously undergoing modifications and updating.

In a subsequent article, Felice discusses set up, calibration and software details.

Further reading


Hewlett Packard, 'Application Note 150 Series'.


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Unipolar positive operation for the DAC-08

The DAC-08 digital-to-analogue converter has excellent performance and provides complementary outputs. Its data sheet shows various schemes for using the outputs with external components to obtain bipolar or unipolar characteristics. There is, however, no apparent method of obtaining a unipolar positive output. The circuit shown for the positive low-impedance equivalent has no provision for complementary output. This circuit is a solution to the problem.

Basic bipolar operation is shown in the data sheet, reproduced in Fig. 1, and a modification of this, Fig. 2, provides the answer. Splitting the two 10kΩ resistors into two halves, with outputs taken from the junctions, effectively shifts the whole output positive by 10V, as indicated in the table.

With this scheme, unipolar positive and bipolar outputs are available from the same circuit.

S Ravindranathan
Naval Physical and Oceanographic Laboratory
Kochi India

Table. Output versus binary input for the unipolar d-to-a

<table>
<thead>
<tr>
<th>Input status</th>
<th>B1</th>
<th>B2</th>
<th>B3</th>
<th>B4</th>
<th>B5</th>
<th>B6</th>
<th>Eo</th>
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Fig. 1. Basic bipolar operation, as shown in Motorola’s data sheet for the DAC-08.

Fig. 2. Modifying the circuit of Fig.1 provides both unipolar positive and bipolar working at the same time.
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Trickle-charged emergency light

Emergency lights designed to switch on when the mains fails have the disadvantages that it can be expensive to repair them when necessary and to keep them in batteries. This design uses an old 12V car battery, but the circuit may be modified to take a 6V battery.

With normal mains operation, the battery charges through $D_1$, the 6V bulb indicating the fact. Since $Tr_1$ is bottomed by way of $D_2$ and the smoothing components $CR_2, Tr_2$ is turned off. Consequently the 12V lamp is off. A mains failure cuts $Tr_1$ off, so $Tr_2$ conducts and the 12V lamp comes on.

Transistor $Tr_2$ is a power type on a heat sink and the 12V bulb may be replaced by three 6V, 8W bulbs for a longer life.

I Rahman
Islamabad
Pakistan

200MHz spectrum analyser displays to -75dB

Used with an ordinary oscilloscope, this circuit forms a spectrum analyser for the 0-200MHz range of frequencies. To simplify examination of the wanted frequency, span and centre frequency controls are arranged to make signals in the middle of the trace stay there as span decreases.

Amplifiers $A_1$ and $A_2$ produce a sawtooth waveform, symmetrical about the 12V rail, which is amplified in $A_3$ with control of amplitude for span and offset for centre frequency. The discharge pulse from $A_2$ for the integrator also goes to the oscilloscope as a timebase trigger.

Driven by the sawtooth, the Mini Circuits POS-400 voltage-controlled oscillator provides a linear voltage/frequency output over the 180-380MHz range and drives the SLB-1 double-balanced mixer.
Electrostimulator

Providing a constant-frequency, variable-amplitude train of pulses in response to a probe coming in contact with a low-resistance biological point, this circuit is also usable as a voltage detector or to measure resistance.

Oscillator IC1a drives the monostable IC1c2b, this part of the circuit only operating when its voltage supply is provided by IC2a. When the passive probe touches a biologically active point of low resistance, the output of IC2a rises and turns on the oscillator, being prevented from affecting the rest of the circuit by D1. Resistor R2 determines the input sensitivity and R1 and R3 are frequency and pulse width adjustments respectively.

To provide an indication of operation, particularly at low frequency, the IC1 oscillator modulates a second oscillator in IC2a,b, which drives the sounder, transistor and led. Component values shown give a 10-150Hz train of 0.5 to 5ms pulses.

Vasily Borodai
Zaporozhje
Ukraine

Symmetry of response is maintained by the CFSK ceramic filters, which also determine the bandwidth. A limitation is the fairly slow response of the output of the NE605. It is usable, but reduces the amplitude of the display at faster sweep rates. It may be that the NE625, which is pin-compatible and faster, would improve matters.

Glyn Roberts
Walsall
West Midlands

Using c-mos ICs, this circuit emits a constant-frequency pulse train to stimulate biologically active points. No switching is needed, since it only operates when the probe touches the point.

Spectrum of local FM broadcast band, 88 to 108MHz, top, and 20MHz square wave, demonstrating sweep linearity, bottom.
This very simply built generator is effectively a set of commercial voltage-controlled oscillators with some power and control circuitry. It provides good performance in a novel arrangement, which can be modified to suit individual needs.

Output is from 50Ω at 5dBm, varying about 1.5dBm over the frequency range, harmonics are at less than –20dBc and phase noise around –100dBc at 10kHz.

Two switched rails carry the tuning voltage, which may be adjusted on each by potentiometers or varied by an externally applied sawtooth for a sweep. A meter, driven by voltage followers, indicates the tuning voltage and therefore the frequency fairly roughly.

A set of seven oscillators from Mini Circuits covers the 25-1025MHz band in roughly 2.1 steps and forms the core of the circuit, each oscillator being followed by an attenuator to define the 50Ω impedance at the BNC outputs rather more exactly than do the oscillators. This also allows individual output levels to be adjusted, since they do vary somewhat. Again, the attenuators come from MiniCircuits.

Since, even if all the oscillators were to be switched into circuit at the same time, current requirement would be only 140mA, the power supply need only be fairly modest, although the tuning voltage must be well filtered to maintain a low noise level.

A BICC-Vero Eurocard 03-2989L suits the circuit well. Note that a ground plane is essential. Björn Nilsson Mijas Spain

Make sure that V+ is as clean as possible. Decouple each tuning port pin with 10nF but do not use an electrolytic capacitor — ed.

Using a set of seven commercial vcos and attenuators, this generator covers a very wide band, tuning being either by ten-turn potentiometers on either of two switched voltages or externally to give a sweep.

Special offer to EW readers

Electronics World readers can obtain a set of oscillators for this design at £60.65 excluding VAT. This represents over 10% discount on the normal price. For each PATx surface-mount attenuator (MATx equivalent) required add £2.95 excluding VAT. Send your order requesting one KPOS2EW set plus any attenuators you want at £2.95 each, together with a cheque or postal order payable to Mini Circuits Europe, Dale House, Wharf Road, Frimley Green, Camberley, Surrey GU16 6LF. Credit card orders are acceptable. Phone 01252 835094, fax 01252 837010.
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CIRCLE NO. 135 ON REPLY CARD
Simulation for mixed signals

Bashir Al-Hashimi explains how mixed-mode circuit simulation packages differ from their solely analogue or digital predecessors.

Computer simulation packages have become an important part of the design and implementation cycle of electronic systems. The result is fewer prototypes, shorter design cycle, and significantly better quality products.

Simulation packages divide into three types: analogue, digital, and mixed signal simulators. Almost all analogue simulators are based on the popular Spice—an acronym for simulation program with integrated circuit emphasis. Spice was developed in the 1970s in the University of California. This type of simulator predicts the frequency and time response of analogue circuits.

The simulation results are graphs of amplitude, phase against frequency in the case of frequency analysis and graphs of amplitude against time for transient analysis. This shows that Spice can be used to simulate the functions of network analysers and oscilloscopes in practice.

A digital simulator enables the designer to perform a timing analysis of a digital circuit. The result is a timing diagram or a truth table of the circuit. In practice, this is similar to using a logic analyser.

