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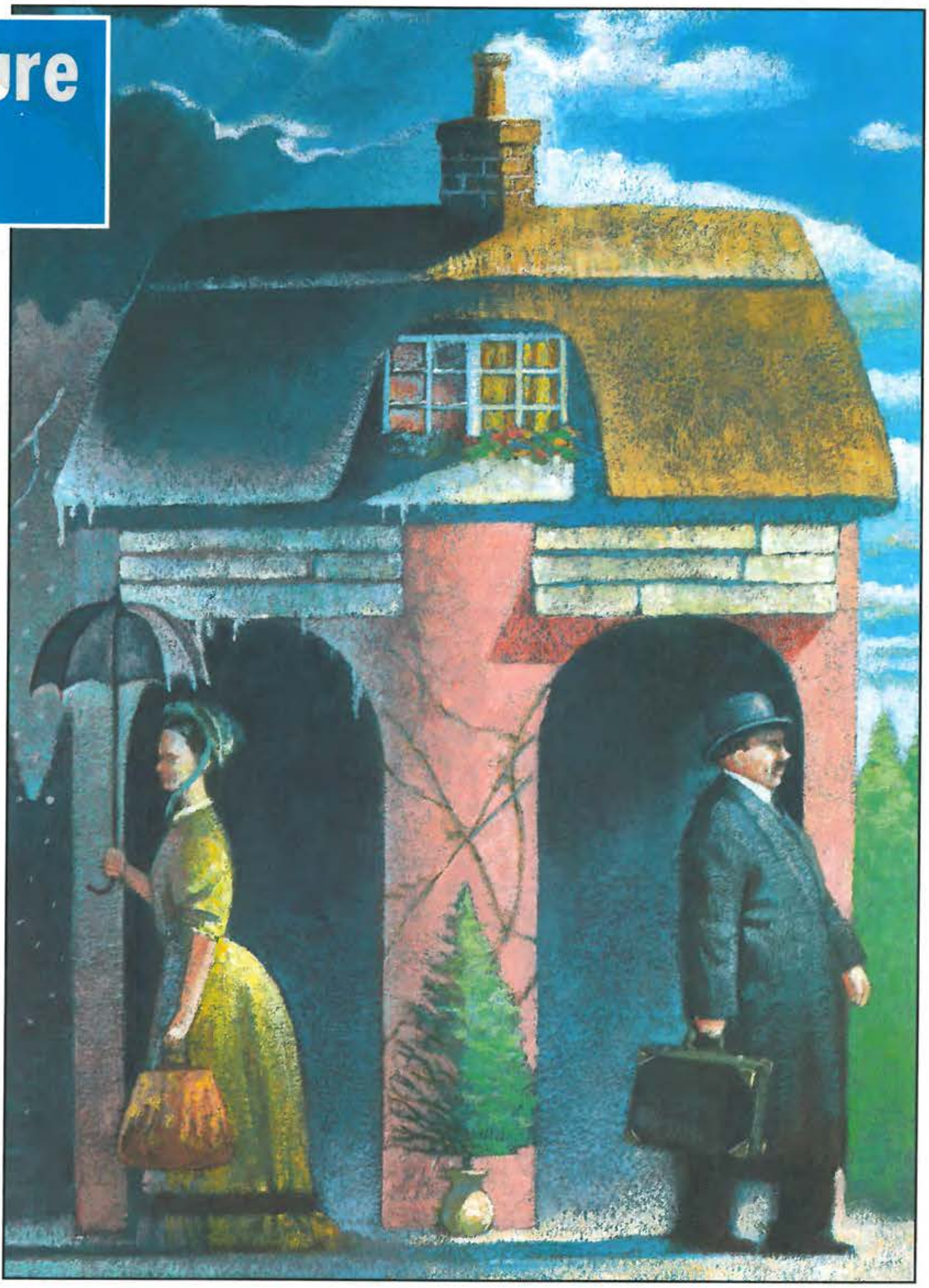

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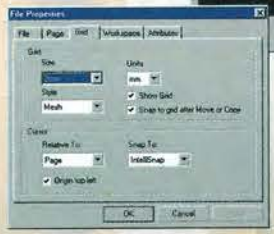

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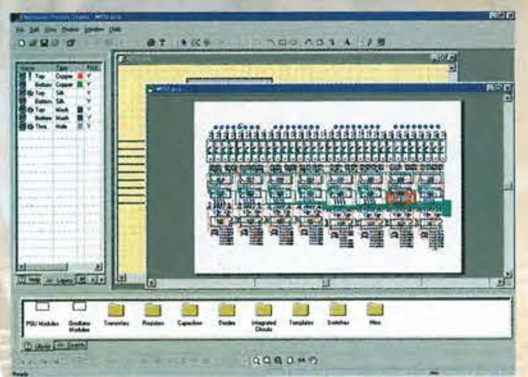

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CIRCLE NO. 101 ON REPLY CARD

CONTENTS

Illustration **Anna-Louise Barclay**

There was a time when it was predicted that OTA technology would become dominant. It hasn't, but it is probably used a lot more than you think. See page 82.

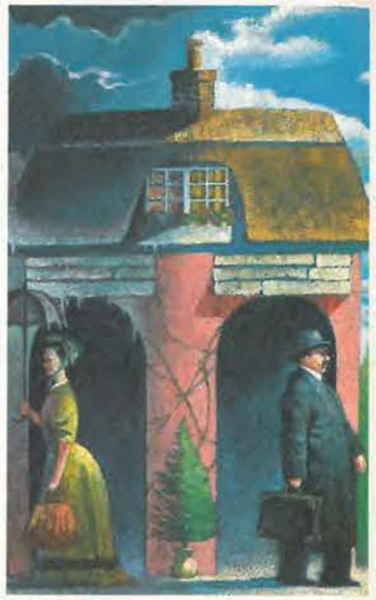

Illustration **Hashim Akib**

This display uses a touch-screen technique that involves sputtering on to glass. Read about it, and a host of other new products page 49.

Small, simple and battery operated, this temperature logger wakes up from time to time to take a reading. When it's full, it connects to a PC for data uploading. See page 26.

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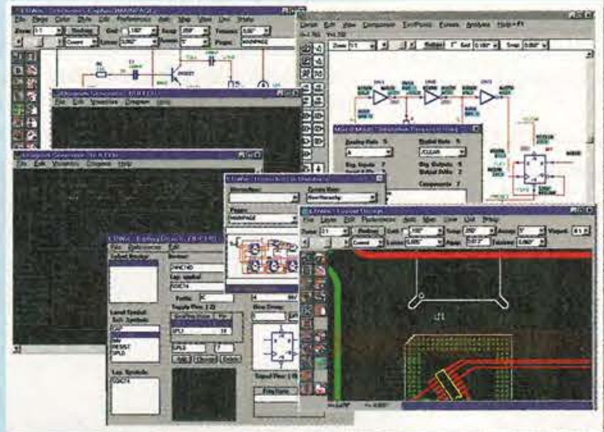

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CIRCLE NO. 105 ON REPLY CARD

Sugar spurns wireless Internet explosion

Wireless Internet access devices will never be a mass market, says Alan Sugar, chairman of Amstrad, in a chilling judgement for proponents of the 'wireless revolution'.

"I'm not saying that it's not a wonderful achievement to be able to send e-mail from the middle of a field," Sugar says, "it's a great scientific achievement – but so what? In practical terms it's no use whatsoever".

Sugar's attitude is in marked contrast to other electronics industry leaders.

Brian Halla, National Semiconductor's CEO, says the industry is "undergoing a shift in product leadership from PCs to wireless communications applications and Internet infrastructure products".

Earlier this month, Dr Tsugio Makimoto, corporate chief technologist of Hitachi told the European Electronics '99 conference that wireless connectivity would spawn a 'Nomadic Age'.

The UMTS Forum reckoned that a big application for 3G (3rd generation) mobile phone networks would be the provision for travellers of "on the pause" access to the Internet and e-mail through wireless links.

Moreover Matsushita, Nokia, Ericsson and Motorola have joined up with Psion to pursue wireless connectivity in portable devices.

Last month Microsoft and BT joined forces to provide services and platforms for wireless Internet access, and 3Com announced a similar initiative with Nokia to add wireless capabilities to the PalmPilot.

Sugar will have none of it. "It will be a niche product for highly technically literate people, but there won't be a mass market for it," he said.

Sugar has a ready explanation for all the hype about wireless Internet access.

"It's fashionable to talk about new technologies among people driven by stock options because then people think you're going to sell hundreds of thousands of the product and that drives your share price up," said Sugar, "the game is to drive your

First bidirectional semiconductor laser

Bell Labs has unveiled what it claims is the world's first bidirectional semiconductor laser.

The laser emits light at two different wavelengths, or colours, depending on the direction of current flow through it. The direction can be changed by switching between negative and positive voltage applied across the device. Normally current can flow in only one direction in semiconductor lasers.

"It's a radically new concept," said Federico Capasso, head of Bell Labs semiconductor physics research department. "This is one laser that behaves as if it were two."

The laser could find use in detecting pollutants in the atmosphere or for increasing the capacity of light-wave communications systems.

The lasers are grown by molecular beam epitaxy, a crystal-growth technology developed at Bell Labs which involves building new materials one atomic layer at a time.

Back and forth... Bell Labs' researchers Deborah Sivco (left) and Claire Gmachl use the bidirectional semiconductor laser.

Europe is failing to address high-tech skills crisis

European companies are failing to tackle the shortage of skilled high-tech workers, according to US market research company IDC.

Senior analyst Andrew Milroy said that the lack of a European strategy for resolving the high-tech skills crisis is making the situation worse. European companies could be hurt by other countries which are addressing the skills shortage. For example, in the US large numbers of foreign workers are being allowed to work for local companies.

The shortage of skilled IT professionals is predicted to grow from five per cent in 1998 to almost 20 per cent in 2002. The shortage is made worse by increasing demand for staff with Internet skills as companies focus on the Internet for a wide variety of management and sales applications.

"Demand for skilled labour will continue to grow year on year," said Milroy. "Soon the demand will significantly outstrip supply, leading to inflated salaries, increasing staff turnover, and therefore higher operating costs and lower profit margins."

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CIRCLE NO. 106 ON REPLY CARD

West Midlands plans high-tech cluster

The West Midlands is the latest region to catch the high-tech bug in a plan to set up a technology cluster of businesses.

Advantage West Midlands, the region's development agency, has put together a plan to inject new life into the area, which it will present to deputy prime minister John Prescott.

Anthony Parish, director general of the Federation of Electronics Industry (FEI), commented: "It is a jolly good idea, there is room for a lot more of this kind of thing. Whether you can create a Cambridge-type phenomenon by spending and government edict, I don't know but you have to try."

However, Parish added that the country still suffered from a lack of entrepreneurs and a capital market to support them.

He said there was also a shortage of people coming out of university with the right type of training. Recently Cadence's design centre in Livingston, Scotland cut back on recruitment plans because of a serious shortage of engineers.

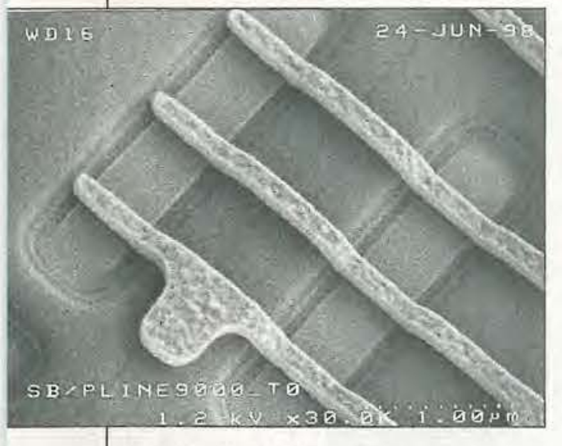
Referring to a joint scheme between BT and Coventry University to promote broadband communications technologies such as ADSL, Advantage West Midlands' chairman Alex Stephenson said: "We have taken the initiative recognising the enormous virtue of an enterprise launched here in the West Midlands, and backing it to the hilt."

"As a result," Stephenson added, "the West Midlands will be at the forefront of developments in ADSL."

Chip transistor size reduction is approaching physical limits

The semiconductor industry's obsession with Moore's Law, doubling transistor count every couple of years, is starting to cause some real headaches for developers of new process technologies. Among these developers is IMEC, Belgium's independent research organisation.

"When we want to scale down technology further, some very severe roadblocks are



Tiny tracks... Getting to 0.13µm features using existing 248nm lithography has resulted in the use of phase shift masks and other complex techniques. IMEC's work on 193nm processing will enable 0.13µm and 0.10µm without these expensive add-ons.

coming up," says Luc Van den hove, v-p of IMEC's silicon process division. "These are scaling optical lithography, a new gate stack, and new back end of line."

New optical lithography means shifting from today's 248nm steppers, being used today for 0.18µm chips, to 193nm lithography.

A better gate stack is needed to avoid breakdown of the gate oxide material as the layer separating the conducting channel from the gate gets thinner.

The new back end of line requires low-k dielectric to reduce capacitance between metal layers and also copper for interconnects.

"These are major challenges," Van den hove warns.

Reducing optical lithography from 248nm to 193nm has received a lot of attention, and IMEC has already started a major programme, working with over 20 firms to develop the steppers, photo-resists and other technologies needed for 193nm processing.

Likewise, work on low-k dielectric and copper interconnect is progressing well.

However, the ace in the pack, with the potential to completely stall the shift to 0.10µm (100nm) and beyond, is the problem of the gate stack – "of crucial importance" describes Van den hove.

IMEC claims gate oxide thicknesses used in a 100nm CMOS process will make devices totally unreliable, unless new materials can be found.

"I'm convinced this is the major roadblock to further development," says Marc Heyns, from the ultra clean processing group.

As silicon manufacturing processes shrink, the gate oxide thickness is also reducing. At 0.13µm, the insulating oxide below the gate is around 2.5nm thick. At 0.10µm this drops to under 2nm - a handful of molecules thick.

While thinning the oxide layer is needed to reduce operating voltage, it offers more chance of defects in the layer forming a conducting path between gate and conducting channel.

According to IMEC's data, oxide breakdown will make 0.10µm devices unreliable because of direct tunnelling of electrons from the conducting channel to the gate.

"As the oxide gets thinner, tunnelling changes from Fowler-Nordheim to direct tunnelling. This dramatically increases leakage and reduces reliability," says Heyns. "Somewhere around the 2.0 to 2.5nm area we're getting into trouble and we need some new materials."

Leakage current through to the gate needs to be reduced by a factor of 100 compared with SiO_2, Heyns believes.

Materials such as tantalum pentoxide have high enough dielectric constant (high-k), but don't work well with conventional CMOS processing.

"These layers need to have a low defect density and accurate thickness control – within one molecule," said Heyns.

To achieve this, IMEC is working with ASM International to develop atomic layer CVD (chemical vapour deposition), which can lay down layers of material a molecule at a time. The team hopes to have a working solution within two years.

However, the problem of breakdown of the insulating oxide layer could also affect devices made using 0.13µm, some of which are starting to appear in pilot lines around the world.

Typical reliability specifications for 0.13µm at 1.5V show that circuits can only operate up to about 90°C. Above this devices become unreliable and might not have the required ten year lifetime.

"It is marginal," admits Herman Maes, v-p of silicon technology at IMEC. "It may be necessary to retrofit some of the high-k results to the 130nm process – not just the 100nm."

Richard Ball *Electronics Weekly*

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CIRCLE NO. 109 ON REPLY CARD

Europe edges into world semiconductor big league

Two European suppliers have made a strong showing in the latest world semiconductor ranking. Infineon Technologies will grow 34 per cent this year to reach No 7 in the world semiconductor league, according to IC Insights, the Arizona market analysts.

ST Microelectronics is also expected to rise fast – a 19 per cent rise this year – thanks to the boom in flash memories. For the first time in its history, ST is expected to crack the $5bn barrier this year, climbing to No.9 spot one place above Philips.

Infineon has benefited hugely from the stiffening price of DRAM. The other big DRAM player in the top ten – Samsung – is forecast to grow 35 per cent this year and is just above Infineon in the league table - at No.6.

Texas Instruments has climbed a slot since last year – from four to three – by selling its analogue and DSP lines into the booming cellular handset market. IC Insights predicts 22 per cent growth for TI this year over last.

Philips grew by just 3 per cent, and Motorola is the only company in the top ten showing zero growth for the year. The US company's second half '99 figures are projected to be 11 per cent down on its first half.

NEC, Toshiba and Hitachi, the three Japanese top ten players, are expected to grow in the middle teens but much of that is because of currency fluctuations. If it were not for exchange rate factors their growth would be in single figures, said IC Insights.

At the beginning of the decade the threesome inhabited the top ten along with Mitsubishi, Fujitsu and Matsushita – a sign of how times have changed for Japan.

World Semiconductor Rankings

1999 Rank		Revenues $bn	99/98 % Change
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2	NEC	8.93	15
3	TI	7.65	22
4	Motorola	7.34	0
5	Toshiba	6.75	16
6	Samsung	5.80	35
7	Infineon	5.24	34
8	Hitachi	5.22	14
9	ST	5.00	19
10	Philips	4.53	3

source: IC insights

Green solder gets green light in Europe

Europe's electronics industry has sent a message to US suppliers that it has opted for tin/silver/copper alloy as a suitable replacement for lead solders.

The lead-free solder decision also has the backing of the UK board makers and contract equipment manufacturers' trade association, the PCIF, as well as solder and materials suppliers.

The International Tin Research Institute (ITRI) made the announcement at the opening of its Soldertec lead-free soldering technology centre.

Lead and other hazardous materials are scheduled to be phased out by the year 2004, under the EU's Waste from Electrical and Electronic Equipment Directive, which is still at the consultative stage. Because its implications are so far reaching, covering so many aspects of waste recovery and recycling, introduction is expected to be pushed back beyond that date.

Free of lead... ITRI maintains that cost is not expected to be a major issue in the introduction of lead-free solder.

Soldertec's director of marketing, Dr Jeremy Pearce said: "We are expecting a large number of enquiries, especially from small to medium size electronics manufacturers."

The centre has also opened an interactive Web site (www.lead-free.org).