As circuits increase in complexity, it is likely that a mix of analogue and digital parts is used for implementation. To predict the performance of such circuits, mixed signal simulators are needed. This article aims to provide an introduction to this type of simulation through a detailed worked example. The simulator PSpice A/D from MicroSim is used for demonstration purposes. For an introduction to circuit simulation and PSpice, read reference 2.

Table 1: Comparison of some commercial mixed-signal simulators.

<table>
<thead>
<tr>
<th>Vendor</th>
<th>Name</th>
<th>Type</th>
<th>Platform</th>
</tr>
</thead>
<tbody>
<tr>
<td>Analogy</td>
<td>Saber</td>
<td>Native</td>
<td>Workstations</td>
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<tr>
<td>Intusoft</td>
<td>ICAP/4</td>
<td>Native</td>
<td>Workstations</td>
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<td>Mentor Graphics</td>
<td>Continuum</td>
<td>Native</td>
<td>Workstations, PCs</td>
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<tr>
<td>Microsim</td>
<td>PSpice A/D</td>
<td>Glued</td>
<td>Workstations, PCs</td>
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<tr>
<td>Viewlogic</td>
<td>Designer</td>
<td>Glued</td>
<td>Workstations</td>
</tr>
</tbody>
</table>

Mixed signal simulator principles

Analogue simulators recognise only analogue nodes where all components connected to the node are analogue. Similarly, digital simulators deal only with digital nodes. A mixed signal simulator on the other hand recognises three types of nodes, namely analogue, digital and interface.

An interface node occurs when a combination of analogue and digital components are connected to it as shown in Fig. 1. Here, node 1 is an analogue node, while nodes 2 and 3 are interface nodes. A mixed signal simulator must translate interface nodes into purely analogue or digital nodes. The translation is achieved using different techniques depending on the simulator. The Saber simulator, for example, uses special models called Hypermodels for the translation.

The PSpice A/D simulator on the other hand, achieves the translation using 1-bit analogue/digital, a-to-d, and digital/analogue, d-to-a interface subcircuits.
Entering circuit details
In order to perform a circuit simulation, the circuit must first be described to the simulator. There are usually two methods of achieving this task, one a netlist, the other schematic capture. The choice of circuit entry depends largely on user preference, with both methods requiring some time to learn. However, it is generally accepted that understanding the basic rules of creating netlists often allow a better appreciation of the simulation process and principles in particular for analogue simulation.

The PSpice A/D simulator supports both types of circuit entry, with the dos version providing the netlist entry, while the windows version provides the schematic capture entry. In this article, the netlist method is used. Note, however, that the simulation of complex digital circuits requires the use of schematic capture. To demonstrate mixed signal simulation, consider the following example.

A mixed simulation example
Figure 3 shows a circuit for producing ramp waveforms. Outputs of the four-bit counter, QD QC QB QA, determine the position of switches S1, S2, S3, ... Logic zero connects the switch to ground, while logic one connects the switch to the inverting input of the amplifier. The amplifier acts as a current-to-voltage converter. Here, the simulator is used to obtain the amplifier output voltage as the four-bit counter goes through its different 16 states.

Listing 1 gives the PSpice netlist of the circuit.

Listing 1

For more information on how PSpice creates and name these nodes, see reference 2.

For example, Fig. 2 shows the circuit of Fig. 1, after PSpice has inserted a-to-d and d-to-a interface subcircuits. Nodes 2 and 3 are now purely analogue nodes. As Fig. 2 shows that the simulator has created two new digital nodes designated 2AtoD and 3DtoA. The PSpice symbol $ represents new nodes occurring as a result of a-to-d and d-to-a interface subcircuits.

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are described using passive component description statements and the voltage reference Vref is described using a dc independent source statement.

Op-amps are modelled in PSpice using X statements. In this example it is assumed that the TL082 op-amp is used with ±15V power supplies. The model of the TL082 device is obtained from the ‘linear.lib’ PSpice library. Switches $S_1, ...$ are voltage controlled and are modelled as follows. First the input, output, control nodes and model name of the switch are described using an S statement.

For example the input and output of the switch $S_1$ are connected to nodes 2 and 0 respectively, while the control is connected to nodes ‘QDBAR’ with respect to ground. PSpice allows you to express nodes as numbers and/or names. For example, the control node of $S_1$ has been given the name ‘QDBAR’, which has been chosen arbitrarily.

Attributes of a switch are defined using a .MODEL statement giving the optional parameters of the switch model. The model parameters are: on and off resistance, and the control voltage for on and off switch state. In this example, all switches have been modelled using the model name ‘SW1’ which has all its parameters set to their model default values.

Now consider the description of the digital parts of the circuit. PSpice A/D has an extensive digital models library including the 74xx series and the 4000 cmos series. The library also has models of data converters and memory devices.

PSpice models digital devices using X statements. For example, the 74HC161 counter is defined by the X2 statement. To be compatible with practical digital devices, PSpice expresses the device nodes using names. The counter output nodes, for example, have been assigned the names QA QB QC QD, as shown in the X2 statement. Note that the parameters $SD_{HI}$ and $SD_{LO}$ in the statement represent a digital one and zero respectively. Finally, the inverters are defined using the X3, X4, X5 and X6 statements.

Simulating input/output

Having described the circuit, the next step is to define the circuit input(s), desired analysis type, and how to display the simulation output results. Clock signals are defined using the U statement as shown in the listing. It has been assumed that the counter clock is 1kHz, or 1ms period. The parameter STIM in the statement is the PSpice symbol for the stimulus generator. Statement (1 1) defines one signal of binary format². Also, the parameters $SD_{DPWR}$ and $SD_{DGND}$ represent digital power supply and ground nodes respectively.

To simulate the output voltage of the amplifier as a function of time, a transient analysis is required. This type of analysis is specified using a .TRAN statement. The netlist shows the transient analysis is performed over a period of 30ms.

The .PROBE command generates a data file for viewing the simulation results graphically. Figure 4 shows the simulated digital and analogue signals of the circuit. The digital signals are the clock and the counter outputs, while the analogue signal is the amplifier output voltage. Note that the digital signals are expressed in terms of logic values (0 and 1), while the analogue signal is described in terms of voltage levels.

A number of interface nodes in the circuit of Fig. 3 occur when the counter outputs QD QC... and the inverter outputs QDBAR... drive the control inputs of the analogue switches $S_1, S_2, ...$ In this case PSpice automatically breaks these nodes into purely analogue nodes using digital/analogue interface subcircuits as mentioned earlier. Figure 5 shows the waveforms of the counter outputs. Note how PSpice changes the output from digital states 0, 1 (top waveforms, Fig. 4) into analogue voltages 0, 5V.

References

Schematic Capture

- Easy to Use Graphical Interface under both DOS and Windows.
- Netlist, Parts List & ERC reports.
- Hierarchical Design.
- Extensive component/model libraries.
- Advanced Property Management.
- Seamless integration with simulation and PCB design.

Simulation

- Non-Linear & Linear Analogue Simulation.
- Event driven Digital Simulation with modelling language.
- Partitioned simulation of large designs with multiple analogue & digital sections.
- Graphs displayed directly on the schematic.

PCB Design

- 32 bit high resolution database.
- Multi-Layer and SMT support.
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Technology. Tel., 01932 254904; fax, 01276 694000; fax, 01276 694800.

The S730 is an 8-bit, serial-output-device, sampling at 3Msamples/s. Both are omos devices with on-chip s/h and provide a reference settling with a sample rate of 2.4Msample/s.

The 7730 is an 8-bit, 8-pin small-outline packages, recently serial analogue-to-digital converters in High-speed a-to-d. SPT7730 and SPT7830 are two 3.3/5V 3Msample/s converters

Please quote “Electronics World” when seeking further information

Feedforward amplifier consists of two for catv use - Motorola's MNN4248 Catv amplifier. For broadband, low -power application at 500MHz is 2dB. Motorola 300ps pulse is 50-200 and fT typically frequencies up to 1GHz. Dc gain on a transistor for low -power application at 5V dram MelCard comes in 4, 8, 16, 32 and makes in values from 22012 to 150k11, although lower values can be supplied to order. The multilayer technique in use provides much less change in resistance than conventional glass-coated types. Thermal time is under 5s and heat dissipation constant 1-2.5mV/°C.

Lens holders. Melles Griot has a new range of bench lens holders for experimental work, which are fitted with easily distinguishable locking knobs and adjuster knobs. Not much more to say, really, except that the holders hold firmly and won’t move until you move them. Melles Griot Ltd., 01223 420071; fax, 01223 425310.

Lasers. A range of visible and infrared lasers is available from UV-Tec in wavelengths of 635, 650, 660 and 670nm, and infrared wavelengths of 780, 830 up to 30mW and 808nm at 5mW. There is also a self-contained lab. version with an emission indicator and power switch. UV-Tec Ltd. Tel., 01252 844880; fax, 01252 844885.

Optical devices
Photo-interrupter. Isocom's IS7302 optical interrupter switch has a 5mm slot and a 0.5mm apature for the light for increased sensing accuracy. Switching time when the beam is interrupted is 3us. A number of case styles is offered and Isocom says it can design styles to suit requirements. Isocom Components Ltd. Tel., 01429 863609; fax, 01429 863581.

Passive components
Chip thermistors. Taiyo Yuden has ntc chip thermistors rated at 63mW in size 0803 and made in values from 220Q to 150kΩ, although lower values can be supplied to order. The multilayer technique in use provides much less change in resistance than conventional glass-coated types. Thermal time is under 5s and heat dissipation constant 1-2.5mW/°C. Taiyo Yuden UK. Tel., 01494 464642; fax, 01494 474743.

Surface-mounted Darlingston. The Zetex FZ7602 Darlingston, an E-line device and SOT223 device, has a minimum hFE figure of 5k at 600mA and 5V, and vice of 100 and 80 and saturation voltage of 0.75V for 0.25A collector current and 2.25mV base current. Zetex plc. Tel., 0161-627 5105; fax, 0161-627 5497.

Surface-mounted electrolysics. Panasonic's VS Series of aluminium, s-m electrolytic capacitors, which are all 5.5mm high, covers the range 25μF-200μF, 4V to 0.1μF-10μF, 50V. The VA series 50V types go up to 47μF and the 6.3V series has values of 1000μF, maximum ripple of 700mA at 120Hz and operating temperature of 85°C. These are water-resistant, Flindt Distribution. Tel., 01530 510333; fax, 01530 510275.

Audio products
Multimedia audio recording. AKM's AK4531 is a multimedia codec for pc-based recording applications, having a two-channel, 16-bit audio codec and a two-channel, 16-bit digital-to-analogue converter, both with an

Passive components

A-to-D and D-to-A converters
High-speed a-to-d, SP7730 and SP7730 are two 3.3V 3Msample/s serial analogue-to-digital converters in 8-pin small-outline packages, recently introduced by Signal Processing Technology. The 7730 is an 8-bit, serial-output-device, sampling at 3Msamples/s. Both are omos devices with on-chip s/h and provide a reference settling with a sample rate of 2.4Msample/s.

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866B dynamic range. It has five-channel stereo and three-channel mono playback mixers. Sampling rates are 4-50kHz for variable audio bandwidth and input can be down to 300Hz. DIP International Ltd., 01223 462244; fax, 01223 467316.

Connectors and cabling

Miniature connectors. Amp Micro-MaTch connectors have a contact spacing of 1.27mm, the range including board and wire connectors to allow a variety of wire-to-board and interboard connections. Versions in the range are somewhat improbable-sounding but, to quote, are “female-on-board, top entry, side entry and surface-mount; male-on-board; male-on-wire, and paddle board connector”. Contact springs are in the board connector rather than, as usual, in the cable connector. In the cable connector is a simple pin with either insulation-displacement section or a kinked solder leg, so separating and optimising the contact force and generation and termination requirements. Gothic Crelton Ltd., 01734 798978; fax, 01734 776095.

Relay sockets. Inelco has a range of DIN sockets and bases for relay mountable on DIN rails, printed boards and chassis and all having finger protection and low flammability material. Examples are a screw base for the two pole 4.7/5mm blade plug-in relays, which can be surface mounted or clipped to a 35mm DIN rail, rated at 240V ac, 10A; and a 14-pin base for 4-20mA miniature relays, timers and controllers rated at 240V ac, 5A. Inelco Ltd., 01734 810799; fax, 01734 810844.

Chip sockets. Robinson Nungest's TSOOP and QFP low-profile surface-mounted sockets are meant for ram, rom and other I/O devices where replacement may be necessary. TSOOP types have 28, 32 or 40 pins at lead spacings of 0.5, 0.55 and 1.27mm, QFP models are in 4x and 8x-100 pin styles on a spacing of 0.8mm. Contacts are rated at 0.1A for the TSOOP and 0.2A; breakdown 500V ac, insulation 1GΩ, contact resistance 60Ω for the TSOOP and 50Ω, Robinson Nungest (Europe) Ltd., 01256 842626; fax, 01256 842673.

Test points. Glass-beaded, board-mounted test points by Williams Hughes consist of a colour-coded bead and phosphor bronze, tinned tag, with 1.3-2.03mm diameter loops for probe connection or to take other components; so raising them above the board surface. The tags are so shaped to avoid damage to through-hole plating and to keep the test points in position when the board is being flow soldered. William Hughes Crelton Ltd., Tel., 01963 363377; fax, 01963 363640.

80-pin styles on a spacing of 0.8mm. Resolution is down to 3mm diameter. Resolution is 1°C to within ±1%, with a response time of 350ms. The instrument is strong enough for factory floor use. Calex Electronics Ltd., Tel., 01525 853800; fax, 01525 853139.

Real-time/storage oscilloscope. VC-6555 by Hitachi is a 100M/sample(s)two channels) and 100MHz bandwidth oscilloscope, with 8Kword capacity for a single channel, delaying sweep and a counter. One-
shot and intermittent events are easily captured and a pre-trigger function shows the leading edge of a triggering waveform. Averaging reduces noise and sweep rate is automatically adjusted for the input frequency. Output is available for hard copy, and, via the RS232 interface, for a pc. Hitachi Denshi (UK) Ltd. Tel., 0181-202 4311; fax, 0181-202 2451.

GPS station clock. Nothing to do with BR, the 88204 GPS Station Clock, available from Steatite, is a time generator whose outputs are synchronised to within 100ns of UTC. It automatically tracks up to six GPS satellites, discriminates its own crystal or rubidium low-noise oscillator and synchronises its outputs to universal time, automatically accounting for leap years and seconds. Several time and frequency outputs are provided, all are RS232 ports and an optional IEEE-488 port. Steatite Insulations Ltd. Tel., 0121-643 6868; fax, 0121-643 2011.