Cost is not expected to be a major issue in the introduction of lead-free solder, rather the adoption of different manufacturing practices in the run up to 2001. Because lead-free processes can be degraded by lead contamination from other sources, the changeover has to be comprehensive. Flow-solder baths and re-work stations, etc., are expected to be replaced in the natural course of events.

In a separate move the director of the PCIF, Brian Haken, has taken up with the EC the issue of the automotive industry's temporary exemption for lead used in car batteries. ■

Paul Gregg

On yer bike... A bike hire scheme in the German city of Munich is using Densitron matrix touch screens to allow authorised users to unlock the electric motor assisted cycles. The Call-A-Bike scheme uses 2000 cycles, which are locked in racks next to telephone boxes. To hire a bike the user phones the scheme operator and gives credit card details in exchange for a four digit PIN. The PIN is entered on the touch screen allowing the lock to be released.

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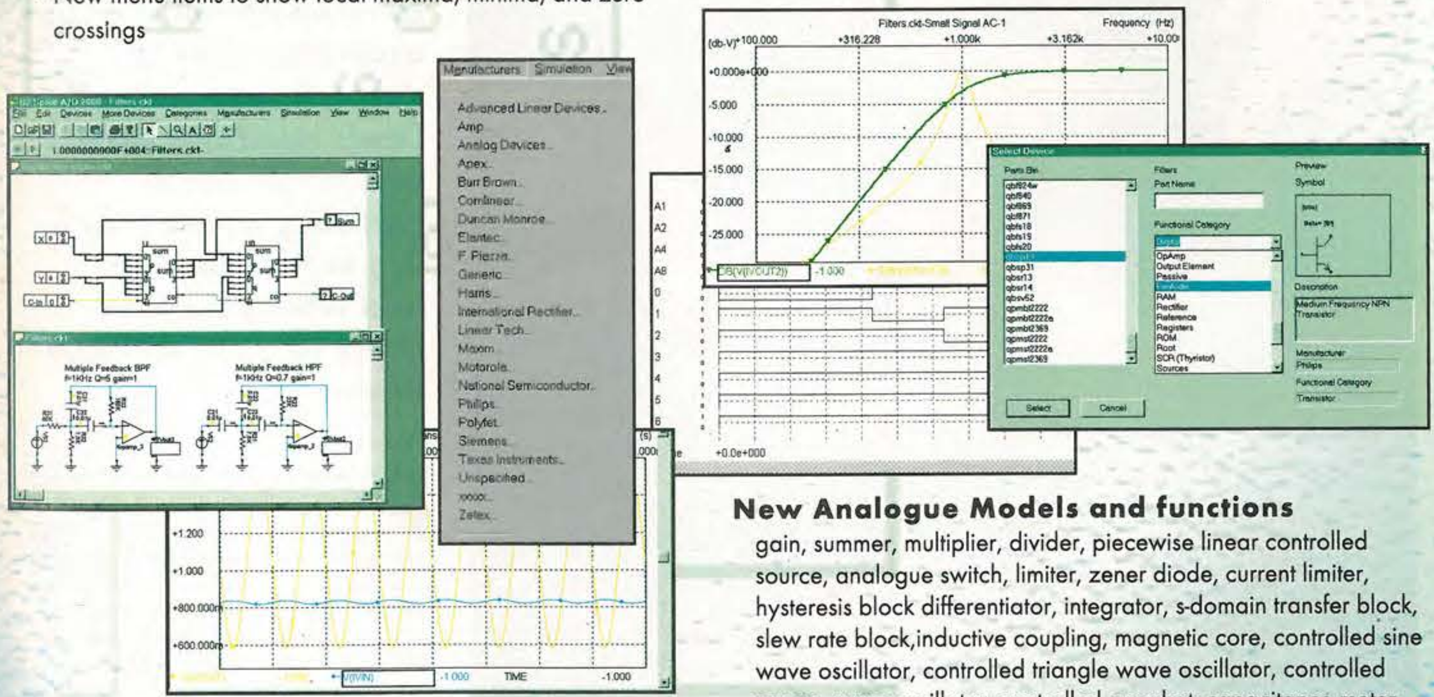

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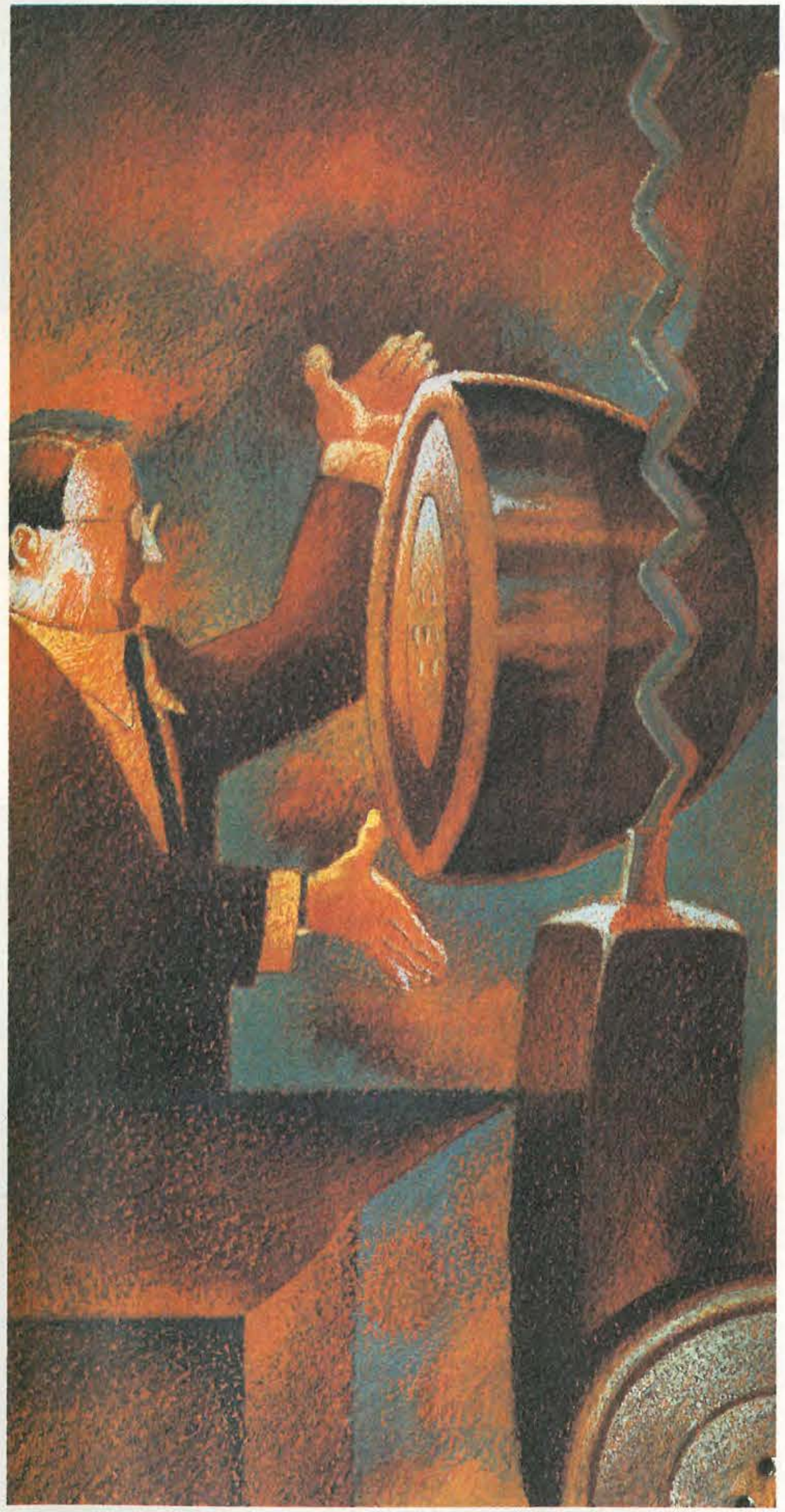

ADSL and beyond

Using the existing telephone network to provide megabit-per-second data rates, ADSL is making high-speed digital communications accessible even to domestic users. Geoff Lewis describes ADSL, and its offshoots, outlining where they are at the moment, and where they are likely to go in the near future.

Proposals for Asymmetrical Digital Subscribers Line, or ADSL, first appeared during 1989. They appeared after it had been recognised that data rates in excess of 10Mbit/s were already being achieved over local area networks using twisted pair cables with similar characteristics to those of the copper cables used for the subscriber's final loop of the plain old telephone system – 'POTS'.

By employing frequency-division multiplex (FDM) using radio frequency carriers up to about 1MHz, it was found possible to add data modulation on to the basic system.

Using FDM, it became possible to provide for a two-way simplex communications system and still provide for the simultaneous use of the normal voice telephone band, including fax, **Fig. 1**. The term 'asymmetrical' is applied to this concept because the two data streams employ different bandwidths and bit rates.

At that time, the concept was seen mainly as a means of delivering video-on-demand (VOD) services via the existing telephone system. The network was expected to act simply as the carrier for third party programme content. Because of the already installed wide-ranging telephone infrastructure, it would be possible to bring the concept on stream fairly quickly.

By comparison, the competing cable TV companies could only provide such a service by using the spectrum wasting process of near-VOD (NVOD), where five or six carrier frequencies are allocated to the same programme but with staggered start times.

During the early 1990s many ADSL trials were carried out around the world. These all proved that the service could deliver VOD with adequate image quality using MPEG compressed television signals over the existing telephone network – and with full VCR type characteristics.

Many other possible services were proposed but the concept failed to influence the broadcasters, the data communicators or the viewing public.

To date, a worldwide ADSL Forum with more than 200 members has been established to drive the system forward. This is now supported by more than fifty ADSL modem manufacturers. Probably the final acceptance of the digital subscriber's line concept depends on its future physical developments.

During the late eighties, the developing convergence between television and computing technology, chiefly in the guise of the Internet, provided a new impetus for ADSL expansion. Baseband modems for connecting to a standard telephone line are now capable of bit rates as high 56kbit/s. This is not far short of the best that that can be provided by the ISDN system, of up to 128kbit/s over the subscriber's final loop.

By using leased lines, the ISDN Basic Rate Access (BRA) can achieve a data rate of 2.048Mbit/s. By comparison, the television system cable modems can operate at up to 30Mbit/s – the typical capacity of a given spur. However, since this bandwidth has to be shared with other users on the spur, the average is seldom more than 2Mbit/s.

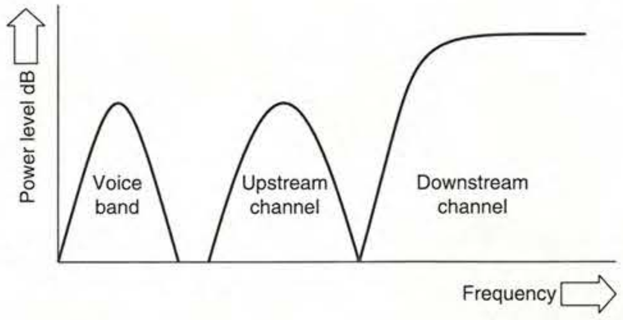

Fig. 1. Asymmetric digital subscribers line channel structure, not to scale. The 'asymmetry' of ADSL is in the bandwidth of the up and downstream channels.

The maximum bit rate for the ADSL system depends chiefly on attenuation produced by the resistance of the local loop, which increases with distance. In addition, the frequency response falls off at about –0.75dB/kHz.

Most importantly for digital signals, the group delay over this section of the line is practically constant thus producing very little pulse distortion and hence very few bit errors. For a reach of about 2.5km the bit rate can be as high as 6.5Mbit/s, falling to about 2.5Mbit/s over 5km.

By 1998 Kingston Communications had installed an ADSL system in the Hull telephone district to provide the first operational public service in Europe. This has a customer base of some 200 000 subscribers.

By mid-2000, British Telecommunications (BT) will have completed the installation of ADSL equipment in 400 digital exchanges to provide a service for up to 6 000 000 domestic and business users. Over the same period, several networks have also been installed in both the USA and Australia.

Up to 1995, most of the standards for ADSL were under the management of the American National Standards Institue (ANSI), with input from the European Telecommunications Standards Institute (ETSI).

During the past few months, the International Telecommunications Union (ITU) has confirmed recommendations for a much wider range of operations. The systems now appear under the global banner of *x*DSL, where *x* covers a range of options which is referred to collectively under ITU Recommendation G.990. The series describes systems that provide subscriber upstream control/demand bit rates as high as 800kbit/s and data downstream at rates of 8Mbit/s.

The xDSL family

The sub-division of the DSL group of systems depends largely on the use of the RF part of the digital spectrum.

Asymmetrical Digital Subscriber's Line (G992.1). ADSL is the original format from which all the others have been derived. The maximum downstream bit rate can be as high as 8Mbit/s, while the subscriber's demand channel can attain 800kbit/s. The local loop reach ranges from 5km at 2Mbit/s up to 8Mbit/s at 1km. The services provide for high speed Internet access, e-mail and video on demand. This system is targeted at the home worker (teleworking), domestic consumers and small businesses.

DSL Lite (G922.2). This represents a simplified version of ADSL whereby the subscribers modem contains only very limited filtering and is designed for user installations. The downstream bit rate can attain 1.5Mbit/s at about 3 to 4.5km reach, with up to 64kbit/s of user bit rate.

High speed Digital Subscriber's Line (G992.1). HDSL1 is designed for business services using two pairs of leased lines to provide E1/T1 data rates in a symmetrical manner, ie, 2Mbit/s in either direction. Repeaters are required for systems with local loop lengths greater than about 5km.

HDSL2 is a variant designed for use in the USA. It is bit-rate adaptive and employs a single pair of leased lines.

ISDN Digital Subscriber's Line. IDSL is a variant of HDSL and is currently under development.

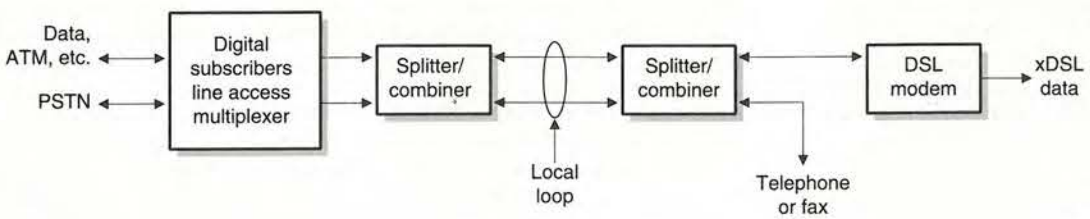

Fig. 2. High-pass/low-pass filter combinations are employed at each end of the local loop of an ADSL network to combine or separate the two services with the minimum of mutual interference.

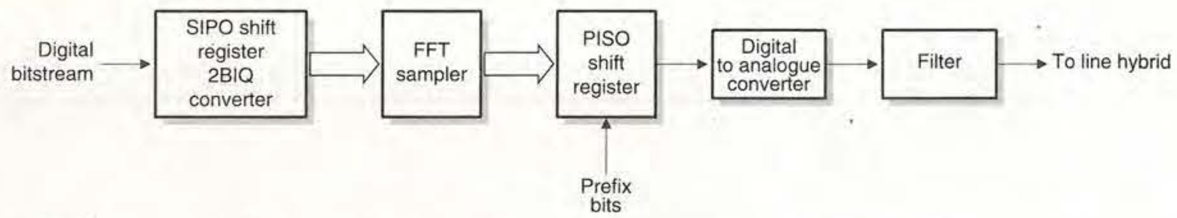

Fig. 3. Basic signal processing sequence of discrete multi-tone modulation – the favourite of three ADSL modulation methods currently under evaluation.

Multi-Rate Symmetrical DSL (G991.1). MR-SDSL is a further concept evolving from HDSL. It currently provides for data rates up to 2.3Mbit/s over a single pair cable. It supports voice and data, together with video-conferencing using MPEG-2 data compression.

Very high speed DSL. VDSL is an asymmetrical system developed for use over short length local loops. It is sometimes seen as a replacement for the 'last mile' or local loop to provide a direct link to optical fibre sections of the network. The downstream bit rates range from 13 to 52Mbit/s, with up to 2.3Mbit/s in the return channel. The downstream local loop reach ranges 1.3km at 13Mbit/s down to 300m at 52Mbit/s.

Aynchronous transfer mode system – ATM

The ATM concept was developed for use in broadband metropolitan area networks (MANs) and optical-fibre systems. Such is its flexibility, it is also capable of being interfaced to SONET (Synchronous Optical NETwork). ATM automatically adjusts the network capacity to meet the system needs and can handle data, voice, video and television signals. These are transferred in a sequence of fixed-length data units called cells.

Common standards definitions are provided for both private and public networks so that ATM systems can be interfaced to either or both.

ATM is therefore a wide-band, low delay, packet-like switching and multiplexing concept that allows flexible use of the transmission bandwidth and is capable of working at data rates as high as 622.08Mbit/s.

Each data packet consists of five bytes of header field plus 48 bytes for user data. The header contains data that identifies the related cell, a logical address that identifies the routeing, forward error control (FEC) bits and bits for priority handling and network management functions. Forward error control applies only to the header as it is assumed that the network medium will not degrade the error rate below an acceptable level.

All the cells of a 'virtual container' follow the same path through the network that was determined during call set-up. There are no fixed time slots in the system so that any user can access the transmission medium whenever an empty cell is available. ATM is capable of operating at bit rates of 155.52 and 622.08Mbit/s. The cell stream is continuous and without gaps.

The position of the cells associated with a particular virtual container is random and depends upon the activity of the network. Cells produced by different streams to the ATM multiplex are stored in queues awaiting cell assignment.

Since a call is only accepted when the necessary bandwidth is available, there is a probability of queue overflow. Cell loss due to this forms one ATM impairment. However, this can be minimized by the use of statistical multiplexers. Bit errors in the header which are beyond the FEC capability can lead to mis-routeing.

Typical ADSL applications

The loop-length limited downstream bit rate can be employed in a time-division multiplex (TDM) mode. For example, by using MPEG-2 digital compression with suitable channel coding, video signals of various grades can transmitted direct to the subscriber's home.

At bit rates of 2Mbit/s, the downstream channel can provide either three video signals of VHS quality together with digitally coded stereo sound, or with MPEG-2 compression. A rate of 6Mbit/s can provide video with a quality equal to that of a broadcast transmission. At 8Mbit/s even high-definition television images are possible.

As an example of the ISDN compatibility of xDSL, the subscriber's upstream channel can operate at either 16kbit/s or 64kbit/s. The downstream can be divided into three times 2.048 Mbit/s, giving 6.144Mbit/s or four times 1.544 Mbit/s, resulting in 6.176 Mbit/s, to meet either the European or North American telecommunications standards (E1/T1 rates).

Current ADSL modems are now equipped with IEEE 1395 (FireWire) and USB (Universal Serial Bus) interfaces so that the system can readily be extended to operate throughout the home. Used in this way, ADSL enables corporate LANs to be extended right into the home worker's domain.

Basic organization of xDSL

As indicated in **Fig. 2**, high-pass/low-pass filter combinations are employed at each end of the local loop of an ADSL network in order to combine or separate the two services with the minimum of mutual interference.

For ADSL-Lite systems, the local terminal filtering is minimal to simplify installation for the subscriber. This reduced bit rate system can suffer from interference and noise from the telephone network, often influenced by the number of items of equipment coupled to the subscriber's network socket. Under difficult conditions, it is possible to fit small low pass filters inside the terminating socket.

At the exchange end of the system, banks of splitter/combiners and modems are installed at the main distribution frame together with the Digital Subscriber's Line Access Multiplexer (DSLAM).

To minimise the effects of both radiated and induced radio frequency interference, ADSL employs a balanced feed of RF signals to the local loop.

Data-packet structure

ADSL has a packetised mode of transmission with the data contained within a frame structure, either in High level Data Link Control (HDLC) or Asynchronous Transfer Mode format. There's more on ATM in the panel entitled 'Aynchronous transfer mode system – ATM'. A typical frame consists of two header bytes and up to 1600 data bytes, delimited by two flag bytes.

Since ATM is used extensively for transporting digital signals within the television studio and production areas, it is logical that it should be used for similar video demands in DSL systems.

While five 'adaption' layers are defined for ATM, each with the prefix AAL, only AAL1 and AAL5 are employed for television distribution. Layer AAL1 is defined for constant bit rate (CBR) production distribution, while AAL5 is defined for consumer quality VOD variable bit rate (VBR) systems. Each MPEG-2 transport stream (TS) packet consists of 204 bytes while each ATM cell contains only 55 bytes. There's more on this in the separate panel.

Thus for VOD on DSL using AAL5, two MPEG-2 transport-stream packets of 408 bytes are loaded into eight ATM cells, with the spare 32 bytes being used for cell sequence addressing and bit error control in the form of a checksum. By comparison, for the production environment, each single MPEG-2 TS packet is loaded into four ATM cells, again with the spare bytes occupying a control and addressing function.

Modulation methods

To provide a suitable transmission format, three modulation techniques are being evaluated. These are Quadrature Amplitude Modulation (QAM), Carrier Amplitude and Phase Modulation (CAP) and Discrete Multi-Tone modulation (DMT). Only DMT is expected to survive the extensive trials.

Discrete multi-tone modulation has much in common with COFDM, or coded orthogonal frequency-division multiplex, as used for terrestrial digital television. The frequency spectrum between 20kHz and about 1.125MHz is divided into nominally 256 carriers spaced by 4.3kHz. In order to minimise interference, not all of these are used. In the time domain, each sub-carrier represents a unique symbol period.

In order to maximise the channel capacity a form of bit-rate reduction coding known as 2BIQ is employed. With this, each pair of binary input digits is converted into 1 of 4 quaternary symbols.

Next, the new bit stream is used to modulate the individual carriers during the symbol period so that the spectrum takes on the appearance of a parallel transmission bus. The channel capacity and signal to noise ratio can now approach the Shannon limiting values.

The number of bits transmitted in a given sub-channel can be varied dynamically depending on the signal-to-noise ratio on that sub-channel. This not only improves the signal quality on a particular line, but also minimises the effect of any cross-talk from other lines.

In the interests of high quality of service, the system employs forward error control in the form of Reed-Soloman coding with the option for adding interleaving to improve the burst error capability.

Figure 3 shows the basic signal processing in a DMT modulator. Each serial digital input signal is first encoded into parallel format and then passed through a fast Fourier-transform processor to convert the frequency-domain signals into time domain values with a sliding time-window effect. These values are then transcoded into serial format and digital to analogue converted before transmission.

As indicated in Fig. 3, a number of prefix bits are added to the bit stream in order to minimize interblock interference. When these bits are dropped in the decoder, the bit stream is broken up to improve the FFT time window definition.

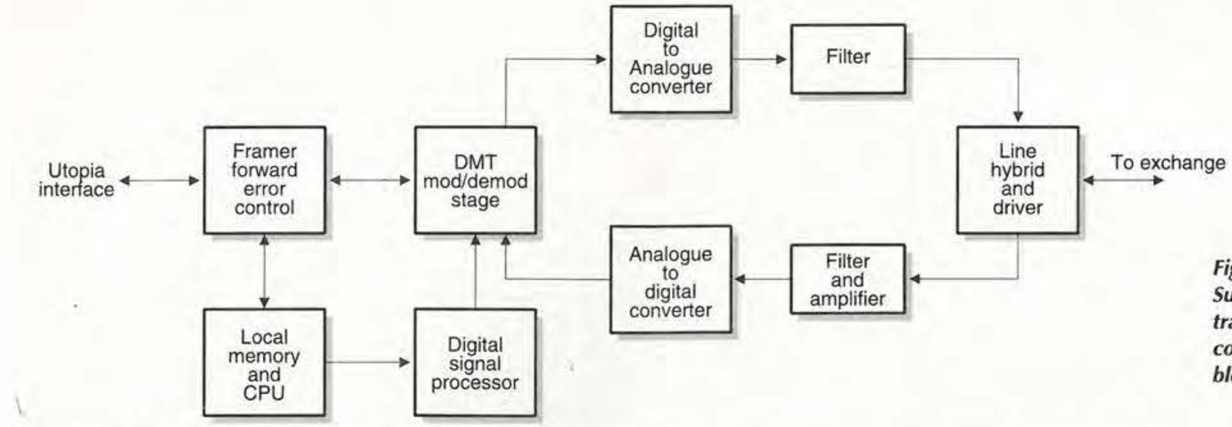

Fig. 4. Subscriber's ADSL transceiver and control function blocks.

Subscriber's transceiver and modem

Figure 4 shows the important operational areas of the subscriber's terminal for both transmit and receive modes. The hybrid transformer provides a balanced feed between the modem and line and helps to minimize cross-talk.

However, the splitter in the form of filters and amplifiers has to separate signals of a few tens of volts on the voice line from RF signals of the few tens of millivolts of the DSL signal. Loop-disconnect and dialling pulses can be particularly troublesome. The high-pass filter section has to contain a dc block to isolate the ringing tones from the DSL section and to prevent a DSL fault from short circuiting the normal POTS system.

Digitally-adaptive equalisers are set under feedback control to suit individual line conditions and characteristics. As well as continuous interference, there are temperature and moisture-dependent variable parameters to contend with. Provision is made so that at each new acquisition, the line amplifier and equaliser gains can be reset under the control of a short bit pattern used as a training sequence to establish the maximum quality of service.

The frequency and power characteristics for each sub-carrier are continually adjusted in both transmit and receive paths, through the use of a feedback mechanism. This monitors the near end and far end cross talk (NEXT and FEXT) in order to minimize inter-carrier interference.

Line characteristics for each narrow sub-channel are practically constant so that the minimum of pulse smearing is created thus improving signal quality. Any impulsive noise that is input as interference is spread over many sub-channels by the FFT processor window so that this form of interference is less likely to create data errors.

RF signals transmitted on the standard telephone network are essentially analogue. As a result, analogue-to-digital and digital-to-analogue converters are incorporated in each modem.

Typically, each modem is based on a single highly integrated chip with embedded microprocessor (CPU) and digital signal processors (DSP). The CPU is commonly based on the reduced instruction set (RISC) design, such as the ARM devices. These not only manage the bit framing and error control, but also perform the modulation and demodulation under software control.

The UTOPIA – or Universal Test and Operations Physical Interface for ATM – has been developed specifically for such applications.

The future of DSL

The last stages of telecommunications liberalisation are likely to occur within the next year or two. This is described as 'unbundling the local loop' to allow third party access to the line subscriber's socket. This is likely to have considerable implications for DSL systems because any such third party will expect to have access to a wide bandwidth in order to make a financial success from providing any new services.

With such third-party access to the network, interference and cross talk between the different service providers will introduce the need for accurate monitoring of power levels and spectrum management. Could this be the driver needed to ensure that optical fibre is introduced into the local loop as a matter of urgency? If so, then the digital subscriber's line will have at last be seen to have arrived. ■

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3. www.adsl.com www.adslforum.com

The UK's congested roads are not only annoying, they can be lethal. Just listen to Radio Five Live's traffic update in the mornings. Help is at hand though, as the big car manufacturers like Jaguar install car crash avoidance systems. Pete Mitchell takes the wheel.

No more bumps?

If you believe the adverts on tv, motorists spend all their time zooming along deserted lanes or across deserts, or parked on cliff tops, never another vehicle in sight. Despite these fantasies, car makers are beginning to consider designing-in electronic aids intended to cope with the reality of UK motoring today, whether on urban streets or motorways: lines of hundreds of vehicles, chugging nose-to-tail, only a few metres apart.

Collisions are a perpetual risk in these conditions, and there has long been talk of adapting the aerospace industry's range-finding techniques into cheap instruments that can give drivers early warning of a dangerous situation – or even apply the brakes automatically.

Governments in a Europe of increasingly dense traffic flows are very keen on this idea. For one thing, minimising collisions will help keep things moving – accidents are one of the principal causes of delay on trunk roads. And range-finders will also enable cars to employ 'intelligent cruise control', microsystems that can adjust the cruising speed of the vehicle to the maximum safe level depending on the speed and distance of cars in front and behind. That would smooth out traffic flow, and reduce the 'bunching' delays regularly seen on heavily used motorways like the M1.

However, car manufacturers have not been convinced that anti-collision gadgetry will give them a competitive edge that is worth the R&D cost, including the risk of market failure. To pay for itself, an "extra" has to bring in more sales by influencing the consumer's choice of vehicle. But having a collision detector won't help you much if the driver behind you doesn't have one, and doesn't manage to brake in time. To be useful, they have to become ubiquitous – and so they need to be cheap to make and based on reliable, standardised components.

Jaguar is to use adaptive cruise control with collision avoidance on its XKR sports car, shown above, starting from next year. The system functions either as a conventional cruise control or, in traffic, maintains a safe distance from the car in front. Developed by Delphi, the system uses a 76GHz microwave radar mounted in the nose of the car. It calculates speeds of vehicles up to 150m ahead. As vehicles in front slow down, the Jaguar's brakes are automatically applied, and audible warnings are given to the driver if the driver in front does something really stupid.

So the European Commission is encouraging manufacturers to get together and agree the best technologies for collision avoidance, rather than develop rival proprietary solutions that could fragment the market and delay its introduction.

The automotive industry is often very slow in transferring technology from academic to applied research and then to the market, according to Wolfgang Gessner, of the EC's Innovation Relay Centre in Germany. "The process is often less linear than that in the car industry," he says. So IRC has established a conference, Advanced Microsystems for Automotive Applications, to promote collaboration on this and other ideas.

This year's AMAA, in Berlin, saw demos of more than 30 different types of microsystem, including one of the first laser collision avoidance devices to be offered as a commercial product. Developed by the Swiss Centre for Electronics and Microtechnology (CSEM) in partnership with Fiat, the OLMO (On-vehicle Laser Microsystem for Obstacle detection) was built into a Fiat Brava for the AMAA demonstration.

There are various options for range-finding sensors, but the OLMO developers opted for laser radar (LIDAR) with an ordinary 1550nm semiconductor laser. This choice means that the device can be not only cheap, but also with a very low power requirement – only 10mW. It is also compact and robust enough to be packaged in a simple injection-moulded unit for mounting anywhere in the car. For the demo, it was built into the left headlamp of the Fiat Brava.

Like all radar, the sensor emits electromagnetic pulses, detects their reflections and calculates the time of flight, making up to 20 measurements a second. The pulse beam is scanned across a 20° field of view ahead of the car, using a swivelling mirror rotating on a special frictionless pivot.

But there's a problem peculiar to automotive range-finding: if there are many cars in the neighbourhood using a similar device, there is a serious danger of crosstalk between the scattering of beams from units on different cars. So the device must also ensure that it detects the right reflection.

To do this, the OLMO group borrowed a signal processing technique from spread-spectrum communications systems like GPS. The instrument modulates each laser pulse with a unique binary code, generated randomly in real time. The back-scattered signal is then cross-correlated with the code sent, selectively killing any crosstalk present in the back-scatter.

This initial 'acquisition' step gives a coarse measurement of the distance to the scattering surface. The instrument then uses this knowledge to perform a much more accurate 'tracking' measurement – only 10cm for distances less than 30m.

These techniques give the device a range of 80m in good visibility, according to CSEM. In poor visibility, the range depends on conditions – it is not as good as microwave radar, but is about 40 per cent better than the human eye, claims the team leader Max Monti.

In fact, Fiat didn't want more than this, says Monti: "They have done simulations that show it's safest that drivers don't have a collision warning system that is much more sensitive than their own eyes, otherwise they start to go too fast."

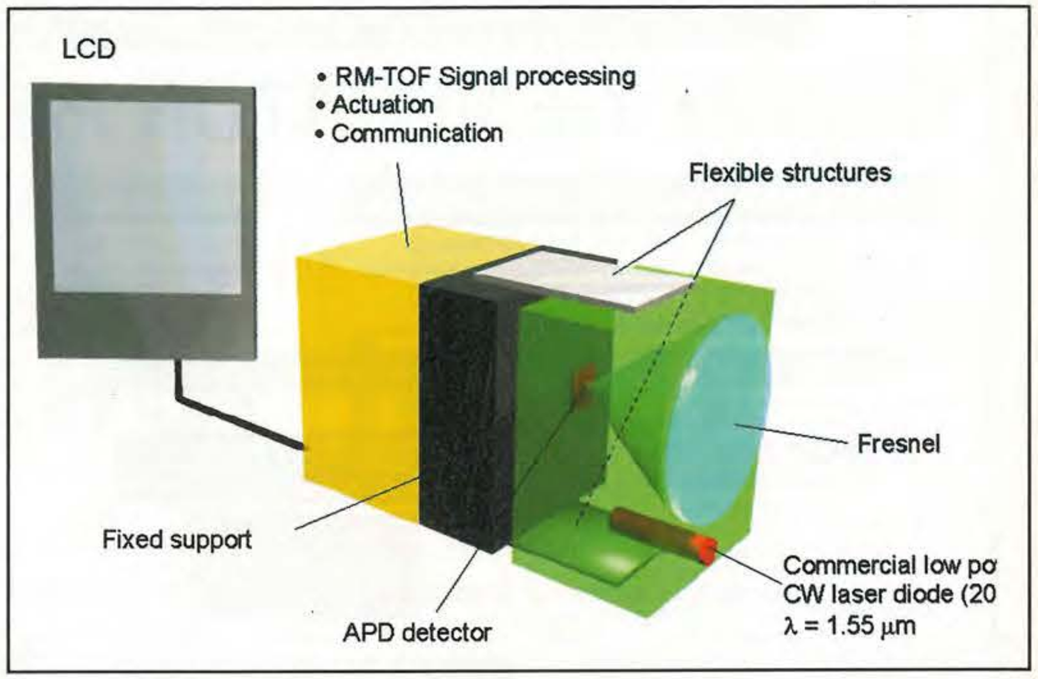
Microwave radar has been the favoured choice so far because of the ready availability of components. Mercedes' top-of-the-range cars already have offer a radar collision-avoidance option, made by ADC. But, according to Monti, it has one fatal drawback: electromagnetic pollution. "Imagine what will happen if in ten years' time every car on the road is emitting microwave radiation", he says, pointing to the current worries about the health effects of these emissions from mobile phone base stations. Whether right or wrong, these fears are real, and are likely to prompt manufacturers to move to optical collision avoidance systems, he says.

However, OLMO has the typical limitations of pioneering work. Its lifetime, at only six months without maintenance, is rather short. It is also too expensive, with production cost estimated at around £250: the usual chicken-and-egg problem of high unit costs at low volumes before economies of scale are established.

So will the manufacturers bite at such sophisticated 'extras', as they did with anti-lock braking? Only if they can see a competitive need: within each class of vehicle, there are fewer and fewer functional differences between the leading car brands, says Jaguar technical specialist Paul Mulvanney. "They all have four wheels and all start and stop reliably. Digital technologies create opportunities to incorporate entirely new features into the basic design".

A lot depends on the manufacturer's public image. Fiat, which is evidently committed to the collision-avoidance idea, may be planning to use it to push a safety-first theme. Monti says that three car manufacturers are interested in his LIDAR, and believes the whole industry will be offering collision-avoidance systems within two years.

Other electronic assists, such as active suspension using multi-axis silicon accelerometers, or piezoelectrically driven fuel injection jets, may appeal more to the technical departments of manufacturers obsessed with what Mr Jeremy Clarkson calls, "sheer raw performance". But it seems unlikely that car makers who encourage drivers to take blind corners at 60mile/h, as Alfa Romeo did in a recent newspaper advertisement, will have collision avoidance instruments high on their priorities list. ■



Collision course... The on-vehicle laser microsystem for obstacle detection – OLMO – was developed by the Swiss Centre for Electronics and Microtechnology in partnership with Fiat. It was first demonstrated in a Fiat Brava. The range finder uses a laser radar with a 1550nm source. Range finding is good for 80m, with accuracy of 10cm when closer than 30m.

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Tanδ capacitor tester

Cyril Bateman has been enhancing his tanδ meter with better protection against charged capacitors. He's also produced an efficient battery regulator providing ±5V from four AA cells.

Recently I measured every board-mounted aluminium electrolytic capacitor in a piece of test equipment. In just ten minutes, using my tanδ meter, I identified 31 capacitors with a tanδ in excess of 0.4. Capacitors measuring such a high tanδ are worn out and should be replaced.[1]

Aluminium electrolytic capacitors are quite different from any other capacitor type in their construction and failure modes. Usually, the capacitance value of a worn out aluminium electrolytic capacitor is little changed, but its series resistance and tanδ will be seriously increased. These failure modes together with the appropriate in-circuit diagnosis methods, were discussed in the May 1999 issue of *Electronics World*.[2]

Design and development of a new and unique prototype in-circuit capacitor tanδ test meter was described in the June issue together with the schematic drawing of the measurement circuit used.