Measuring amplifier. HBM has a small digital amplifier, for the measurement of mechanical quantities, that is for use in laboratory and mobile application. Scout55 operates at 4kHz, being connected by the 6-wire technique to strain gauges, half full and full bridge inductive transducers, piezoresistive and potentiometric transducers and linear variable differential transformers. It has a membrane keyboard protected for hard use, and a ten-digit readout; its RS232C interface allows pc control. HBM United Kingdom Ltd. Tel., 0181-420 7170; fax, 0181-420 7336.

Res. Which is to say, real-time and storage oscilloscope, Hitachi Densha's family of which is now being completed by the introduction of the V2-6645. This instrument is a 100Msamples/s, 25msamples/s (four-channel), 100MHz, 4Kword (single-channel) type with delayed sweep and a 100kHz frequency counter. A standard RS-232 interface allows stored waveform to be transferred to a computer using Hitachi HIKES software and the memory contents are held for up to 10 days. Then power is switched off. Hitachi Denshi (UK) Ltd. Tel., 0181-202 4311; fax, 0181-202 2451.

Low-resistance measurement. Used in conjunction with one's own digital multimeter, the Sutronics M200 enables the measurement of resistance on a most sensitive range of 200ns. Inherent accuracy of the M200 Pass test is typically 0.1%, but final accuracy depends on the dvm in use.

A four-terminal bridge method is used, providing a guaranteed 0.2% accuracy at 20°C for the instrument itself, temperature coefficient being 0.02%/°C from 0°C to 40°C; worst case accuracy is at 25°C. Test current on all but the 202 range is 20mA. Sutronics. Tel./fax, 01929 426400.

Literature

Siemens semiconductors. Technical data on cd from Siemens provides data sheets, s parameters, curves and signal charts, together with the Microelectron training centre, sales information, data for circuit simulation, technical articles and more. The disk is available for dos or Windows and can be installed in networks. There are also files for component comparison and a data retrieval system. BFI IBEKSA Electronics Ltd. Tel., 01622 882467; fax, 01622 882469.

Instrument hire. Livingston Rental, formerly Livingston Hire, has published its first catalogue under the new name. It contains much new equipment and a number of new services in the Instrumentation field. The catalogue is free. Livingston Rental Ltd. Tel., 0181-943 5151; fax, 0181-977 6431.

Data communications. Dataforth Corporation has issued a new catalogue of Industrial Data Communications Products which, in 50pp, provides full specifications, applications and installation data on modems, optical-fibre modulators and rack-mounted modems. Impulse Corporation Ltd. Tel., 01543 466552; fax, 01543 466553.

Thurby Thandar. TTI has a new free catalogue of electronic test gear, which is all developed and made in the UK. New this time is the 1705 a 4.5-digit, dual-display multimeter with an RS-232 interface, true-rms ac measurement, frequency measurement and an optional GPIB interface. There is also a range of counters that handle up to 1.3GHz and a number of instruments for emc testing. Thurby Thandar Instruments Ltd. Tel., 01480 412451; fax, 01480 450409.

Power supplies. From Computer Products, the 1966-7 Power Supply Product Handbook fully describes the company's range of ac supplies, dc-dc converters and dc/ac ring generators, over 1200 in all. Some of them new this time. Also on offer are three free technical guides on thermal management, safety regulations and emc and principles of power conversion. E-mail on jackie.day@cpicc.cpi.sprint.com.

Computer Products, Power Conversion Ltd. Tel., 01494 883113; fax, 01494 883419.

Farnell Components. You can order Farnell's new catalogue and obtain data on the internet (http://www.farnell.co.uk) or, if you suffer from technophobia, the phone from 9am to 6pm. In the 50,000 products, 6000 of them new ones, should satisfy most people's needs, since it includes everything from ordinary passive and active components to network and datacom equipment and top-end instrumentation. Farnell Components Ltd. Tel., 0113 2636311; fax, 0113 2633411.

Printers and controllers

Small print heads. Mitsubishi's W Series 24V dc thermal print heads take the form of a small, flexible printed circuit with a connector, and are intended for incorporation into low-cost mini-printers. They are based on a thick-film process,/cms chips providing shift-register, latch and switching functions. The heads are 5mm high and are in the form of a hard ic moulding, requiring no other cover. Since the drive ic is epoxy coated, print widths are 48mm, 72mm and 96mm in the three types available, with resolution of 800 dots/mm and selectable speed of 3 or 10mss/line. Mitsubishi Electric UK Ltd. Tel., 01707 276100; fax, 01707 278692.

Production equipment

Workstations. Robert Bosch's ESD range of workstations are electrostatically discharged to reduce the number of ics blown by voltages in excess of 100V, which can easily be developed by a worker who shuffles about a bit too much. The workstations are designed to exclude static-dissipative requirements in width and height and are provided with carriers, component containing trays, shelves and a number of other features to reduce not only the shuffle factor but also injuries such as RSI and tendinitis. Robert Bosch Ltd. Tel., 01895 834466; fax, 01895 838548.

Adhesive printing. Loclrite Varnol is a new stencil system which prints different heights of surface-mount adhesive to suit different components in one pass of the squeegee. This is brought about by the use of a flexible stencil providing off-contact printing, which also avoids contaminating the underside of the stencil and therefore cleaning, except when the stencil is changed. Printed dot diameters can be as small as 0.3mm or up to 1.6mm, varying in height from 50microns to 0.6mm. Loclrite UK Ltd. Tel., 01707 821000; fax: 01707 821200.

Transformer kits. For those occasions when no available transformer known to man will do the job, Electrospeed can now offer kits to make your own. Kits consist of a double bobbin section, half of it...
already wound with two independent 11V coils, designed as a pair of twin half shrouts, E and I laminations, and caps and all the other bits, including wire in 12 sizes from 0.2mm to 1.5mm diameter. Four kits cope with ratings of 20, 50, 100 and 200VA. There is a standard turns/v figure for each to give the wanted output voltage and the kits have full instructions and working examples. Electrospeed. Tel., 01703 644555; fax, 01703 610282.

Plug-in switches. Relec's Mascot 9000 series of switched-mode power supplies are cased units plugged straight into a mains outlet and giving 10-40W outputs of 5-24Vdc. Line stabilisation is 0.1% and load regulation below 3% for all models. Output connection is a 2m cord with one of a selection of single or multiheaded jack plugs. Relec Electronics Ltd. Tel., 01962 863141; fax, 01962 855697.

25W and 40W switching psus. XP announces the NAN 25 and NAN 40 low-cost units, available as a pair of twin halves, which use a fixed switching frequency to allow them to meet EN55022 emi regulation below 3%. Both are available for conventional on/off delay timing and, for off delay, no extra power supply is needed. Using either an internal timer or an external pot, timing is adjustable from 0.1 to 100s. The relays are available in 5, 12 or 24Vdc operation. Matsuhashi Automation Controls Ltd. Tel., 01908 231555; fax, 01908 231599.

Switch lens caps. EAO-Highland offers a range of lens caps to make its EAO Series 96 family of cameras, push-button switches look better. The lenses are rounded and come in black(?), red, green, grey, bright orange and blue. There are versions with recesses for one or two lids and for momentary or maintained operation. EAO-Highland Electronics Ltd. Tel., 01444 236000; fax, 01444 236641.

Protection devices
SM transient suppressors. Semiton SM6U series transient voltage suppressors are surface-mounted devices rated at 500W and providing picosecond turn-on and low impedance. With ratings of 5V to 170V, in unidirectional and bidirectional form, peak transient current in avalanche mode is 2.2A to 62.5A for 170V and 5V devices respectively, while in the forward direction, the devices withstand 100A for 1ms. Flint Dynamics Tel., 01530 510333; fax, 01530 510275.