This self-contained prototype meter included a low output impedance 100Hz generator, a floating 9V supply for the PM128 display meter and a stable ±5 volt battery power supply. However, to save space, these additional circuits were not included in the June article.[3]

Since the original design was published, I have made various enhancements to it. The most notable of these is the addition of a further INA118 instrument amplifier, in front of the logic channel LM311 current sensing comparator. Amplification of the tiny

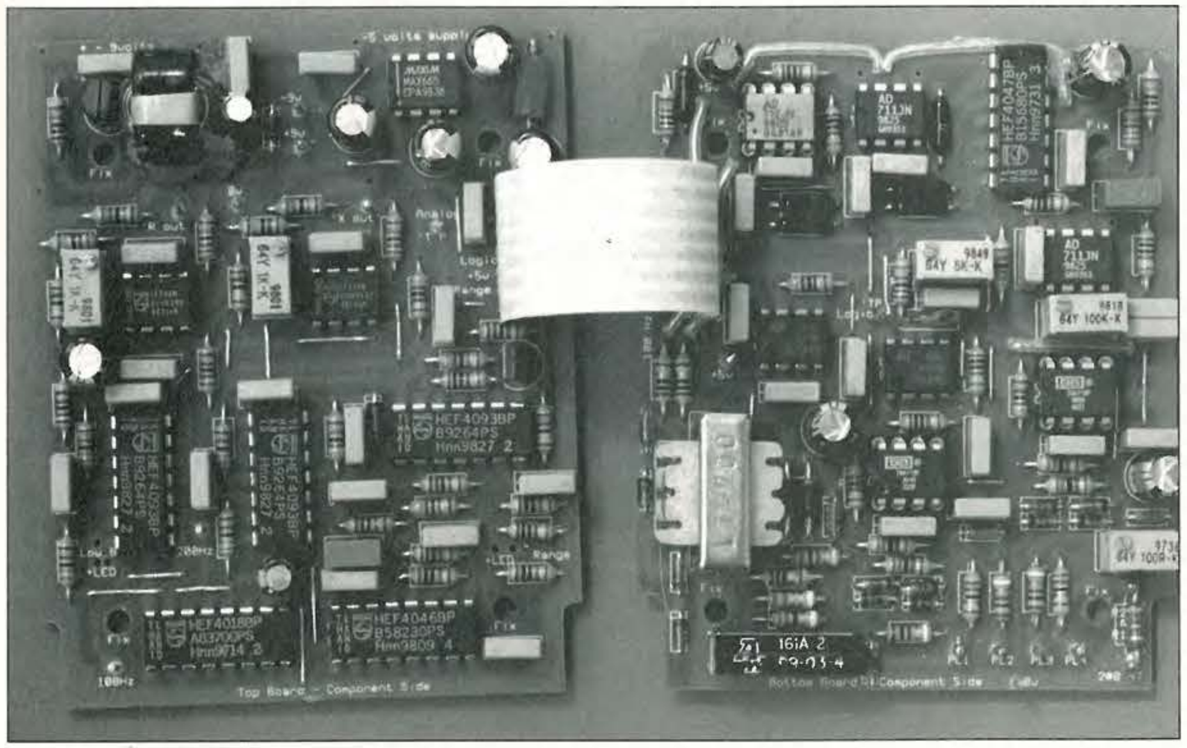


Fig. 1. Complete Tanδ meter on two interconnected, single-sided boards. This view shows the final design, production board. The right hand or 'bottom' board contains all the circuitry of Fig. 2a). The left hand or 'Top' board contains all the circuitry of the Fig. 2b).

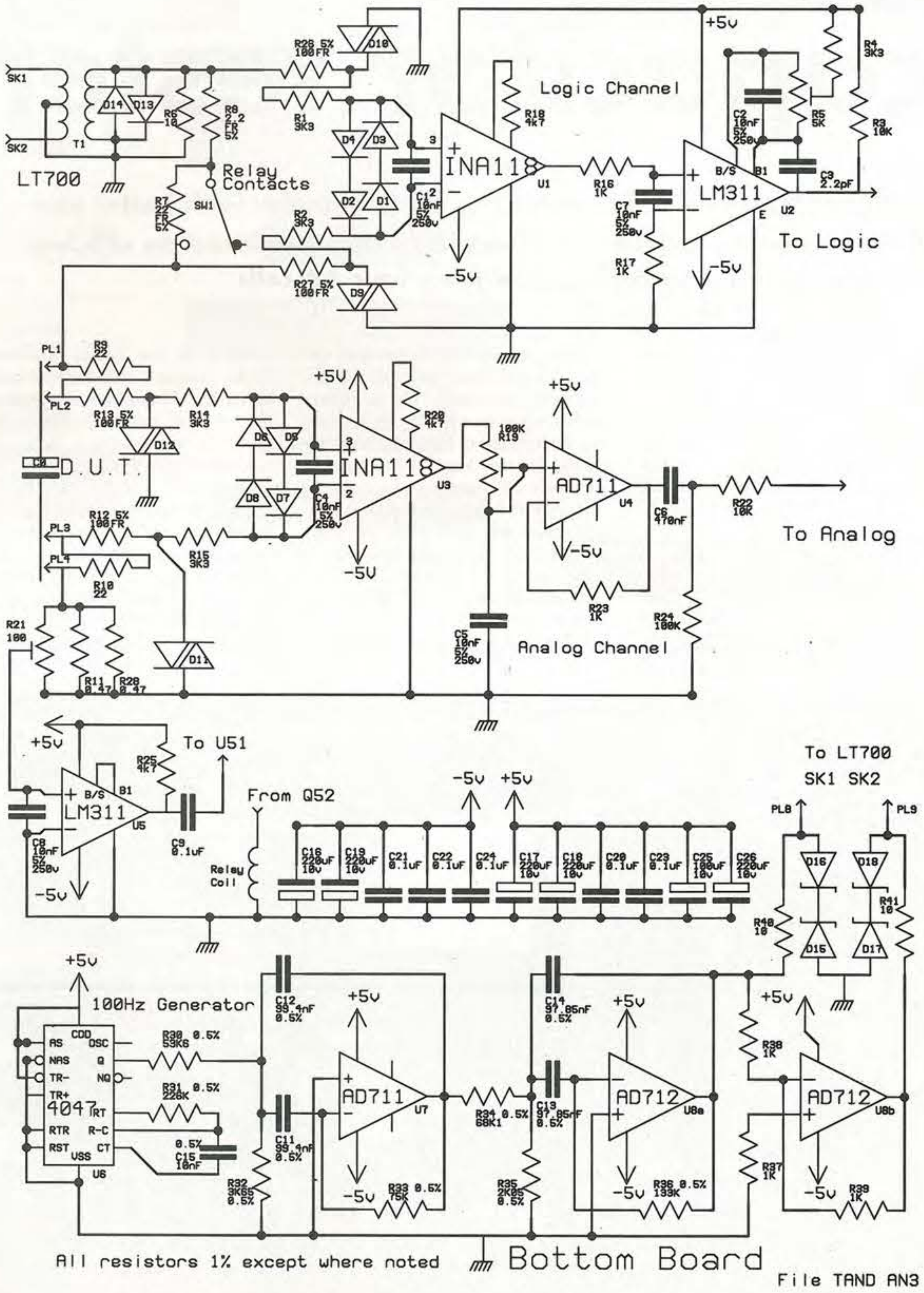
voltage dropped across the current sensing resistors, used to identify the phase of the capacitor's current waveform, improves measurement accuracy particularly when measuring smaller value capacitors.

I chose a standard OKW V155 'Shell' case, measuring 158 by 95 by 45 mm. This uncut case has an externally-accessible battery box taking four AA cells. Smaller cases did not provide enough space for the measurement circuitry, batteries and PM128 display.

The prototype's 140 by 74 mm printed board had to be divided into two sub-boards approximately 85 by 70mm, in order to fit into this case, **Fig. 1**. Both boards are simply interconnected using a seven-way, flat flexstrip jumper cable. In the picture, this cable hides four of the 220μF decoupling capacitors, two on each board.

To complement the above cosmetic improvements, I decided to enhance tanδ meter's 'charged capacitor' protection. In the event these apparently simple changes proved extremely time consuming.

Fig. 2a) Complete schematic, including all supply line decoupling, for the right hand or 'Bottom' board, shown in Fig. 1. Fig. 2b) is the complete schematic, including the floating 9V supply for the PM128 display, the –5V converter and all supply line decoupling, for the left hand or 'top' board in Fig. 1.



Enhanced charged-capacitor protection

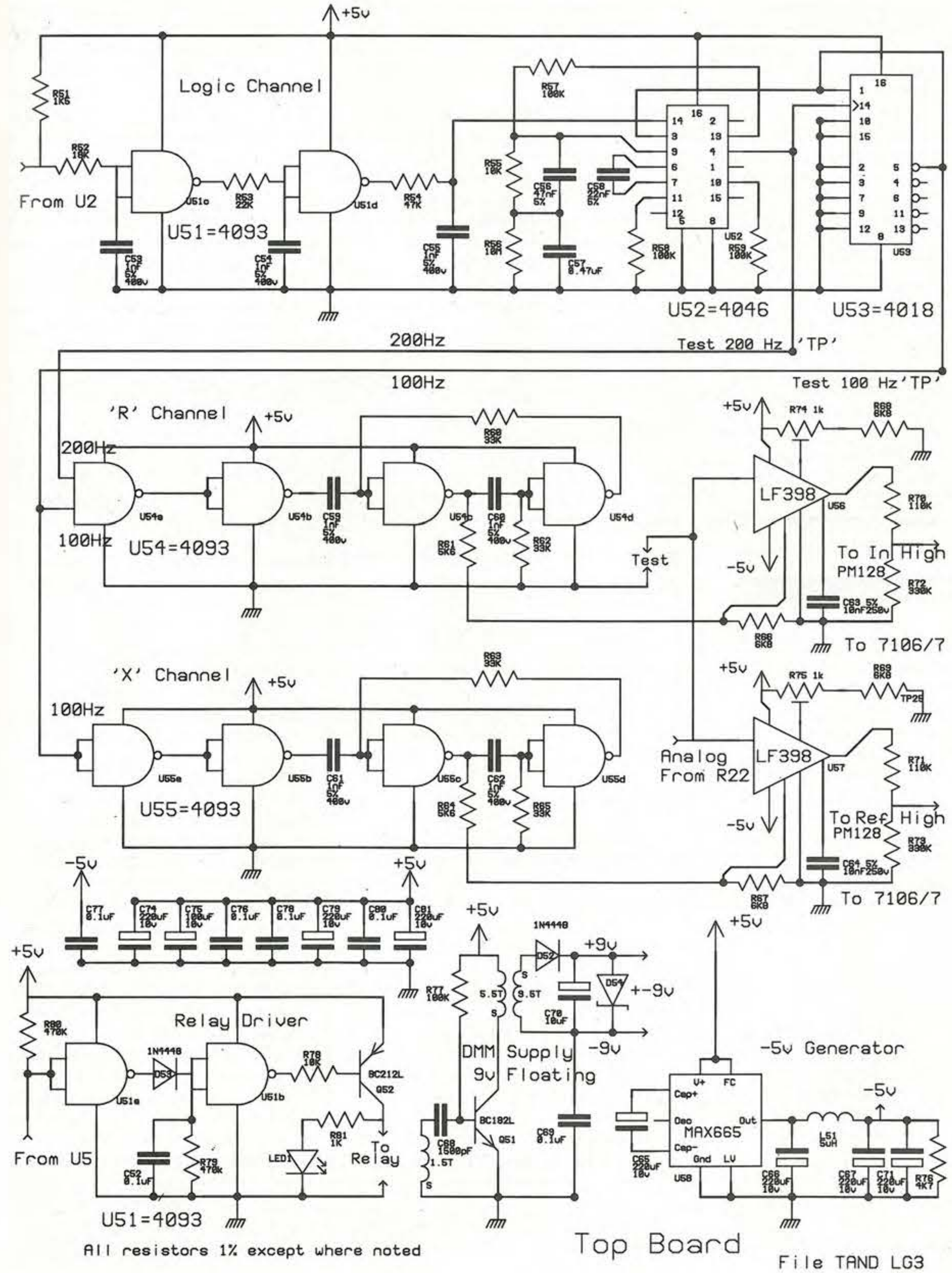
With the range of capacitance values and working voltages used in equipment, it is not possible to provide worst-case protection in a small portable instrument. The tanδ meter is designed and intended to be used on discharged capacitors. There's more on this in the panel entitled 'Worst-case protection'.

Accidental measurement of charged capacitors might however occur so I have provided as much protection as possible within the board size limitations.

Where other protection is not possible, I have used sacrificial fusible resistors. These become open circuit when overloaded, limiting the spread of damaged components and facilitating repairs to the meter.

Measurement circuit protection

Examination of the schematic diagram will reveal that four 1N4150 diodes as series connected back to back pairs, are used across the inputs of IC_1 and IC_3, to limit their differential voltage inputs. These ICs are the two Burr-



Brown INA118 instrument amplifiers previously mentioned. Using two diodes in series ensures that leakage currents do not effect the circuits normal operation, **Fig. 2a)**.

The INA118 amplifiers incorporate internal current limiting protection to ±40 volt inputs, both with and without power being applied.[4]

Enhanced common-mode input protection is provided by a diac to ground, from the junction of a 100Ω fusible resistor and a 3.3kΩ resistor in each INA118 input. These 100Ω fusible resistors become open circuit at around 1A peak. I evaluated many other protection options, but all were rejected. Some failed because leakage currents unbalanced the INA118 inputs. In other cases, photo-voltaic sensitivity to infra-red from mains lighting introduced an error hum voltage into the INA118.

If you are unfamiliar with diacs, they are four layer devices intended as low-cost triggers for triacs. Being four layer they exhibit similar 30V breakdown voltages regardless of polarity. Diacs have negligible capacitance and virtually no measurable leakage current at the meter's normal working voltages. The particular diacs used were selected because their blue coating inhibits the photo-voltaic problem. Also they can handle a 2A peak current.

These INA118 in-amps are by far the most expensive integrated circuits used in the meter. The low-cost diacs protect the INA118s by clamping their inputs to some 35V maximum. Ultimately, they blow open the four 100Ω fusible resistors to minimise further damage should a seriously charged capacitor be measured.

100Hz generator protection

The 10:1 step-down ratio of the LT700 transformer provides the low source impedance test current essential when measuring high value capacitors. Any voltage from a charged capacitor presented to its secondary winding, results in a voltage transient on its primary windings. Unconstrained, this transient could destroy the AD712 used as the 100Hz push-pull output amplifier.

To limit such voltage transients, two 1N4002 diodes $D_{13,14}$, are fitted across the secondary winding. These do not affect normal operation. Two pairs of back-to-back 3.9V zener diodes, D_{15-18} are also used to limit voltages on the LT700 primary windings,

The above measures localise any circuit damage in the event a charged capacitor is measured, but the two range resistors also the range relay contacts, cannot be protected.

Current sensing range resistors

The instant a charged capacitor, or a live voltage, is contacted by both test prods, this voltage appears at $PL_{1,2}$ and hence at one end of R_7 and the relay contact. The other end is held near 0 volts, via R_8 and the transformer, clamped by these 1N4002 diodes.

The meter defaults to its 'low' range, so initially the relay contacts are open. When subject to a charged capacitor of sufficient capacitance, the sustained discharge current causes the meter to switch to its 'high' range, closing the relay contacts which now protect R_7. A much increased discharge current passes through the relay contacts and R_8.

In normal use measuring discharged capacitors, the highest current passing through R_8 is 50mA. Practical tests, discharging a 1000μF capacitor, precharged to various voltages, directly into R_7 suggest this 22Ω resistor fails when subject to a single pulse of 2A peak. Similar discharge tests into R_8 indicate failures occurring at 9A peak.

Using a fuse between PL_1 and R_7 does not help. The fastest acting 'FF' fuses pass a large overload current for a finite time before they rupture. During this time, the range resistors and relay contacts are unprotected. Voltage from a charged capacitor causes the range resistors and relay contacts to pass many times their normal current.

Resistors $R_{7,8}$ can become open circuit and the relay contacts weld together before the fuse acts. The cold resistance of a fuse increases the series resistance of the generator circuit, reducing the maximum capacitance that the meter can measure.

Using a solid state, bi-directional, transient protection diode across PL_{1-4}, to quickly discharge the test capacitor, also does not help. The resistance of the test leads cannot drop significant voltage. The charged capacitor voltage still appears instantaneously at one end of R_7 and the relay, even though extremely large discharge current flows in the test leads.

Both methods have been evaluated experimentally using a test board with charged capacitors. Resistor R_8 failed at much the same test voltage level with and without these devices. Using two resistors in parallel each of double value, did not increase the sustainable voltage.

Continued on page 75...

Worst case protection

Aluminium electrolytic capacitors of several hundred microfarads are regularly used on rectified 230V AC mains in low-cost switched-mode power supplies.

While fully charged, such capacitors can store lethal amounts of energy. This energy is released almost instantaneously when contacted by the meter's test probes. Peak discharge currents easily exceed 100A. Measurement of such high voltage charged capacitors must be avoided.

Aluminium electrolytic capacitors are very low impedance devices. They can only be measured for impedance, tanδ or ESR using low-impedance measuring circuits. Adding resistance in the test circuit to restrict capacitor discharge current is simply not possible.

Using fuses in series with the test leads does not help. Even the fastest class 'FF' fuses pass a large overload current for a finite time before they rupture. During this time the measurement circuit is unprotected.

The alternative approach of 'crow-barring' the capacitor voltage to ground also does not help. The fastest semiconductor transient protection diodes take a finite time to turn on and the capacitor takes more time to discharge. At the instant both test probes contact the charged capacitor, that voltage is instantaneously transferred to the measurement circuit.

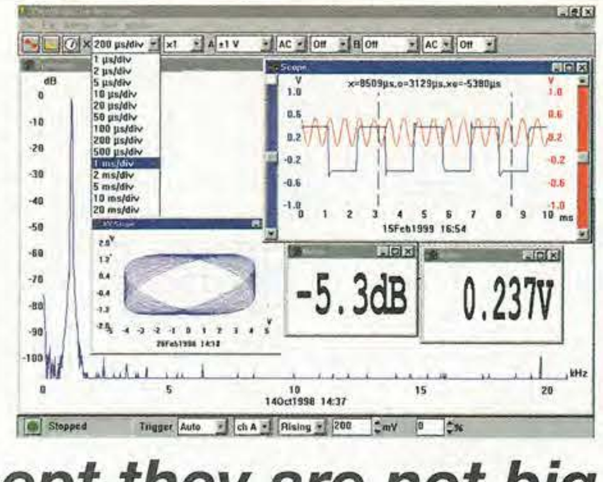
I have experimentally confirmed these effects by discharging 1000μF capacitors, charged at various voltages up to 50 volts, into the fusible resistors mounted in a test board. Using both series 'FF' fuses and shunt 5.8V bi-directional 140A-rated transient-diode protectors, the 22Ω and 2.2Ω. 0.5W fusible resistors still become open circuit.

The 2.2Ω fusible resistors typically fail at 20 to 25V, the 22Ω failing near 50V. Similar results were found testing these resistors both with and without the series fuse and shunt transient diode protector circuits.

Using lesser, 57A-rated transient diode protectors, the high peak discharge currents destroyed the protection diode, in the process damaging my nickel-plated test probes.

Since these fuse and transient diodes did not enhance the charged capacitor protection, they are not used in the final meter design.

They are exactly the same as big, expensive oscilloscopes...



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HP141T+ 8552B IF + 8555A 10 MC/S-18GHzS - £1000.
HP8443A Tracking Gen Counter 100KHz-110Mc/s - £200
HP8445B Tracking Preselector DC to 18GHz - £250.
HP8444A Tracking Generator • 5-1300Mc/s - £450.
HP8444A OPT 059 Tracking Gen • 5-1500Mc/s - £650.
HP35601A Spectrum Anz Interface - £300.
HP4953A Protocol Anz - £400.
HP8970A Noise Figure Meter + 346B Noise Head - £3k.
HP8755A+B+C Scalar Network Anz Pl - £250 + MF 180C - Heads 11664 Extra - £150 each.
HP3709B Constellation ANZ £1,000.
HP11715A AM-FM Test Source - £350.
FARNELL TVS70MKII PU 0-70V 10 amps - £150.
MARCONI 6500 Network Scaler Anz - £500. Heads available to 40GHz many types in stock.
Mixers are available forANZs to 60GHz.
HP6131C Digital Voltage Source + -100V½ Amp.
HP5316A Universal Counter A+B.
Marconi TF2374 Zero Loss Probe - £200.
Racal/Dana 2101 Microwave Counter - 10Hz-20GHz - with book as new £2k.
Racal/Dana 1250-1261 Universal Switch Controller + 200Mc/s Pl Cards and other types.
Racal/Dana 9303 True RMS Levelmeter + Head - £450.
TEKA6902A also A6902B Isolator - £300-£400.
TEK CT-5 High Current Transformer Probe - £250.
HP Frequency comb generator type 8406 - £400.
HP Sweep Oscillators type 8690 A+B + plug-ins from 20Mc/s to 18GHz also 18-40GHz.
HP Network Analyser type 8407A + 8412A + 8601A - 100Kc/s - 110Mc/s - £500 - £1000.
HP 8410-A-B-C Network Analyser 110Mc/s to 12 GHz or 18 GHz - plus most other units and displays used in this set-up - 8411a-8412-8413-8414-8418-8740-8741-8742-8743-8746-8650. From £1k.
Racal/Dana 9301A-9302 RF millivoltmeter - 1.5-2GHz - qty in stock £250-£400.
Racal/Dana Modulation Meter Type 9009-9008 - 8Mc/s - 1.5GHz - £150/£250 - 9009A £350.
Marconi RCL Bridge type TF2700 - £150.
Marconi Microwave 6600A 1 sweep osc., mainframe with 6650PI - 18-26.5 GHz or 6651 PI - 26.5-40GHz-£750 or PI only £600. MF only £250.
Gould J3B test oscillator + manual - £150.
Marconi 6155A Signal Source-1 to 2GHz - LED - £400.
Barr & Stroud Variable filter EF3 0.1Hz-100Kc/s + high pass + low pass - £150, other makes in stock.
Racal/Dana 9300 RMS voltmeter - £250.
HP 8750A storage normalizer - £400 with lead + S.A. or N. A
Marconi mod meters type TF2304 - £250 - TF2305 - £1,000.
Racal/Dana counters-99904-9905-9906-9915-9916-9917-9921-50Mc/s-3GHz - £100 - £400 - all fitted with FX standards.
HP180TR. HP181T, HP182T mainframes £300 - £500.
HP432A-435A or B-436A-power meters + powerheads to 60GHz - £150 - £1750 - spare heads available.
HP3586A or C selective level meter - £500.
HP86222A+B Sweep Pl -01-2.4GHz + ATT £1000-£1250.
HP86290A+B Sweep Pl-2 - 18GHz - £1000 - £1250.
HP8620C Mainframe - £250. IEEE £350.
HP8165A Programmable signal source - 1MHZ - 50Mc/s - £1k.
HP3455/3456A Digital voltmeter - £400.
HP5370A Universal time interval counter - £1k.
HP5335A Universal counter - 200Mc/s-£1000.
HP3552A Transmission test set - £350.
TEKTRONIX 577 Curve tracer + adaptors - £900.
TEKTRONIX 1502/1503 TDR cable test set - £400.
HP8699B Sweep PI YIG oscillator .01 - 4GHz - £300. 8690B MF-£250. Both £500.
Dummy Loads & Power att up to 2.5 kilowatts FX up to 18GHz - microwave parts new and ex equipt - relays - attenuators - switches - waveguides - Yigs - SMA - APC7 plugs - adaptors etc. qty. in stock.
B&K Items in stock - ask for list.
Power Supplies Heavy duty + bench in stock - Farnell - HP - Weir - Thurlby - Racal etc. Ask for list. Large quantity in stock, all types to 400 amp - 100Kv.
HP8405A Vector voltmeter - late colour - £400.
HP8508A Vector voltmeter - £2500.
HP8505A Network Anz 500KHz-1.3GHz - £1000.
HP8505A + 8502A or 8503A test sets- £1200 - £1500.
HP8505A + 8502A or 8503A + 8501A normalizer - £1750-£2000.
Phillips 3217 50Mc/s oscilloscopes - £150-£250.
Wavetek-Schlumberger 4031 Radio communication test set

LIGHT AND OPTICAL EQUIPMENT
Anritsu ML93A & Optical Lead Power Meter - £250.
Anritsu ML93B & Optical Lead Power Meter - £350.
Power Sensors for above **MA96A - MA98A - MA913A** - Battery Pack **MZ95A.**
Anritsu MW97A Pulse Echo Tester.
Pl available - **MH914C 1.3 - MH915B 1.3 - MH913B 0.85 - MH925A 1.3 - MH929A 1.55 - MH925A 1.3GI - MH914C 1.3SM - £500 + one P.I.**
Anritsu MW98A Time Domain Reflector.
Pl available - **MH914C 1.3 - MH915B 1.3 - MH913B 0.85 - MH925A 1.3 - MH929A 1.55 - MH925A 1.3GI - MH914C 1.3SM - £500 + one P.I.**
Anritsu MZ100A E/O Converter.
+ MG912B (LD 1.35) Light Source + MG92B (LD 0.85) Light Source £350.
Anritsu MZ118A O/E Converter.
+MH922A 0.8 O/E unit **+ MH923 A1.3** O/E unit £350.
Anritsu ML96B Power Meter & Charger £450.
Anritsu MN95B Variable Att. 1300 £100.
Photo Dyne 1950 XR Continuous Att. 1300 - 1500 £100.
Photo Dyne 1800 FA. Att £100.
Cossor-Raytheon 108L Optical Cable Fault Locator 0-1000M 0-10KM £200.
TEK P6701 Optical Converter 700 MC/S-850 £250.
TEK OF150 Fibre Optic TDR - £750.
HP81512A Head 150MC/S 950-1700 £250.
HP84801A Fibre Power Sensor 600-1200 £250.

HP8158B ATT OPT 002+011 1300-1550 £300.
HP81519A RX DC-400MC/S 550-950 £250.
STC OFR10 Reflectometer – £250.
STC OFSK15 Machine jointing + eye magnifier – £250.

COMMUNICATION EQUIPMENT
Anritsu ME453L RX Microwave ANZ – £350.
Anritsu ME453L TX Microwave ANZ – £350.
Anritsu MH370A Jitter Mod Oscillator – £350.
Anritsu MG642A Pulse Patt Gen. £350.
System MS02A Timer & Digital Printer – £500.
Complete MS65A Error Detector.
Anritsu ML612A Sel Level Meter – £400.
Anritsu ML244A Sel Level Meter – £300.
W&G PCM3 Auto Measuring Set – £300.
W&G SPM14 Sel Level Meter – £300.
W&G SPM15 Sel Level Meter – £350.
W&G PS19 Level Gen – £500.
W&G DA20+DA1 Data ANZ £400.
W&G PMG3 Transmission Measuring Set – £300.
W&G PSS16 Generator – £300.
W&G PS14 Level Generator – £350.
W&G EPM-1 Plus Head Milliwatt Power Meter – £450.
W&G DLM3 Phase Jitter & Noise – £350
W&G DLM4 Data Line Test Set – £400.
W&G PS10 & PM10 Level Gen. – £250.

MISCELLANEOUS ITEMS
HP 3852A Data Acquisition Control Unit + 44721A 16ch input £1,000.
HP 4261 LCR meter – £650.
HP 4274 FX LCR meter – £1,500.
HP 4951A Protocol ANZ – £500.
HP 3488 Switch Control Unit + PI Boards – £500.
HP 75000 VXI Bus Controllers + E1326B-DVM-quantity.
HP 83220A GSM DCS/PCS 1805-1990MC/S convertor for use with 8922A – £2,000.
HP 1630-1631-1650 Logic ANZ's in stock.
HP 8754A Network ANZ 4-1300MC/S + 8502A + cables – £1,500.
HP 8754A Network ANZ H26 4-2600MC/S + 8502A + Cables – £2,000.
HP 8350A Sweeper MF + 83540A PI 2-8.4GHZ + 83545A PI 5.9-12.4GHZ all 3 – £3,500.
HP MICROWAVE TWT AMPLIFIER 489A 1-2GHZ-30DB – £400.
HP PREAMPLIFIER 8447A 0.1-400MC/S – £200. Dual – £300.
HP PREAMPLIFIER 8447D 0.01-1.3GHZ – £400.
HP POWER AMPLIFIER 8447E 0.01-1.3GHZ – £400.
HP PRE + POWER AMPLIFIER 8447F 0.01-1.3GHZ – £500.
HP 3574 Gain-Phase Meter 1HZ-13MC/S OPT 001 Dual – £400.
MARCONI 2305 Modulation Meter-50KHZ-2.3 GHZ – £1,000.
MARCONI 2610 True RMS Meter – £450.
MARCONI 893B AF Power Meter (opt Sinad filter) – £250-£350.
MARCONI 6950-6960B Power Meters + Heads – £400-£900.
MARCONI SIGNAL SOURCE-6055-6056-6057-6058-6059 – FX Range 4-18GHZ– £250-£400.
RACAL 1792 COMMUNICATION RX – £500 early – £1,000 – late model with back lighting and byte test.
RACAL 1772 COMMUNICATION RX – £400-£500.
PLESSEY PR2250 A-G-H COMMUNICATION RX – £500-£900.
TEK MODULE MAINFRAMES – TM501-502-503-504-506-TM5003-5006.
TEK PI 5010-M1 – Prog Multi Interface – £250. **FG** Prog 20MC/S Function Gen – £400 – **S1** Prog Scanner – £250 – **DM** Prog DMM – £400.
TEK 7000 OSCILLOSCOPE MAINFRAMES – 7603-7623-7633-7834-7854-7904-7904A-7104 – £150-£1,000.
TEK 7000 PI's – 7A11-7A12-7A13-7A18-7A19-7A22-7A24-7A26-7A29-7A42-7B10-7B15-7B53A-7B80-7B85-7B92A-7D15-7D20.
TEK 7000 – 7S11-7S12-7S14-7M11-S1-S2-S3A-S4-S5-S6-S51-S53-S54.
HP POWER SUPPLIES – 6621A-6623A-6624A-6632A-6652A. Qtys available. Also 6000 types £POA.

RADIO COMMUNICATION TEST SETS
BULK PURCHASE ONLY FROM JOHNS RADIO
HP 8920A RF Communication Test Sets – Opts available 001-3-4-5-7-11-12-14-H13-K13. £1,500-£1,750.
HP8920A with opt 002 Spectrum anz plus tracking generator plus opts. 001-3-4-5-11-12-014 available in part includes syn sig generator - digital oscilliscope distortion meter - mod meter - RF power meter etc. £2,500.

SPECIAL OFFERS
MOTOROLA R2600A plus **RLN4260A** RF Test Set – £1,800.
MARCONI 2955 RF Test Sets-1000MC/S – £1,000 each.
MARCONI 2958 RF Test Sets-1000MC/S – £1,000 each.
MARCONI 2960 RF Test Sets-1000MC/S – £1,250 each.
MARCONI 2955A RF Test Sets-1000MC/S – £1,500 each.
MARCONI 2960A RF Test Sets-1000MC/S – £1,500 each.
ANRITSU MS555A2 Radio Comm Anz-1000M/Cs – £750 each.
MARCONI 2019A SYNTHESIZED SIGNAL GENERATORS – 80KC/S-1040MC/S – AM-FM – £400 inc. instruction book – tested.
MARCONI 2022E SYNTHESIZED SIGNAL GENERATOR – 10KC/S-1.01GHZ AM-FM – £500 inc. instruction book – tested.
R&S APN 62 LF Sig Gen 0.1Hz – 260KHz c/w book – £250.

WE KEEP IN STOCK HP and other makes of RF Frequency doublers which when fitted to the RF output socket of a S/Generator doubles the output frequency EG.50-1300MC/S to 50-2600MC/S price from £250 – £450 each.

SPECTRUM ANALYZERS
HP 3580A 5HZ-50KHZ – £750.
HP 3582A Dual 0.2HZ-25.5KHZ – £1,500.
HP 3585A 20HZ-40MC/S – £3,500.
HP 3588A 10HZ-150MC/S – £7,500.
HP 8568A 100HZ-1.5GHZ – £3,500.
HP 8568B 100HZ-1.5GHZ – £4,500.
HP 8590B 9KC/S-1.8GHZ – £4,500.
HP 8569B 10MC/S (0.01-22GHZ) – £3,500.
HP 3581A Signal Analyzer 15HZ-50KHZ – £400.

TEK491 10MC/S-12.4GHZ + 12.4-40GHZ – £500.
TEK492 50KHZ-21GHZ OPT 2 – £2,500.
TEK492P 50KHZ-21GHZ OPT 1-2-3 – £3,500.
TEK492AP 50KHZ-21GHZ OPT 1-2-3 – £4,000.
TEK492BP 50KHZ-21GHZ – £3,000-£4,000.
TEK495 100KHZ-1.8GHZ – £2,000.
HP 8557A 0.01MC/S-350MC/S – £500 + **MF180T** or **180C** – £150 – **182T** – £500.
HP 8558B 0.01-1500MC/S – £750 – **MF180T** or **180C** – £150 – **182T** – £500.
HP 8559A 0.01-21GHZ – £1,000 – **MF180T** or **180C** – £150 – **182T** – £500.
HP 8901A AM FM Modulation ANZ Meter – £800.
HP 8901B AM FM Modulation ANZ Meter – £1,750.
HP 8903A Audio Analyzer – £1,000.
HP 8903B Audio Analyzer – £1,500.
MARCONI 2370 SPECTRUM ANALYZERS – HIGH QUALITY – DIGITAL STORAGE – 30HZ-110MC/S Large qty to clear as received from Gov – all sold as is from pile complete or add £100 for basic testing and adjustment – callers preferred – pick your own from over sixty units – discount on qtys of five.
A EARLY MODEL GREY – horizontal alloy cooling fins – £200.
B LATE MODEL GREY – vertical alloy cooling fins – £300.
C LATE MODEL BROWN – as above (few only) – £500.

OSCILLOSCOPES
TEK 465-465B 100MC/S + 2 probes – £250-£300.
TEK 466 100MC/S storage + 2 probes – £200.
TEK 475-475A 200MC/S-250MC/S + 2 probes – £300-£350.
TEK 2213-2213A-2215-2215A-2224-2225-2235-2236-2245-60-100MC/S – £250-£400.
TEK 2445 4ch 150MC/S + 2 probes – £450.
TEK 2445A 4ch 150MC/S + 2 probes – £600.
TEK 2445B 4ch 150MC/S + 2 probes – £750.
TEK 468 D.S.O. 100MC/S + 2 probes – £500.
TEK 485 350MC/S + 2 probes – £550.
TEK 2465 4ch-300MC/S – £1,150.
TEK 2465A 4ch-350MC/S – £1,550.
TEK 2465ACT 4ch-350MC/S – £1,750.
TEK D.S.O. 2230 -100MC/S + 2 probes – £1,000.
TEK D.S.O. 2430 -150MC/S + 2 probes – £1,250.
TEK D.S.O. 2430A -150MC/S + 2 probes – £1,750.
TEK D.S.O. 2440 -300MC/S + 2 probes – £2,000.
TEK TAS 475-485 -100MC/S-200MC/S-4 ch + 2 probes – £900-£1.1K.
HP1740A – 100MC/S + 2 probes – £250.
HP1741A – 100MC/S storage + 2 probes – £200.
HP1720A – 1722A – 1725A – 275MC/S + 2 probes – £300-£400.
HP1744A – 100MC/S storage – large screen – £250.
HP1745A – 1746A – 100MC/S – large screen – £350.
HP54100A – 1GHz digitizing – £500.
HP54200A – 50MC/S digitizing – £500.
HP54501A – 100MC/S digitizing – £500.
HP54100D – 1GHZ digitizing – £1,000.