Switches and relays
Time-delay relays. Matsushita offers the TR/TS ultra miniature time-delay relays, which will switch 10-10 to 1kVA on each of four separate contacts arranged in various forms of open and closed. Each TC contact will switch at 250V, although the relays measure only 34 by 34 by 10.8mm. Both are available for conventional on/off delay timing and, for off delay, no extra power supply is needed. Using either an internal timer or an external pot, timing is adjustable from 0.1 to 100s. The relays are available in 5, 12 or 24Vdc operation. Matsuhashi Automation Controls Ltd. Tel., 01908 231555; fax, 01908 231599.

Transducers and sensors
Shaft encoder. Digicots are simple, self-contained shaft encoders, designed for use in xy detection, motor speed monitoring, pick-and-place of components and many other areas. They are non-contacting, panel-mounted types, producing a two-channel, quadrature digital output proportional to rotation at speeds up to 10,000rpm/min. Power needed is under 40mA at 5V. Line counts on the internal decoder are 24-bit/pixel, with a resolution of 0.1 degree. Fairchild Ltd. Tel., 01234 217704; fax, 01234 217083.

Vision systems
Graphics accelerator. Ark Logic offers the ARK-2000M/T-64 board, a graphics and video accelerator, which supports the graphics functions such as BITBLT, line draw, pattern-BLT, colour expansion and raster-ops used by graphics user interfaces. Resolutions are from 640 by 480 to 1600 by 1200 and colour depths of 4:2:bits/pixel, and the board comes with a set of drivers. Amiga Technology Ltd. Tel., 01256 330301; fax, 01256 330302.
Computer board-level products

3U single-card pc. On a single 3U VME bus card, Ovation’s VME 486 PC supports all standard pc i/o functions such as keyboard interface, two serial i/o ports, Centronics/parallel Io interface, floppy and hard disk interfaces and a high-resolution VGA display. An optional module allows up to 16Mbyte of dram in simms, Ethernet connection and dual-port static ram expansion to 1Mbyte. Ovation Systems Ltd. Tel., 01844 279638.

Single-board computer. Apex II from Blue Chip is a single-board computer for industrial and embedded applications, having processor, video and peripherals on one 5V board. It takes 486SX up to 72-pin simm form. There is 1Mbyte of flash memory to give solid-state disk storage, and two further sockets for more of the same. Local bus SVGA video control has 512Kbyte of vram to give 640 by 480 colour lcd. Blue Chip Technology. Tel., 01244 520222; fax, 01244 531043.

Data acquisition

Modular, PC-compatible system. Yokogawa’s DARWIN, a modular family of data acquisition instruments, will compose a complete data-logging system from a range of input measurement and scanning modules, output modules such as computer interfaces and alarms, and hard-copy recording. Each system may be specified as a stand-alone unit with up to 40 channels or as an expandable system with from 10 to 300 channels. The standard software handles data logging and configuration, while an enhanced package is available that provides functions allow for signal processing before measurements are passed to display and alarms. Built-in maths functions allow for signal processing before measurements are passed to the pc. Martron Instruments Ltd. Tel., 01494 459200; fax, 01494 535002.

Development and evaluation

Thermal development. Redpoint Thermalloy has a design kit containing a selection of active and passive heatsinks, materials and accessories, with data to help with the thermal management of microprocessor-based designs. The kit includes the Thermalloy Cooling Module, which is a pin-fin type with a fan, in 5V and 12V versions. It all comes in a carrying case. Redpoint Thermalloy Ltd. Tel., 01793 615366; fax, 01793 615396.

Software

Emc measurement. ESxS-K1 from Rohde & Schwarz is a Windows-based package for emc measurement in small and medium-sized companies. It supports R & S emi test receivers and esases the user’s task by providing dialogues and masks for entering data. Features include single-keystroke measurement, graphic display of test results and scan data, comparison of limit lines with peak search function and average-value detection. Rohde & Schwarz UK Ltd. Tel., 01252 811377; fax, 01252 811447.

High-level language debugger. Kineti Microsystems offers the Flex software which incorporates an integrated macro language and editor for 8-bit, 16-bit and 32-bit applications. It operates independently of hardware providing the information from the embedded system and can be used with in-circuit emulators and background debug mode hardware, rom monitors and simulators for use in the future. It is also independent of file formats. Noral Micrologics Ltd. Tel., 01254 682092; fax, 01254 688647.

Industrial control. From PC Soft International comes Wizcon 5, a Scada package in true 32-bit Windows 95 native code. (Scada is short for ‘supervisory control and data acquisition’.) It is designed for use in both stand-alone applications and in systems for manufacturing and management information, including such applications as control and supervision of processing and manufacture, building automation and public utilities. New features in v.5 are online diagnosis and de-bugging and cluster libraries of animations for many processes. The company offers the package with no extras, all options being included. PC Soft International. Tel., 01233 503838; fax, 01233 513687.
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**Atmel 8051 FLASH Microcontroller Range**

<table>
<thead>
<tr>
<th></th>
<th>8951</th>
<th>8952</th>
<th>1051</th>
<th>2051</th>
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<tr>
<td>FLASH code ROM</td>
<td>4K</td>
<td>8K</td>
<td>1K</td>
<td>2K</td>
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<tr>
<td>RAM</td>
<td>128</td>
<td>256</td>
<td>64</td>
<td>128</td>
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<td>Timer/Counter (16-bit)</td>
<td>2</td>
<td>3</td>
<td>15</td>
<td>15</td>
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<td>Serial Port</td>
<td>YES</td>
<td>YES</td>
<td>NO</td>
<td>YES</td>
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<tr>
<td>Interface</td>
<td>5</td>
<td>5</td>
<td>5</td>
<td>5</td>
</tr>
<tr>
<td>Pins (DIL/PLCC)</td>
<td>40/44</td>
<td>40/44</td>
<td>20</td>
<td>20</td>
</tr>
</tbody>
</table>

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In the second extract from his book *Simplified design of micropower and battery circuits*, John Lenk shows how to apply the MAX712/713 and other ICs in a wide variety of charging circuits.

If you read Part I of this article in the March issue, you should have a good understanding of how the MAX712 and 713 work. These are highly integrated charger ICs specifically designed for use with NiCd and NiMH cells.

Here is a basic simplified design procedure for using these ICs and others in practical circuits. The following steps assume that the maximum power dissipation is not exceeded in your application, that the charge current does not exceed 600mA, and that the battery stack has 11 cells or fewer.

**Check recommendations:** Follow the battery manufacturer’s recommendations for maximum charge currents, as well as charge-termination methods for the specific batteries in an application. Table 1 provides some general guidelines. When the charge rate is greater than 2C, use both temperature and voltage-slope/time. With charge rates below C/2, use the MAX712 for NiMH batteries. Use the MAX713 for all NiCd applications.

**Decide on a charge rate:** A charge rate of C/3 (where C is the battery capacity) charges the battery in about 3h. Current in milliamps required to charge at this rate is calculated as:

\[ I_{\text{FAST}} = \frac{\text{capacity of battery in mAh}}{\text{charge time in hours}} \]

Depending on the battery, charging efficiency can be as low as 80%, so a C/3 fast-charge could take 3h 45min. This has nothing to do with the efficiency of the IC, but reflects the efficiency with which electrical energy is converted to chemical energy within the battery.

**Choose an external dc power source:** An example of this is a wall-cube. Minimum output voltage, including ripple, must be greater than 6V and at least 1V higher than the maximum battery voltage.