MICROWAVE COUNTERS – ALL LED READOUT
EIP 351D Autohet 20Hz-18GHz – £750.
EIP 371 Micro Source Locking – 20Hz-18GHz – £850.
EIP 451 Micro Pulse Counter – 300MC/S-18GHz – £700.
EIP 545 Microwave Frequency Counter – 10Hz-18GHz – £1K.
EIP 548A Microwave Frequency Counter – 10HZ-26.5GHz – £1.5k.
EIP 575 Microwave Source Locking – 10Hz-18GHz – £1.2K.
EIP 588 Microwave Pulse Counter – 300MC/S-26.5GHz – £1.4K.
SD 6054B Micro Counter 20HZ-24GHZ – SMA Socket – £800.
SD 6054B Micro Counter 20HZ-18GHZ – N Socket – £700.
SD 6054D Micro Counter 800MC/S-18GHz – £600.
SD 6246A Micro Counter 20Hz-26GHz – £1.2K.
SD 6244A Micro Counter 20Hz-4.5GHz – £400.
HP5352B Micro Counter OPT 010-005-46GHz – new in box – £5k.
HP5340A Micro Counter 10HZ-18GHz – Nixey – £500.
HP5342A Micro Counter 10HZ-18-24GHz – £800-£1K – OPTS 001-002-003-005-011 available.
HP5342A + 5344S Source Synchronizer – £1.5K.
HP5345A 500MC/S 11 Digit LED Readout – £400.
HP5345A + 5354A Plugin – 4GHz – £700.
HP5345A + 5355A Plugin with 5356A 18GHz Head – £1K.
HP5385A 1GHz 5386A-5386A 3GHz Counter – £1K-£2K.
Racal/Dana Counter 1991-160MC/S – £200.
Racal/Dana Counter 1992-1.3GHz – £600.
Racal/Dana Counter 9921-3GHz – £350.

SIGNAL GENERATORS
HP8640A – AM-FM 0.5-512-1024MC/S – £200-£400.
HP8640B – Phase locked – AM-FM-0.5-512-1024MC/S – £500-£1.2K. Opts 1-2-3 available.
HP8654A – B AM-FM 10MC/S-520MC/S – £300.
HP8656A SYN AM-FM 0.1-990MC/S – £900.
HP8656B SYN AM-FM 0.1-990MC/S – £1.5K.
HP8657A SYN AM-FM 0.1-1040MC/S – £2K.
HP8657B SYN AM-FM 0.1-2060MC/S – £3K.
HP8660C SYN AM-FM-PM-0.01-1300MC/S-2600MC/S – £2K.
HP8660D SYN AM-FM-PM-0.01-1300MC/S-2600MC/S – £3K.
HP8673D SYN AM-FM-PM-0.01-26.5 GHz – £12K.
HP3312A Function Generator AM-FM 13MC/S-Dual – £300.
HP3314A Function Generator AM-FM-VCO-20MC/S – £600.
HP3325A SYN Function Generator 21MC/S – £800.
HP3325B SYN Function Generator 21MC/S – £2K.
HP8673-B SYN AM-FM-PH 2-26.5 GHz – £6.5K.
HP3326A SYN 2CH Function Generator 13MC/S-IEEE – £1.4K.
HP3336A-B-C SYN Func/Level Gen 21MC/S – £400-£300-£500.
Racal/Dana 9081 SYN S/G AM-FM-PH-5-520MC/S – £300.
Racal/Dana 9082 SYN S/G AM-FM-PH-1.5-520MC/S – £400.
Racal/Dana 9084 SYN S/G AM-FM-PH-.001-104MC/S – £300.
Racal/Dana 9087 SYN S/G AM-FM-PH-.001-1300MC/S – £1K.
Marconi TF2008 AM-FM-Sweep 10KC/S-510MC/S – £200 Fully Tested to £300, as new + book + probe kit in wooden box.
Marconi TF2015 AM-FM-10-520MC/S – £100.
Marconi TF2016A AM-FM 10KC/S-120MC/S – £100.
Marconi TF2171/3 Digital Synchronizer for 2015/2016A – £50.
Marconi TF2018A AM-FM SYN 80KC/S-520MC/S – £500.
Marconi TF2019A AM-FM SYN 80KC/S-1040MC/S – £650-£1K.
Marconi TF2022E AM-FM SYN 10KC/S-1.01GHz – £1K-£1.2K.
R & S SMPD AM-FM-PH 5KHz-2720MC/S – £3K.
Anritsu MG3601A SYN AM-FM 0.1-1040MC/S – £1.2K.

ITEMS BOUGHT FROM HM GOVERNMENT BEING SURPLUS. PRICE IS EX WORKS. SAE FOR ENQUIRIES. PHONE FOR APPOINTMENT OR FOR DEMONSTRATION OF ANY ITEMS, AVAILABILITY OR PRICE CHANGE. VAT AND CARRIAGE EXTRA. ITEMS MARKED TESTED HAVE 30 DAY WARRANTY. WANTED: TEST EQUIPMENT-VALVES-PLUGS AND SOCKETS-SYNCROS-TRANSMITTING AND RECEIVING EQUIPMENT ETC.

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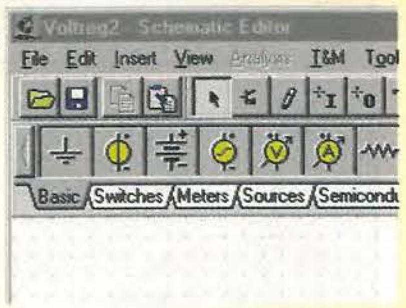

Are you familiar with CAD? If not, you may think that you still need to get involved with writing unintuitive listings. Not so. Drag and drop analogue components and digital ICs into your circuit diagram then see how your circuit performs using simulated instruments including function generator, multimeter, XY plotter, oscilloscope, signal analyser, logic analyser and digital signal generator. Once you've found your way around, you'll soon be including things like noise and distortion analyses in your summaries.

Feature outline

- Fully 32bit software for fast performance under Win 95, 98 or NT
- 10000+ built in components and 1000+ manufacturer made components in SPICE sub-circuit format
- DC analysis includes nodal voltages, DC transfer characteristic and temperature analysis
- AC analysis includes nodal voltages, time function, and AC transfer characteristic
- Symbolic analysis includes DC and AC result, semi-symbolic DC and AC result, AC transfer, semi-symbolic AC transfer, Poles and Zeros , and semi-symbolic transient.
- Other analyses include transient, digital step by step, digital timing analysis with glitch control, and noise analysis
- Bode plots, Nyquist diagrams, poles and zeros, transient responses, temp. sweeps, etc in the diagram window.
- Virtual instruments include function generator, multimeter, XY plotter, oscilloscope, signal analyser, logic analyser and digital signal generator.
- Enhanced SPICE 3F5 compatible algorithms
- Massive range of component models including resistor, potentiometer, capacitor, charged capacitor, inductor, energy storing inductor, coupled inductors, diode, Zener, LED, lamp, motor, transformer, transmission line, bipolar transistor (NPN/PNP), MOS transistor (enhancement and depletion mode, N and P channel), JFET (N and P channel), thyristor, triac, diac, ideal OPAMP, OPAMP, current source, voltage source, current generator, voltage generator, controlled sources (CCCS, CCVS, CCVS, VCVS), and a full complement of digital components, AD and DA converters, timers, analogue control blocks, seven segment display, keypads, nonlinear coils, transformers, relays, darlington transistors, optocouplers and voltage regulators and much much more..
- New sub-circuits can contain both SPICE macros or schematics
- Automatic shape creation for subcircuits
- Schematic symbol editor for creating custom components
- Customisable component toolbar – add your own components
- SPICE macros can be used directly from Tina's schematic editor
- Parameter extractor program to calculate model parameters from catalogue or measurement data
- Library manager for extending and maintaining manufacturer made component libraries
- New multimeter, oscilloscope and signal analyser components can be added directly onto schematics
- Average and RMS values for arbitrary periodic waveforms
- Enhanced Fourier analysis with both peak and RMS calculation
- AC analysis now with RMS calculation
- Conversion of symbolic analysis results into code executable in the interpreter
- Specify temperature for individual components
- New fault simulation on digital circuits
- Multimedia tutorials demonstrate key concepts and features
- Supplied on CD-ROM with printed documentation.
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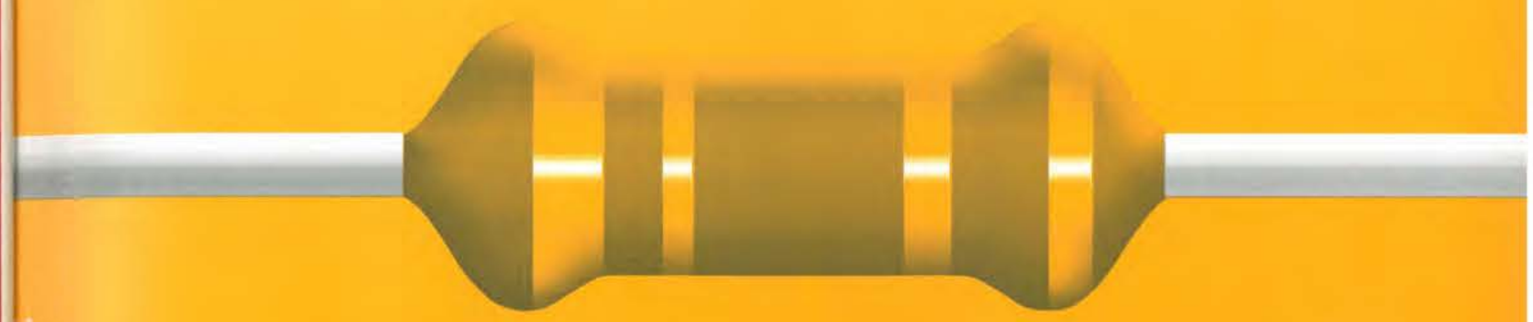

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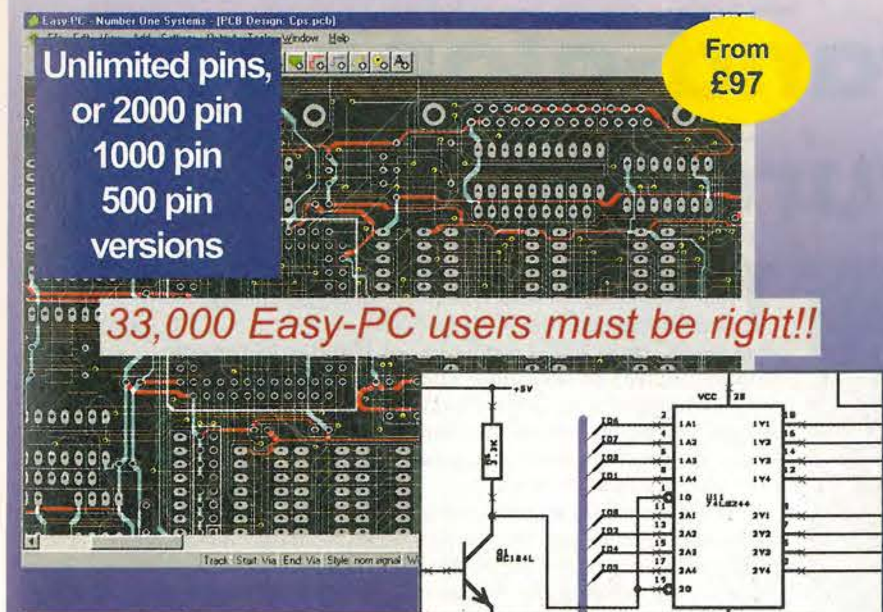

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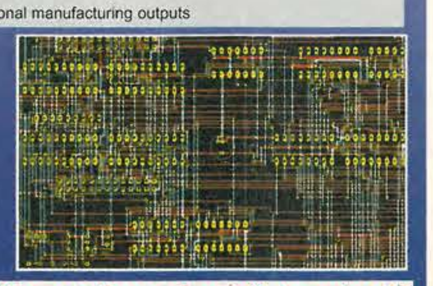

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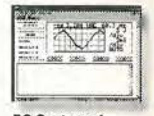

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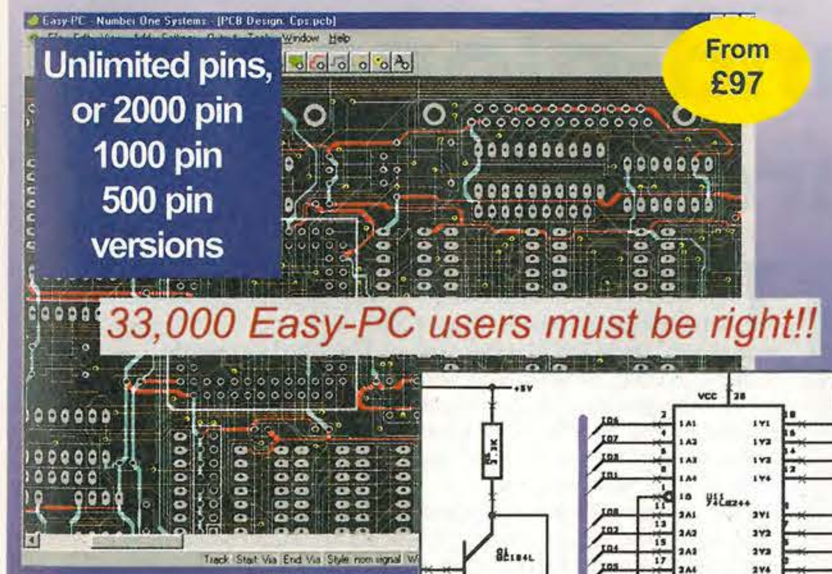

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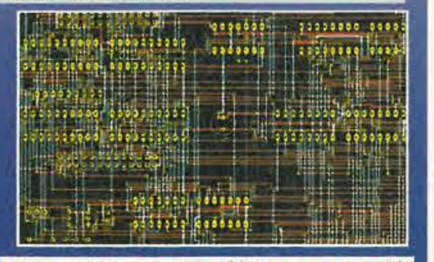

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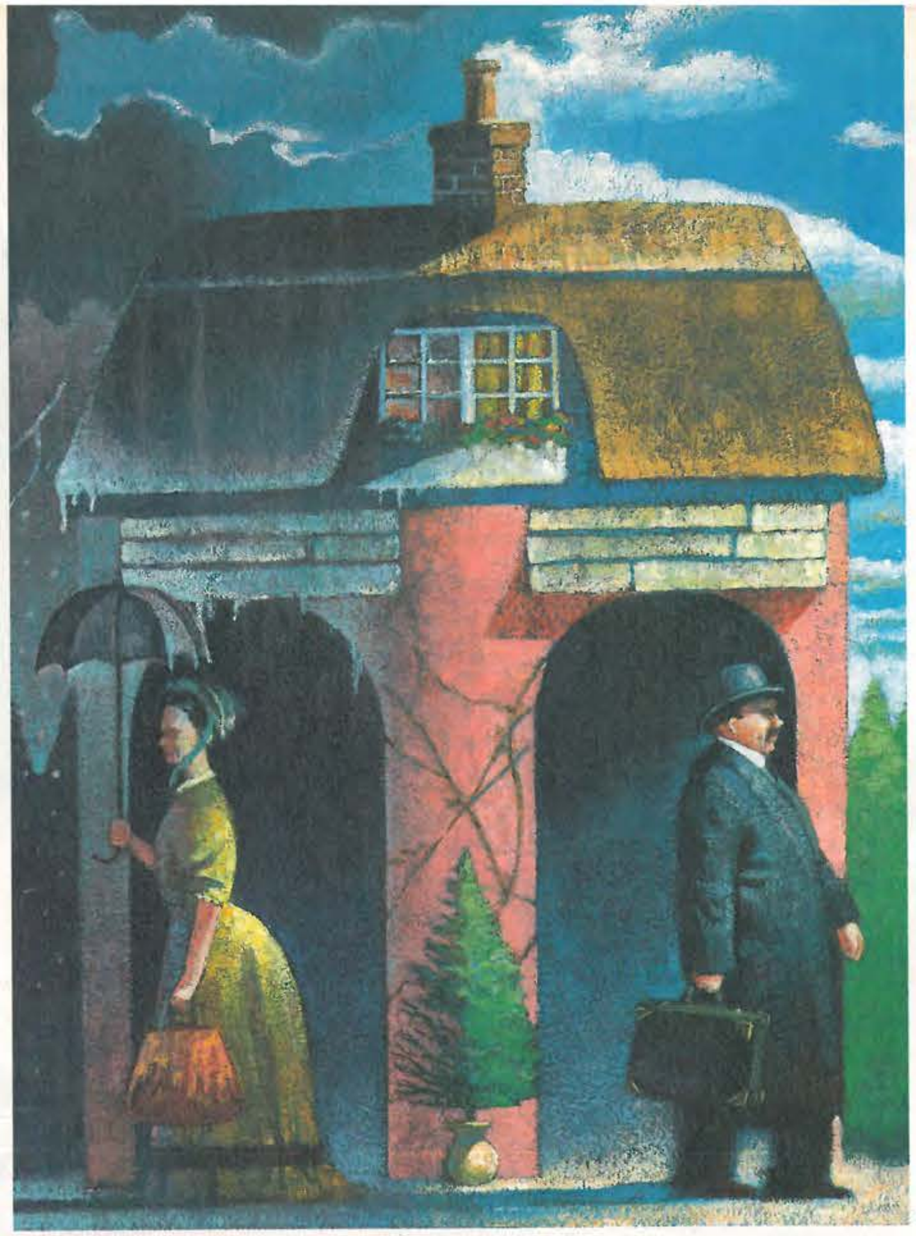
A temperature sensing IC with digitiser, RTC, memory and serial i/o makes the hardware needed for a stand-alone temperature logger exceedingly simple. Pei An's low-power design covers –40 to 85°C, feeding the data into a PC at the end of a session so for analysing and storage via a spreadsheet for example.



One IC stand-alone temperature logger

Fig. 1. Temperature logging is made simple – and small – using a sensor with integral digitising, control and serial i/o.

This stand-alone temperature logger incorporates a Dallas DS1615 temperature recorder IC. It connects to a PC via its RS232 port for initialising and data downloading. After initialisation, the logger is disconnected and placed in the environment where temperature needs recording, ready to make measurements.

If measurements do not need to start right away, the logger can be programmed to sleep for a while before it starts to log temperature. Scanning interval can be varied from 1 minute to 255 minutes. Up to 2048 temperature records can be stored in the on-board memory.