**Know the power dissipation:** Calculate the worst-case power dissipation of the power p-n-p transistor \( T_1 \) and diode \( D_1 \) in Fig. 2, “Putting the power back in, Part I” *EW*, March, pp. 223-228) in watts, using:

\[ PD = \frac{(\text{max wall-cube voltage} - \text{min battery voltage}) \times \text{charge current in amps}}{} \]

**Check pin-strap settings:** Consult Tables 2 and 3 to determine the pin-strap settings. For example, with a battery of six cells, both PGM0 and PGM1 (pins 3, 4) should be left open. For a charge time of 90min with the charge-slope limit enabled, PGM2 and PGM3 (pins 9, 10) should be connected to REF (pin 16).

**Table 1. End of charge.**

<table>
<thead>
<tr>
<th>Charge</th>
<th>NiCd cells</th>
<th>NiMH cells</th>
</tr>
</thead>
<tbody>
<tr>
<td>&gt;2C</td>
<td>DV/Do and</td>
<td>DV/Do and/or</td>
</tr>
<tr>
<td></td>
<td>temperature</td>
<td>temperature</td>
</tr>
<tr>
<td>2C to C/2</td>
<td>DV/Do and/or</td>
<td>Max712/3</td>
</tr>
<tr>
<td></td>
<td>temperature</td>
<td>temperature</td>
</tr>
<tr>
<td>&lt;C/2</td>
<td>DV/Do and/or</td>
<td>Max712</td>
</tr>
<tr>
<td></td>
<td>temperature</td>
<td></td>
</tr>
</tbody>
</table>

**Fig. 1. NiCd/NiMH charger with step down regulator (Maxim Battery Management Circuit Collection, 1994, p. 2).**
Simple fast-charger with linear-regulator

Figure 1 shows the MAX713 connected to provide a fast-charge with linear regulation to both NiCd and NiMH batteries.

Note that Fig. 1 is similar to Fig. 2 in Part I. The major difference is that the external pass transistor in Fig. 1 is a MJD2955 and the resistor at pin 15 (V+) is called RSHUNT instead of R1, as in Fig. 2, Part I.

The circuit of Fig. 1 solves two closely related problems found in powering small portable systems – firstly charging the battery and secondly switching over from battery power to ac power when an external ac-to-dc adapter is plugged in. The MAX713 supplies the system load current while the battery is being charged by sensing and dynamically regulating the battery current.

Using a linear regulator instead of a switch-mode circuit is a good approach for small systems: palm-top computers having low-voltage ac-to-dc adapters and low-wattage battery packs with 5V, 9V, and 12V outputs are common examples. The linear regulator approach is also effective for battery back up in non-portable systems, such as large file servers.

The choice between linear and switch-mode is usually a matter of what is an acceptable power dissipation in the regulator pass transistor. For instance, fast-charging three 750mAh NiCd cells from 9Vdc at a 1C rate produces a worst-case dissipation of about 5W. This dissipation is too high for most hand-held applications.

The MAX713 must be programmed for the desired number of cells and charging time. Use the pin straps, as described above, with Tables 2 and 3. The circuit will charge up to 16 cells.

Operating area of the MAX713/MJD2955 circuit, Fig. 2, shows that if the input-output differential can be kept low, the operating current – i.e. charging current plus load current – can be 1A. Input voltage range is battery voltage plus 1V, up to 20V, with a 5V minimum. Supply current when not charging is 5μA maximum. Charging current is limited only by maximum power dissipation PD, and efficiency is battery voltage divide by source voltage multiplied by 100%. For example, if the supply voltage is 6V, used to charge a 3V battery, efficiency is 3/6x100=50%.

Where heat-sinking is not practical, fast-charging large batteries in compact enclosures raises the problem of temperature rise. The battery-charger current source must have adequate efficiency to prevent temperature rise. This usually means some form of switch-mode current source, if the charging plus load currents – i.e. operating current – is greater than about 1A.

Fast-charger with switch-mode current source

The MAX713 can be connected as a switch-mode regulator to provide a fast-charge for both NiCd and NiMH batteries, Fig. 3. The operating area of the switch-mode circuit, Fig. 4, shows that the input voltage range is battery voltage plus 1.5V, up to 20V, with a 7V minimum. Efficiency is 80% with an input of 12V and two cells being charged at 1A. Note that the circuit of Fig. 3 is programmed for two cells, with PGMO strapped to V+, PGMO1 open, and for a time-out of 90min, i.e. PGM2 and PGM3 strapped to RET.

Control loop for the circuit is a variable-frequency time, which senses and regulates current through the battery. Battery current is measured by the 0.08Ω resistor (RSENSE), and this sense signal is compared to an

---

Table 2. Number of cells

<table>
<thead>
<tr>
<th>No of cells</th>
<th>PGM1</th>
<th>PGM0</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>V+</td>
<td>V+</td>
</tr>
<tr>
<td>2</td>
<td>open</td>
<td>V+</td>
</tr>
<tr>
<td>3</td>
<td>REF</td>
<td>V+</td>
</tr>
<tr>
<td>4</td>
<td>BATT-</td>
<td>V+</td>
</tr>
<tr>
<td>5</td>
<td>V+</td>
<td>open</td>
</tr>
<tr>
<td>6</td>
<td>REF</td>
<td>open</td>
</tr>
<tr>
<td>7</td>
<td>open</td>
<td>REF</td>
</tr>
<tr>
<td>8</td>
<td>BATT-</td>
<td>REF</td>
</tr>
<tr>
<td>9</td>
<td>V+</td>
<td>REF</td>
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<td>10</td>
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<tr>
<td>11</td>
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<td>REF</td>
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<tr>
<td>12</td>
<td>BATT-</td>
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<tr>
<td>15</td>
<td>REF</td>
<td>BATT-</td>
</tr>
<tr>
<td>16</td>
<td>BATT-</td>
<td>BATT-</td>
</tr>
</tbody>
</table>

Table 3. Maximum charge time.

<table>
<thead>
<tr>
<th>Timeout (min)</th>
<th>A-to-d τA sampling (seconds)</th>
<th>Slope charge limit</th>
<th>PGM3 pin</th>
<th>PGM2 pin</th>
</tr>
</thead>
<tbody>
<tr>
<td>22</td>
<td>21s</td>
<td>Disabled</td>
<td>V+</td>
<td>open</td>
</tr>
<tr>
<td>22</td>
<td>21s</td>
<td>Enabled</td>
<td>V+</td>
<td>REF</td>
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<td>33</td>
<td>21s</td>
<td>Disabled</td>
<td>V+</td>
<td>V+</td>
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<td>45</td>
<td>21s</td>
<td>Disabled</td>
<td>V+</td>
<td>BATT-</td>
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<tr>
<td>45</td>
<td>21s</td>
<td>Enabled</td>
<td>open</td>
<td>open</td>
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<tr>
<td>45</td>
<td>21s</td>
<td>Enabled</td>
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<td>REF</td>
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<tr>
<td>45</td>
<td>21s</td>
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<td>open</td>
<td>BATT-</td>
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<tr>
<td>66</td>
<td>42s</td>
<td>Disabled</td>
<td>open</td>
<td>V+</td>
</tr>
<tr>
<td>66</td>
<td>42s</td>
<td>Enabled</td>
<td>open</td>
<td>REF</td>
</tr>
<tr>
<td>90</td>
<td>42s</td>
<td>Disabled</td>
<td>open</td>
<td>REF</td>
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<td>90</td>
<td>42s</td>
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<td>168s</td>
<td>Disabled</td>
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</tr>
<tr>
<td>180</td>
<td>168s</td>
<td>Enabled</td>
<td>BATT-</td>
<td>REF</td>
</tr>
</tbody>
</table>

Fig. 2. Operating area of MAX713/MJD2955 5S circuit (Maxim Battery Management Circuit Collection, 1994, p. 2).