Temperature to be measured is in the range –40 to 85°C with an error of ±2°C. In my version, power is supplied by a 3V lithium button cell, which should last for three years or so. My implementation is shown in **Fig. 1**.

The hardware side

Figure 2 shows the circuit diagram, which includes the DS1615, two transistors and a number of passive components. Functionally, the circuit divides into two sections, namely the data acquisition/storage section and the RS232 transceiver section.

The data acquisition/storage section is based on the DS1615. This is an intelligent IC, capable of receiving various control commands from its built-in universal asynchronous receiver and transmitter, or UART, port. Once the DS1615 is initialised properly, it becomes a stand-alone temperature logger.

The asynchronous port on the

DS1615 is a standard UART i/o port. Buffering is needed to make the UART's inputs and outputs compatible with the PC's COM port input and output lines. While the unit is connected to the RS232 port during initialisation and data downloading, the PC provides a +5V supply rail to the temperature logger circuit.

Pressing the switch starts temperature logging. Diode D_4 flashes four times once a logging mission is started. An optional low-current piezo-electric sounder can also gives four beeps to signify the start of a logging session.

In my unit, the on-board battery is a CR1616 3V lithium button cell with a capacity of typically 50mAh.

Overview of the DS1615

Dallas's DS1615 is a highly-integrated single-chip temperature recorder, **Fig. 3a**). It comprises a temperature sensor, an analogue-to-digital converter, a real-time clock, a non-volatile memory and a serial interface, Fig. **3b**).

Temperatures from –40 to 85°C in 0.5°C increments are read by the device with a measurement accuracy of ±2°C. The real time clock/calendar is in BCD format and counts seconds, minutes, hours, date, month, day of the week and year with leap year compen-

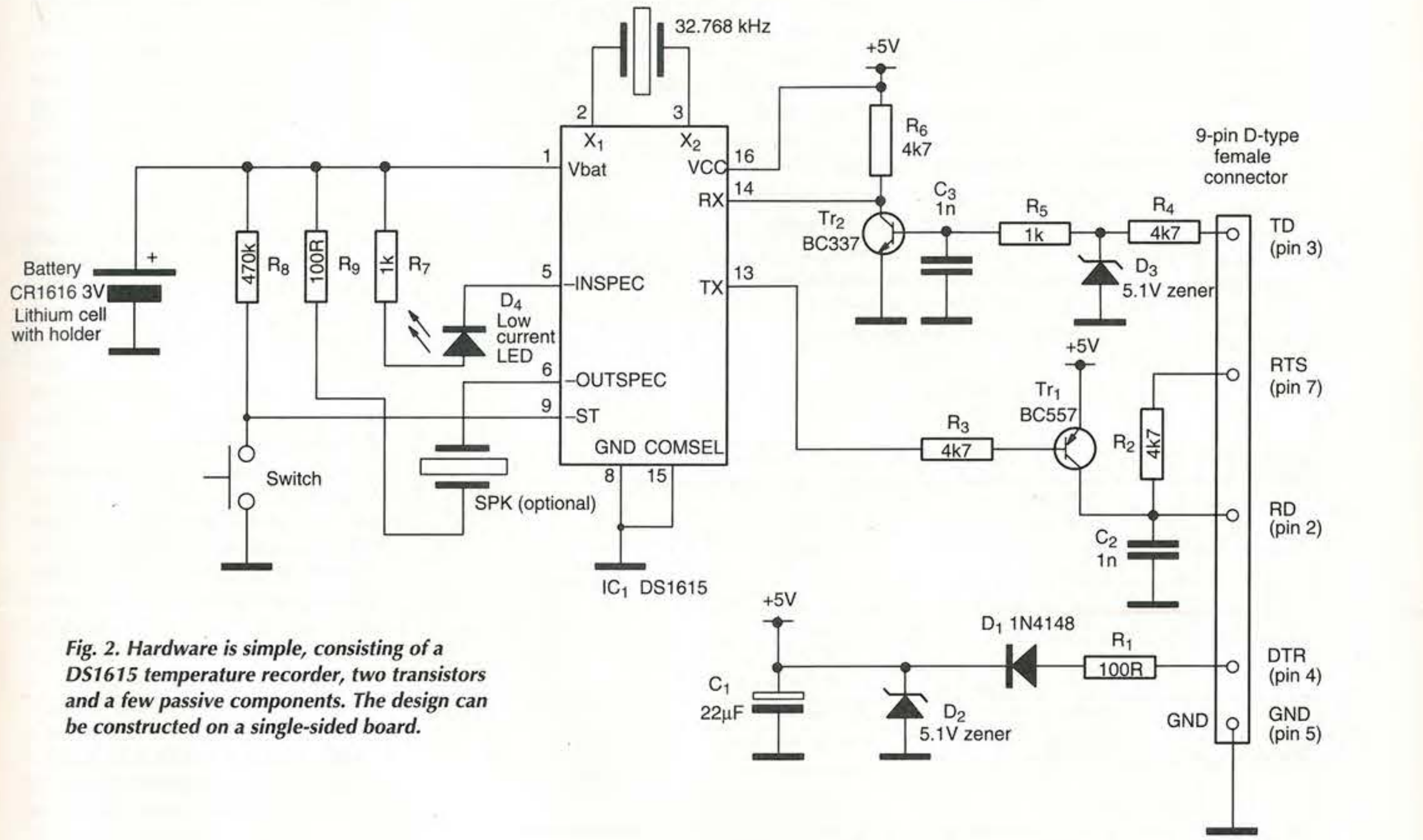
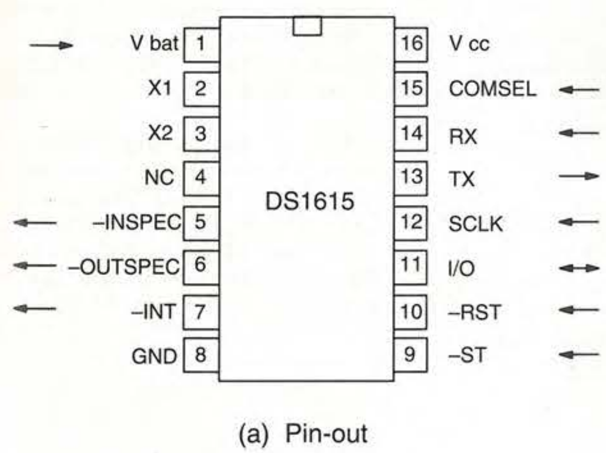
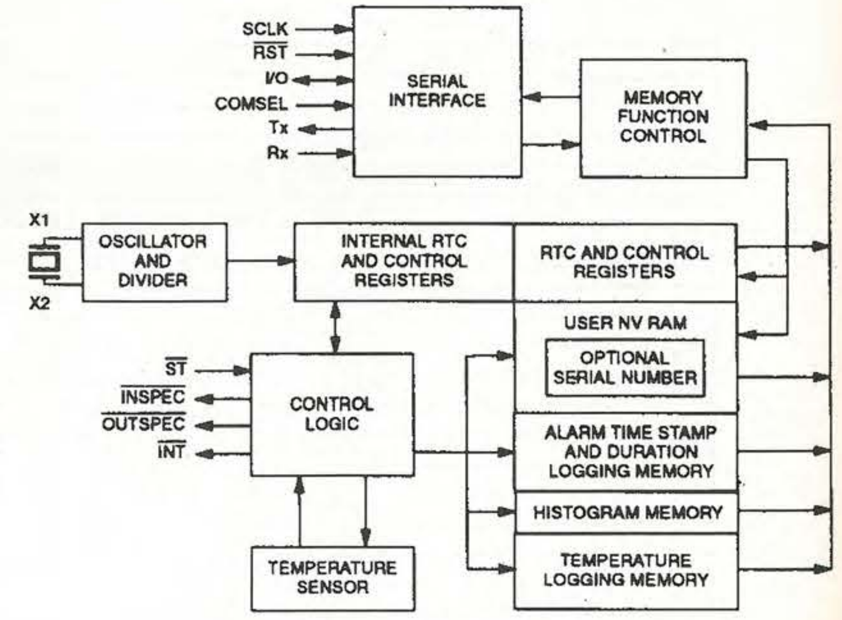


Fig. 2. Hardware is simple, consisting of a DS1615 temperature recorder, two transistors and a few passive components. The design can be constructed on a single-sided board.

Fig. 3. The DS1615 is a 16-pin device as shown in a). Its internal block diagram is shown in b).



(a) Pin-out



(b) Internal block diagram

Address	Contents	Page
0000H TO 001FH	RTC AND CONTROL REGISTERS	PAGE 0
0020H TO 003FH	(RESERVED)	PAGE 1
0040H TO 005FH	USER NV RAM	PAGE 2
0060H TO 0217H	(RESERVED FOR FUTURE EXTENSIONS)	PAGE 3 TO PAGE 16 (EXCLUDING LAST 8 BYTES OF PAGE 16)
0218H TO 021FH	SERIAL NUMBER (OPTIONAL)	PAGE 16 (LAST 8 BYTES)
00220H TO 027FH	ALARM TIME STAMPS AND DURATIONS	PAGE 17 TO PAGE 19
0280H TO 07FFH	(RESERVED FOR FUTURE EXTENSIONS)	PAGES 20 – 63
0800H TO 087FH	TEMPERATURE HISTOGRAM (63 BINS OF 2 BYTES EACH)	PAGE 64 TO PAGE 67
0880H TO 0FFFH	(RESERVED FOR FUTURE EXTENSIONS)	PAGES 68 – 127
1000H TO 17FFH	TEMPERATURE DATALOG MEMORY (64 PAGES)	PAGE 128 TO PAGE 191
1800H AND HIGHER	(RESERVED FOR FUTURE EXTENSIONS)	PAGE 192 AND HIGHER

ADDR.	BIT 7	BIT 6	BIT 5	BIT 4	BIT 3	BIT 2	BIT 1	BIT 0	FUNCTION
00	0	10 Seconds			Single Seconds				Real Time Clock Registers
01	0	10 Minutes			Single Minutes				
02	0	12/24	10 h A/P	10 h	Single Hours				
03	0	0	0	0	0	Day Of Week			
04	0	0	10 Date		Single Date				
05	0	0	0	10 m.	Single Months				
06	10 Years				Single Years				
07	MS	10 Seconds Alarm			Single Seconds Alarm				Real Time Clock Alarm
08	MM	10 Minutes Alarm			Single Minutes Alarm				
09	MH	12/24	10 ha. A/P	10 h. alm.	Single Hours Alarm				
0A	MD	0	0	0	0	Day Of Week Alarm			
0B	Low Temperature Threshold								Temperature Alarm
0C	High Temperature Threshold								
0D	Number Of Minutes Between Temperature Conversions								Sample Rate
0E	$\overline{EOSC}$	CLR	0	SE	RO	TLIE	THIE	AIE	Control
0F	(reads 00h)								Reserved
10	(reads 00h)								Reserved
11	Current Temperature								Temperature
12	Start Delay Register (LSB)								Start Delay
13	Start Delay Register (MSB)								Start Delay
14	TR	MEM CLR	MIP	SIP	LOBAT	TLF	THF	ALMF	Status
15	Minutes								Start Time Stamp
16	Hours								
17	Date								
18	Month								
19	Year								
1A	Low Byte								Current Samples Counter
1B	Medium Byte								
1C	High Byte								
1D	Low Byte								Total Samples Counter
1E	Medium Byte								
1F	High Byte								
20–3F	(Read 00H)								Reserved

Fig. 4. The 1615 has 192 pages of internal memory. Each page consists of 32 memory cells. Figure 4a) shows the structure of the memory, while b) shows functions of memory cells in the first memory page – i.e. page 0.

sation. It is also year 2000 compatible.

The device can be programmed to start a temperature logging session via a push button, or under control of a host PC through the serial interface. After a session is started, the DS1615 can be made to wait for a pre-programmable delay ranging from 0 minutes to 45.5 days before taking its first temperature measurement.

After the first temperature measurement, the IC becomes idle. After a preset interval, it wakes up and measures temperature again. Up to 2048 temperature records can be stored in the on-board non-volatile memory. The device also produces a temperature histogram in 63 bins with 2.0°C resolution.

Control functions can also be carried out directly by the chip. It has programmable high and low-temperature trip points. If the measured temperature is outside the pre-set temperature range, the device can be made to flag an interrupt.

Communications with the host can carried out via a three-wire synchronous serial interface or via an asynchronous serial interface compatible with a standard UART. A 16-bit cyclic-redundancy check generator (CRC) is also provided to safeguard data read from the DS1615 in the UART communication mode.

Details of the DS1615 can be found in the manufacturer's data sheets mentioned later.

The 1615 in detail

Pin name and functions of the 1615 are described in **Table 1**.

Figure 4a) shows the internal memory organisation of the DS1615. Each location is eight bits wide. The complete memory locations are from 0000_{16} to $17FF_{16}$ – in total 6144 memory cells. These are segmented into 192 pages and each page has 32 memory cells.

Memory locations – also called registers – in page 0 are shown in Figure **4b**). Page 0 contains a real-time clock, control registers, a status register and other registers containing operation parameters.

Useful registers are the sample-rate register that determines the sampling rate; start-delay registers that determine a time delay in minutes before the first temperature reading is taken after a logging mission is started. Current temperature is stored in temperature register at address 11_{16}.

Page 2 is a user's non-volatile memory area where the user can write an identification string up to 32 characters long into the logger.

Temperature histogram bins start at page 64 and end at page 67 – in total four pages. Temperature records start from page 128 and use 64 pages.

Users can only write to some memory cells in page 0 and all memory cells in page 1. All other memory cells are read-only. The DS1615 provides a write-a-byte command and a read-a-page command. The former writes a single byte into a specific memory location. The latter allows a user to read a complete page, 32 bytes in one go.

Locations for temperature records. Memory at 0011_{16} stores the most-recently measured temperature. Locations in pages 128 to 191 store temperature records. The value in each 8-bit memory cell is converted into temperature using,

Temperature=(0.5×*byte value in a memory cell*–40)°C

Clock and calendar. Time and calendar information is accessed by reading/writing the appropriate RTC registers in page 0. The data is in binary-coded-decimal form. It is possible to set the clock for either 12-hour or 24-hour mode.

The DS1615 contains a time-of-day alarm whose registers are at 0007-$000A_{16}$. Bit 7 of each alarm register is a mask bit. When all mask bits are logic 0, an alarm occurs once per week when the values stored in time-keeping registers 0000_{16} to 0003_{16} match the values stored in the alarm registers.

Control and status. Location $000E_{16}$ holds the control register, whose bit functions are given in **Fig. 5a)**. It is used to set up the operating modes. Bit functions of the status register at 0014_{16} are given in Fig. **5b)**.

Serial interface. The DS1615 provides two types of serial interface: asynchronous and synchronous.

The asynchronous interface involves two data lines, TX and RX, and has a standard UART data format. Its bit rate is 9600 and it communicates via eight-bit word lengths terminated with one stop bit. Parity checking is not used.

The synchronous serial interface is a three-wire bus consisting of –RST, SCLK and I/O (data input/output). Since this design involves communicating with a PC's asynchronous COM port, the synchronous interface is

Msb

-EOSC	CLR	0	SE	RO	TLIE	THIE	AIM

-EOSC – 1: disable oscillator 0: enable oscillator

CLR – 1: enable memory clear 0: disable memory clear

SE – 1: a non-zero value in Sample Register. -ST pin held low 0.5 s to start logging
0: write a non-zero value to Sample Register to start logging

RO – 1: over write data from beginning if temperature memory is full
0: Stop logging data after the temperature memory is full

TLIE – 1: enable temperature low interrupt at -INT pin 0: Disable temperature low interrupt

THIE – 1: enable temperature high interrupt at -INT pin 0: Disable temperature high interrupt

AIM – 1: enable alarm interrupt at -INT pin 0: Disable alarm interrupt

(a) Bit functions of the control register

Msb

TR	MEM CLR	MIP	SIP	LOBAT	TLF	THF	ALMF

TR – 1: current temperature is ready in the Current Temp Register. 0: not ready

MEM CLR – 1: memories are cleared. 0: memory not cleared.

MIP – 1: logging mission is in progress. 0: logging mission is not in progress

SIP – 1: temperature sampling is in progress. 0: sampling is not in progress

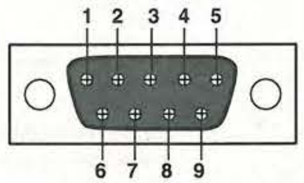
LOBAT – 1: low battery voltage. 0: good battery power

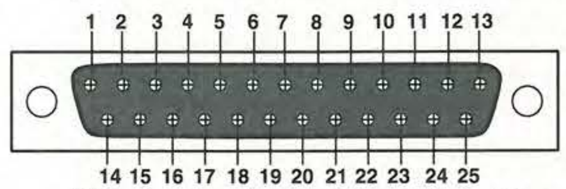
TLF – Temperature low flag

THF – Temperature high flag

ALMF – Alarm flag

(b) Bit functions of the status register

Fig. 5. Bit functions of the control and status registers in memory page 0.

(a) 9-pin male socket viewed from the back of the computer



(b) 25-pin male socket viewed from the back of the computer

Pin functions of the RS232 connectors

25 pin	9 pin	Name	Direction (for PCs)	Description
1		**–Prot**	-	Protective ground
2	3	**TD**	OUTPUT	Transmit data
3	2	**RD**	INPUT	Receive data
4	7	**RTS**	OUTPUT	Request to send
5	8	**CTS**	INPUT	Clear to send
6	6	**DSR**	INPUT	Data set ready
7	5	**GND**	-	Signal ground (common)
8	1	**DCD**	INPUT	Data carrier detedt
20	4	**DTR**	OUTPUT	Data terminal ready
22	9	**RI**	INPUT	Ring indicator
23		**DSRD**	I/O	Data signal rate detector

Fig. 6. Pin-out and functions of the RS232 port on a PC.

redundant. If you wanted the 1615 to talk to a microcontroller though, the synchronous interface may well be more appropriate.

Command sets. The DS1615 controller recognises five commands that are written into the chip via its serial interfaces, **Table 2**. To write a byte into the device, the host PC first sends 22_{16} to the controller via the RS232 port. An address byte is sent next. Finally, the host PC sends a data byte.