Fig. 3. NiCd/NiMH charger with step-down regulator (Maxim Battery Management Circuit Collection, 1994, p. 3).
ANALOGUE DESIGN

internally generated 250mV threshold. The difference is amplified by a gain of eight, and the resultant error signal appears at the current-sense amplifier output on the CC pin.

A second high-gain stage between CC and DRV compares the error signal to the MAX713 +2V reference, switching the MTP120P05 p-channel fet on or off to regulate battery current.

The circuit operates in switch-mode rather than in the linear mode because of hysteresis introduced by the feedback divider and 33pF capacitor connected to CC. The capacitor injects charge into the CC node each time the p-channel fet turns on or off. This raises the error signal slightly above or below the +2V reference, and overdrives the second gain stage to ensure a fast-switching drive signal to the fet.

Current available for charging is 3A. This is determined by the value of $R_{SENSE}$ connected at the BATT- pin ($0.25/3A = 0.0833$).

Lower currents allow smaller external components. For example, for a 1A charger, $R_{SENSE} = 0.25/1A = 0.225$. A 1A Schottky such as the IN5818s can be substituted for the IN5821s, and a Sumida CD75470 surface-mount inductor can be substituted for the Gowanda part. The Sumida inductor has a value of 47µH and is capable of carrying 1A.

Also, higher input voltage can be accommodated by adding a level-shifter between DRV and the fet driver transistors (2N3904/2N3906) and changing the 390Ω shunt resistor value.

**Fig. 4. Operating area of MAX713/step-down circuit (Maxim Battery Management Circuit Collection, 1994, p. 3).**

**Fig. 5. Constant-current battery charger (Linear Technology, Linear Technology, Linear Applications Handbook, 1993, p. AN51-11).**

**Fig. 6. Dual-rate battery charger (Linear Technology, Linear Applications Handbook, 1993, p. AN51-12).**

Constant-current battery charger

A Linear Technology LT1171 can be connected for fly-back operation to provide a constant-current battery charger, Fig. 5. Note that this circuit can be used for either NiCd or NiMH technologies. It does not detect when full battery charge is reached. Nor does the circuit indicate a full charge. However, it does have an important advantage – battery voltage can be higher or lower than the input voltage. This is because of the fly-back configuration.

For example, a 16V battery stack may be charged from a 12V automobile battery. Charge current is sensed by R4 and set at about 600mA. Resistors R5 and R6 limit the peak output voltage when no battery is connected.

Diode D3 prevents the battery from discharging through the divider network when the charger is off. Transistor Tr3 provides for electronic shutdown of the charger (if needed).

Dual-rate battery charger

The LT1171 can also be connected to provide a dual rate battery charger, Fig. 6. Again, the circuit can be used for either NiCd or NiMH cells, but it does not detect when full battery charge is reached. Also, the input voltage must be higher than...
the battery voltage for charging to occur. The primary advantage of this charger is efficiency – at around 90% when charging at maximum output current. Because of this efficiency, no heat sinks are needed on either the LT1171 or the diodes.

A logic signal causes toggling between a high-charge rate, up to 2A, or a trickle rate. An LT1006 amplifier senses the current in the battery and drives the feedback pin of the LT1171. The entire control circuit is bootstrapped to the LT1171 and floats at the switching frequency. This means that stray capacitances must be minimised in component layout.

A transistor sets the gain on the LT1006 by shorting or opening resistor R1. For the values shown, this changes the charge rate between 100mA and 1A.

**Programmable battery charger**

The LT1171 can be connected to provide a programmable charger for use with either NiCd or NiMH batteries, Fig. 7. This circuit does not detect when full battery charge is reached, and the input voltage must be higher than the battery voltage.

But the charging current is directly proportional to the program voltage, which can be controlled digitally via d-to-a converters if desired. A small sense resistor between the bottom side of the battery and its ground senses battery charging current. This is compared with the program voltage. From this comparison, a feedback signal is developed to drive the LT1171 Vc pin, thus controlling the charge current. Typical efficiency during high charge is 90%, thus eliminating the need for heat sinks in most applications.

**Four-cell regulator/charger for NiCd cells**

It is possible to connect the LTC1155 dual power-mosfet as a four cell charger/regulator, Fig. 8. The LTC1155 has the ability to deliver 12V of gate drive to two n-channel power mosfets when powered from a 5V supply with no external components. This ability, coupled with micropower current demands, makes the LTC1155 well suited for high-side...
switching applications. These generally require more expensive p-channel mosfets.

The circuit is well suited for a notebook-computer power-supply system powered by a four-cell nickel-cadmium battery pack. Such a charger consumes very little board space when the LTC1155 and three power mosfets are housed in SO packages. But Tr3 and Tr4 must be provided with proper heat sinks.

One-half of the LTC1155 is the battery-pack charging. The 9V, 2A current-limited wall unit is switched directly into the battery pack through an extremely low-resistance mosfet switch, Tr2. Gate-drive output pin 2 from the 1155 generates about 13V of drive to fully switch both Tr1 and Tr2. Voltage drop across Tr2 is about 170mV at 2A, so it can be surface mounted to save board space.

Inexpensive thermistor RT1 measures the battery temperature and latches the LTC1155 off when the temperature rises to 40°C. This is done through operation of the LT1018 window comparator.

Input to the LT1018 is determined by the RT1 resistance, which, in turn, is controlled by battery temperature. When the temperature rises to 40°C, drain-sense input of the 1155, pin 1, goes low. Use of a window comparator between the RT1 thermistor and the 1155 also ensures that very cold battery packs below 10°C are not quick-charged.

Transistor Tr1 drives an indicator lamp during quick-charge to let the computer user know that the battery pack is being charged. When the battery temperature rises to 40°C, the 1155 latches off, and the battery-charge current flowing through R9 drops to 150mA.

A four-cell NiCd battery pack produces about 6V when fully charged, dropping to about 4.5V when the batteries are nearly discharged.

The second half of the 1155 provides gate-voltage drive, pin 7, for a low-voltage drop mosfet regulator. The LT1431 controls the gate of Tr4 and provides a regulated 5V output when the battery is above 5V. When battery voltage drops below 5V, Tr4 acts as a low-resistance switch between the battery and the regulator output.

A second power mosfet, Tr3, connected between the 9V supply and the regulator output, by-passes the main regulator when the 9V supply is connected. This means that the computer power is taken directly from the ac line while the charger wall unit is connected.

Regulation for both Tr3 and Tr4 is provided by LT1431 which also maintains a constant 5V at the regulator output. Diode string D1-D4 ensures that Tr3 conducts all of the regulator current when the wall unit is plugged in by separating the two gate voltages by about 2V.

Resistor R14 acts as current sensor for the regulator. When voltage drop between the second drain-sense input, pin 8, and the supply, pin 6, rises above 100mV, the regulator latches off at 3A. Resistor R10 with C3 provides a short delay, and the microprocessor can restart the regulator by turning the second input, pin 5, off and then back on.

When the battery voltage drops below 4.6V, the regulator is switched off by the microprocessor. Stand-by current for the 5V 2A regulator is less than 10μA, and the regulator is switched on again when the battery voltage rises during charging.
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Power dissipation within the notebook computer is generally quite low. As a result of fast-charging the battery pack, the current-limited wall unit dissipates most of the power. Transistor \( T_2 \) dissipates less than 500mW, while resistor \( R_6 \) dissipates about 700mW.

Transistor \( T_4 \) dissipates about 2W for a very short period when the batteries are fully charged but dissipates less than 500mW as soon as the battery voltage drops to 5V. The three ICs dissipate little power, though \( T_3 \) can dissipate as much as 7W if the full 2A output current is required and the unit is powered from the wall unit.

**Thermally-based NiCd-cell charger**

Thermocouples can be used to sense cell and ambient temperatures simultaneously. Figure 9, shows such an arrangement and Fig. 10 its charge characteristics. The LT1006 amplifier, \( A_1 \), furnishes the amplification necessary for millivolt-level thermocouple signals.

To understand operation of the circuit, assume that the battery pack is discharged. The battery thermocouple is directly mounted on one of the cells in the pack, while the ambient thermocouple is thermally insulated and mounted on a mass. This mass may be a frame or other part of the equipment. Both thermocouples are at the same temperature, and both produce the same voltage.

Under these conditions, the thermocouples are phase-matched, negative to negative, as shown. Their outputs cancel, and \( A_1 \) sees an input of OV.

The 10kΩ offset-adjustment trimmer is set to introduce enough input-offset so that the \( A_1 \) output swings positive, turning on the transistor. Current flows from the supply, through the battery pack, and to ground via the 250pF shunt. The low-impedance shunt minimises losses, cost, and complexity.

Voltage across the shunt rises to about 625µV, which is the amount of off-set set by the potentiometer. This voltage is fed back to inverting input minus of \( A_1 \) and forms a basic amplifier-servo loop. The loop controls about 2.5A through the battery pack.

Low-resistance shunts can be constructed in a simple, inexpensive way using a small length of wire or a circuit-board trace, Fig. 11. The type and length of wire determines the shunt resistance, which can be altered to produce the desired charging characteristics. Figure 11 also shows the details for wire and circuit-board shunts. In both wire and pcb-track cases, the shunt should have separate Kelvin style connections for sensing so that the high current does not affect the readings.

The battery heats as it charges and this heat is picked up by the battery-mounted thermocouple. Temperature difference between the two thermocouples determines the voltage, appearing at the \( A_1 \) non-inverting input. As battery temperature rises, this small negative voltage becomes larger. Note that a 1°C difference between the thermocouples equals 40µV. Amplifier \( A_1 \) gradually reduces the current through the battery pack to maintain a balance between the inverting and non-inverting inputs.

The effect of this action over time, Fig. 10, is that the battery charges at a high rate until heating occurs, and then the circuit tapers or slopes the charge.

Circuit values given in Fig. 9 limit the battery surface-temperature rise over ambient to about 15°C.

In a thermally based NiCd charger for use with batteries committed to ground, Fig. 12, the transistor is connected as a common emitter, so the inputs to amplifier \( A_1 \) are reversed. However, operation is the same as for the thermally controlled NiCd charger circuit.

Note that in Figs 9 and 12, the trimmer may be eliminated by specifying an LT1006 set at manufacture to the desired offset value. The small shunt-sense voltage, of a few hundred microvolts at most, requires a high-quality ground for accurate results. This ensures that the large current flow through the transistor does not combine with ground-return impedances to create errors. The servo cannot tell the difference between voltages developed across a poor ground and those produced by the shunt.

In practice, all returns should be brought directly back to the supply common terminal.

Figures 9 and 12 both force the transistor to dissipate some power, particularly in the middle of the charge curve, and the transistor is connected as a common emitter, so the inputs to amplifier \( A_1 \) are reversed.
heat may be a problem in very small enclosures. This is typical of many micropower circuits.

A circuit can be designed to eliminate this problem - though requiring a much more elaborate configuration, Fig. 13. This circuit is similar to the other circuits, except that an additional IC, A2, is placed between A1 and the output transistor Tr1.

Op-amp A2 functions as a duty-cycle modulator. Transistor Tr1 - a power fet in this case - operates in the switched mode, delivering duty-cycle modulated-current pulses to the battery pack.

The RC network of R7/C4 filters the switching waveform to dc. Resistors R8 and R9 present a balanced source impedance to A1, and capacitor C2 sets gain roll-off.

The circuit relies on the source impedance of the wall transformer to limit current through Tr1 and the battery pack. This parameter may be set when specifying the transformer.

If the charging source has low impedance, a circuit can be used where the output is essentially a step-down switching regulator, Fig. 14. The 74C04s provide phase inversion, and drive, for Tr1, which is a p-channel mosfet.

Lead-acid battery charging

Lead-acid rechargeable gel cells are attractive because of their high energy-density per-unit volume. They have a long life expectancy when treated properly, but often suffer premature failure because of improper charging.

A charging circuit, Fig. 15, needs precise non-linear temperature compensation, constant-voltage charging with constant-current over-ride, and high efficiency over a wide range of input and battery voltage.

The basic charger is a fly-back design to allow operation with input voltages above or below battery voltage. Switching IC LT1171 operates at 100kHz and can deliver up to 15W into the battery. A dual op-amp is used to control constant voltage and constant current modes.

Acting as a current limiter, A1 turns on when charging-current through R7 exceeds a preset limit determined by R3, R6, and R5. This current limit is included to prevent excess charge current for heavily discharged batteries. Losses in R7 are kept low because the voltage drop across R7 is kept to several hundred millivolts.

Lead-acid batteries have a non-linear negative temperature coefficient, which must be accurately compensated to ensure long battery life and full charge capacity. Resistor R5 is a linear positive-temperature-coefficient thermistor. This characteristic is converted to the required non-linear characteristic by parallel connection with R3. Combination R2, R3, and R4 multiplies the 1.244V feedback level of the LT1171 to the proper 2.35V level required by one cell at 25°C.

One-half of A1 is used as a buffer to drive the resistor network. This allows large resistors, R9 and R10, to be used for the cell-multiplier string. Resistor R9 is set at 200kΩ for each series cell over one. Current through R9 is only 12µA, so R9 can be left permanently connected to the battery. Resistor R10 is added to give the charger a finite output resistance of about 25mΩ per cell in a constant-voltage mode to prevent low-frequency hunting.

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Dataman S4

Compare the Dataman S4 with any other programmer and you'll see why it's the world's undisputed number one.

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Dataman-48

Our new Dataman-48 programmer adds PinSmart® technology to provide true no-adaptor programming right up to 48-pin DIL devices. Dataman-48 connects straight to your PC's parallel port and works great with laptops. Coming complete with an integral world standard PSU, you can take this one-stop programming solution anywhere!

As with S4, you get free software upgrades and technical support for life, so now you don't need to keep paying just to keep programming.

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Try the Dataman S4 or Dataman-48 without obligation for 30 days. If you do not agree that these are the most effective, most useful, most versatile additions you can make to your programming toolbox, we will refund your money in full.

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The current device library contains over 1800 of the most popular logic and memory devices including GALs, PALs, CEPALs, RALs, 8 and 16-bit EPROMs, EEPROMs, PEROMs, FLASH, BOOT-BLOCK, BIPOLAR, MACH, FPGAs, PICs and many other Microcontrollers. We even include a 44-pin universal PLCC adaptor.

If you need to program different packaging styles, we stock adaptors for SOP, TSOP, QFP and SDIP. The Dataman-48 is also capable of emulation when used with memory emulation pods.

Order your Dataman programming solution today via our credit card hotline and receive it tomorrow. For more detailed information on these and other market leading programming products, call now and request your free copy of our new colour brochure.

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