To read a page, the host PC sends 33_{16} to the controller. Then most and least-significant addresses of the start of a page are sent. After the DS1615 receives these two address bytes, it sends back 32 data bytes from the selected page plus cyclic redundancy check bytes. Bytes are sent continuously, i.e. without handshaking.

To read current temperature, the host PC sends 55_{16} first. If the logger is not in data logging mode, i.e. MIP=0, temperature is measured and the value is stored in the temperature register in page 0.

Since a temperature measurement takes time, it is advisable to poll the temperature-ready bit, TR, of the status register. This bit being logic high indicates that a temperature measurement is completed and the value is written into the temperature register. If the controller is in the data logging mode though, i.e. MIP=1, the command is simply ignored.

To clear on-board memory, the host first sends 64_{16} to the control register, sending the CLR bit high. Then the host sends the 'clear' command, $A5_{16}$, to the controller. After the memory is cleared, the MEM CLR bit in the status register goes high. Only after the clear command is completed can the sample-rate register be written with a non-zero value to start another logging mission.

RS232 transceiver unit

Pin designations of the PC's RS232 port are given in **Fig. 6**. In this design, some of the port's lines are used

Table 1. Pin functions of the DS1615 temperature recorder.

Pin	Mnemonic	Description
1	V_{bat}	Battery input for a 3V lithium cell. When $V_{cc}>V_{bat}$, the chip is powered by V_{cc}. When $V_{bat}>V_{cc}$, it is powered by V_{bat} Note that the serial interface is always powered by Vcc.
2	X_1	Connection to a 32.768kHz crystal with 6pF internal load capacitance.
3	X2	
4	NC	Not used
5	–INSPEC	Together with Pin 6, signals status of the operation of DS1615.
6	–OUTSPEC	See pin 5.
7	–INT	Open-drain output, active low, stays while a status bit in the status register signals an interrupt and the corresponding interrupt-enable bit in the control register is set.
8	GND	Ground
9	–ST	Start/status button input. When enabled as the data-logging start control, this pin tells the DS1615 to begin recording temperature at a regular and pre-set interval. It must be held low for 0.5s. It is also used to poll the status of logged data on pins 5 and 6.
10	–RST	Reset line of 3-wire synchronous interface.
11	I/O	Data input/output line in the 3-wire synchronous interface.
12	SCLK	Clock line in the 3-wire interface.
13	TX	Transmit line in the asynchronous interface.
14	RX	Receive line in the asynchronous interface.
15	COMSEL	Select Communication Mode. When high, 3-wire interface is selected. When low, asynchronous interface is selected.
16	V_{cc}	+5V power supply.

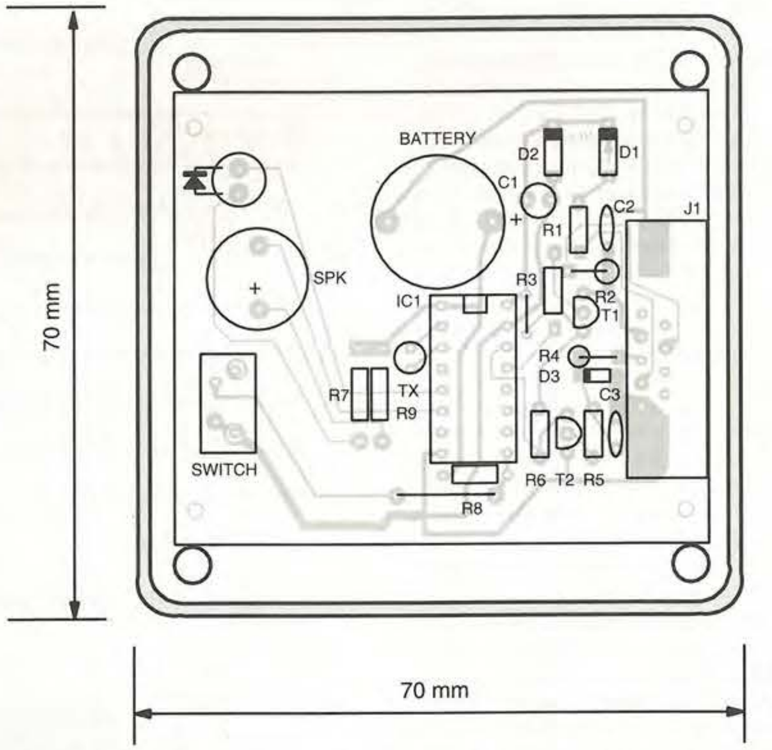


Fig. 7. The single-chip temperature logger can be constructed on a single-sided board and housed in a vented box. This figure shows the component layout of the PCB.

Table 2. The DS1615 recognises five commands, written into the chip via one of its two serial interface.

Command	Function	Description
22_{16}	Write byte	Write a single byte into a memory cell in pages 0 and 2
33_{16}	Read page	Read 32 memory cells in a page in one go
44_{16}	Specification test	Poll temperature extremes
55_{16}	Read temperature	Measure temperature immediately and store it in the current-temperature register in page 0. Mission-in-progress bit MIP in the status register should be logic low.
$A5_{16}$	Clear memory	Clears temperature record, histogram, temperature alarm, current samples, start time stamp, start delay and sample rate registers. Clear-enable bit CLR in the control register should be logic high.

List 1. Main program for the temperature logger in Visual Basic 5.

```
Function Read_page(Address_MSB As Byte, Address_LSB As Byte)
Dim outbyte(10) As Byte
outbyte(0) = 51 'read page command
MSComm1.Output = outbyte
outbyte(0) = Address_MSB
MSComm1.Output = outbyte
outbyte(0) = Address_LSB
MSComm1.Output = outbyte
End Function
Function Load_Data(address As Byte, Data As Byte)
Dim outbyte(10) As Byte
outbyte(0) = 34 'write a single byte into memory command
MSComm1.Output = outbyte
outbyte(0) = address
MSComm1.Output = outbyte
outbyte(0) = Data
MSComm1.Output = outbyte
End Function
Function Clear()
Dim outbyte(10) As Byte
outbyte(0) = 165 ' clear memory of logger
MSComm1.Output = outbyte
End Function
Function BCD(ByVal BIN As Integer) As Integer
    BCD = (BIN \ 10) * 16 + BIN Mod 10 'convert a binary byte into
            BCD
End Function
Function BIN(ByVal BCD As Integer) As Integer
    BIN = ((BCD And 240) / 16) * 10 + (BCD And 15) 'convert a BCD
            data into a binary byte
End Function
Private Sub Command1_Click()
    'erase memory, set Real Time Clock, set the start method as push
            button start
    Sleep (1000)
    Scan_interval = Val(Text1.Text)
    dummy = Load_Data(14, 64) 'set CLR bit in the Contro Register
    dummy = Clear 'clear memory command
    Sleep (10) 'delay a short while
    dummy = Load_Data(14, 16) 'SE bit=1, press Start button to start
            logging mission
    dummy = Load_Data(0, BCD(Second(Time))) 'Real Time Clock, 6 bytes
    dummy = Load_Data(1, BCD(Minute(Time)))
    dummy = Load_Data(2, BCD(Hour(Time)))
    dummy = Load_Data(3, BCD(WeekDay(Date)))
    dummy = Load_Data(4, BCD(Day(Date)))
    dummy = Load_Data(5, BCD(Month(Date)))
    dummy = Load_Data(6, BCD(Year(Date) - 1900))
    dummy = Load_Data(7, 0) 'Real time clock alarm, second
    dummy = Load_Data(8, 0) 'Real time clock alarm, minute
    dummy = Load_Data(9, 0) 'Real time clock alarm, hour
    dummy = Load_Data(10, 0) 'Real time clock alarm, day of week
    dummy = Load_Data(11, 0) 'low temperature limit
    dummy = Load_Data(12, 0) 'upper temperature limit
    dummy = Load_Data(18, 0) 'Delay number MSB=0, no delay
    dummy = Load_Data(19, 0) 'Delay number LSB=0, no delay
    dummy = Load_Data(13, Scan_interval) 'load scan interval in
            DS1615
    DoEvents
    Label2.Caption = "Initialisation completed. Logger can be
            disconnected for PC. Press Start Button to begin
            temperature logging"
End Sub
Private Sub Command2_Click()
Dim Scan_number As Single
Dim i As Integer, j As Integer, l As Integer
    Command2.Enabled = False
    Command1.Enabled = False
    MSComm1.InBufferCount = 0
    'read 1st page
    dummy = Read_page(0, 0) 'read page
    Sleep (50)
    DoEvents
    Logger_second = BIN(Data(1))
    Logger_minute = BIN(Data(2))
    Logger_hour = BIN(Data(3))
    Logger_date = BIN(Data(5))
    Logger_month = BIN(Data(6))
    Logger_year = BIN(Data(7))
    Scan_interval = Data(14)
    Start_minute = BIN(Data(21 + 1))
    Start_hour = BIN(Data(22 + 1))
    Start_date = BIN(Data(23 + 1))
    Start_month = BIN(Data(24 + 1))
    Start_year = BIN(Data(25 + 1))
    Scan_number = Data(29) * 256# * 256# + Data(28) * 256# + Data(27)
    If Scan_number = 2048 Then Scan_number = 2048
    DoEvents
    Label2.Caption = "Logging mission starts from: " +
            Str(Start_hour) + ":" + Str(Start_minute) + "   " +
            Str(Start_year) + "-" + Str(Start_month) + "-" +
            Str(Start_date) + Chr$(13) + "Number of temperature
            records stored in memory: " + Str(Scan_number)
    'read histogram data
    For i = 1 To 4
    dummy = Read_page(8, 32 * (i - 1))
    Sleep (50)
    DoEvents
       For j = 1 To 16
       hist(16 * (i - 1) + j) = 256# * Data(2 * j) + Data(2 * j - 1)
    Next j
    Next i
    'read temperature records
    For i = 1 To 8
    For j = 1 To 8
    dummy = Read_page(15 + i, 32 * (j - 1))
    Sleep (50)
    DoEvents
    For l = 1 To 32
    Temp((i - 1) * 256 + (j - 1) * 32 + l) = 0.5 * Data(l) - 40
    Next l
    DoEvents
    Label1.Caption = Str((i - 1) * 8 + j)
    Next j
    Next i
    'save data onto hard disk
    If Text2.Text = "" Then Text2.Text = "c:\Temp" & Day(Date) & "_"
            & Month(Date)
    filename = Text2.Text & ".DAT" 'Text2.text is the filename
    Open filename For Output As 1
    Print #1, "Data is downloaded at (PC time)  "; Time; "   "; Date
    Print #1, "Logger local time                                  ";
            Logger_second; ":"; Logger_minute; ":"; Logger_hour; "
            "; Logger_date; "-"; Logger_month; "-"; Logger_year
    Print #1, "Scan Interval: "
    Print #1, Scan_interval
    Print #1, "Logging start time stamp (minute hour day month year)"
    Print #1, Start_minute; Start_hour; Start_date; Start_month;
            Start_year
    Print #1, "Total number of temperature records (2048 maximum)"
    Print #1, Scan_number
    'save temperature log data
    Print #1, "Temperature records (Log number, temperature in °C)"
    For i = 1 To Scan_number
    Print #1, i; Temp(i)
    Next i
    'save histogram data
    Print #1, "Histogram data (temperature bin, number of temperature
            scans in the bin)"
    For i = 1 To 64
    Print #1, -40# + (i - 1) * 2#; hist(i)
    Next i
    Close #1
    'Label2.Caption = "Data download and saved into a file"
    Command2.Enabled = True
    Command1.Enabled = True
End Sub
Private Sub Command3_Click()
End
End Sub
Private Sub Command6_Click()
Dim outbyte(10) As Byte
outbyte(0) = 85 ' 55h is to read current temp
MSComm1.Output = outbyte
Sleep (1000)
dummy = Read_page(0, 0)
Label3.Caption = Format(0.5 * Data(18) - 40#, "00.0") & "°C"
End Sub
Private Sub Form_Load()
Com_number = 0
Do
Com_number = (InputBox$("Input 1,2,3 or 4 to select" & Chr(13) &
            "COM1, COM2,COM3 or COM4", "Select a COM port"))
If Com_number = "" Then End
Com_number = Val(Com_number)
Loop Until Com_number <> 0
MSComm1.CommPort = Com_number 'Assign a COM port
MSComm1.InBufferSize = 35
MSComm1.DTREnable = True 'DRT goes high
MSComm1.OutBufferSize = 1 'output buffer size = 1
MSComm1.InputMode = comInputModeBinary ' data format: binary form
MSComm1.RThreshold = 32
MSComm1.InputLen = 1  'read one byte a time
MSComm1.PortOpen = True 'open the port
End Sub
Private Sub MSComm1_OnComm()
'if 32 byte received, read data from the buffer
If MSComm1.CommEvent = comEvReceive Then
For i = 1 To 32
    Data(i) = AscB(MSComm1.Input) 'read data from input buffer and
            assign to DATA array
Next i
MSComm1.InBufferCount = 0
End If
End Sub
```

Program list in Module1.bas

```
Global Com_number As Variant, Scan_interval As Byte
Global Data(32) As Integer
Global Temp(2048) As Single
Global hist(64) As Single
Global Start_minute, Start_hour, Start_day, Start_month, Start_year
            As Integer
Global sel As Byte
Declare Sub Sleep Lib "kernel32" (ByVal dwMilliseconds As Long)
```

unconventionally. Transmit data line TD sends data from the PC to the logger. It is clamped by a zener diode, D_3, inverted by Tr_2 and then fed into the receive-data pin, RX, of the controller.

Fig. 8. Screen dump of the Visual Basic 5 demonstration software.

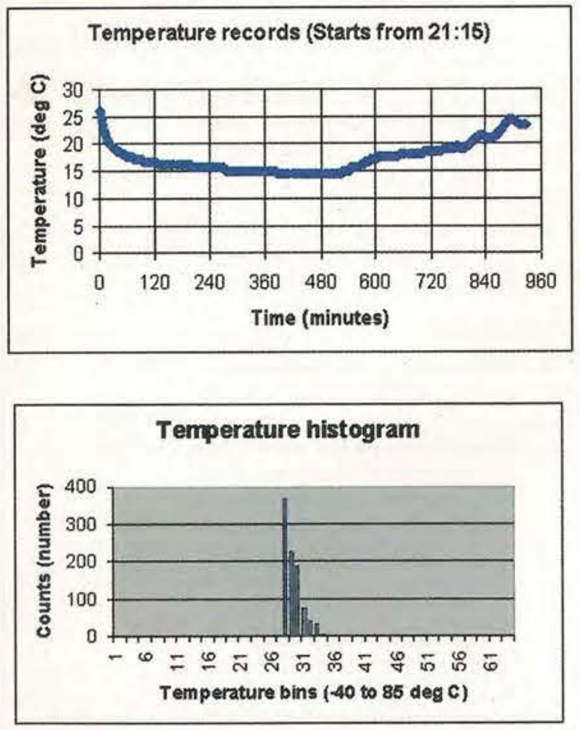


Fig. 9. Plots of temperature record data and histogram from a typical test. Both diagrams are produced using Excel.

Technical Support

A complete designer's kit is available from the author. It includes the above-mentioned VB5 software source code and EXE files. Make your enquiry to Dr.Pei An, 11 Sandpiper Drive, Stockport, SK3 8UL, UK. Tel/Fax: 44(0)1614779583. E-mail: Pan@intec-group.co.uk.

Data Sheets and application notes for the DS1615 are available from the Dallas web site: http://www.dallas.com.

Silver Birch Marketing Limited is one of the Dallas semiconductor's distributors in the UK, telephone 01480 812806.

Data from the logger is read by the receive data line of the COM port, namely RD. The TX pin of the controller is fed into a voltage translator circuit based on Tr_1 and the output from that circuit is fed into the RD line of the RS232 port.

The DTR line of the RS232 port supplies power to the logger. From Fig. 2 you can see that voltage at DTR is regulated to +5V by D_2 to generate a power supply for the DS1615. The RTS line of the port supplies –9V supply to the RS232 voltage translator.

The circuit can be constructed on a single-sided board as in Fig. 7. The circuit board can be housed in a vented ABS box, Fig. 1. The overall size of my implementation of the logger is 70mm by 70mm by 28mm.

Software control

Visual Basic 5 software demonstrating how the controller can read is shown in **List 1**. From **Fig. 8**, you can see that there are four control buttons on the screen: 'Initialisation" to initialise the logger, 'Download' to download data from the logger, 'Get temp' to read current temperature when the logger is not in a temperature logging mission, and 'Quit'.

A scanning interval from 1 to 255 minutes is keyed into the text box above the 'Initialisation' button. Before data downloading, a file name should be specified in the text box above the 'Download' button.

Visual Basic 5 provides a Microsoft COMM 5.0 control routine called 'MSComm', which is used to control a PC's COM port. The serial port should be configured for 9600 baud with 8-bit word length and one stop bit. All these parameters can be set in the 'Properties' section of the MScomm control routine. Other parameters are set within the program (see 'From_load' procedure in the program).

Initialising the logger

Initialising the logger is quite simple. It consists of the following stages.

- Enable memory clear – write 64_{10} into the control register to set the CLR bit.
- Issue a 'Clear' command.
- Select start method – start with push button or via the host PC. This is selected by setting the SE bit of the control register. In the present program, the push-button start is selected.
- Initialise real-time clock.
- Write various operation parameters into the data logger such as temperature limits, start delay, etc.
- Write a non-zero value to the sample-rate register.

If the logger is programmed to start using the start button, a logging mission will start as soon as the start button is pressed for more than 0.5 second. If the logger is programmed to start under the control of the host PC, after the sample rate is written into the register, a logging mission starts straight away. In both cases, once a logging mission is started, the on-board LED flashes four times.

An example of how the VB5 program initialises the logger is given in the subroutine 'Command1_Click()'. The logger is programmed to start logging once the start button is pressed. The start delay is set as 0. The first temperature measurement begins when the second register in the RTC rolls from 59 to 00. The delay between two scans can be selectable from 1 to 255 minutes.

Reading a page from the logger consists of the following stages.

- Issue a 'read-page' command.
- Send MSB address of the start of the page.
- Sending LSB address of the start of the page.
- Receive 32 data samples, plus CRC bits if appropriate, via the RS232 port.
- Read 32 data bytes, plus CRC bytes if appropriate, from the RS232 buffers into program variables.

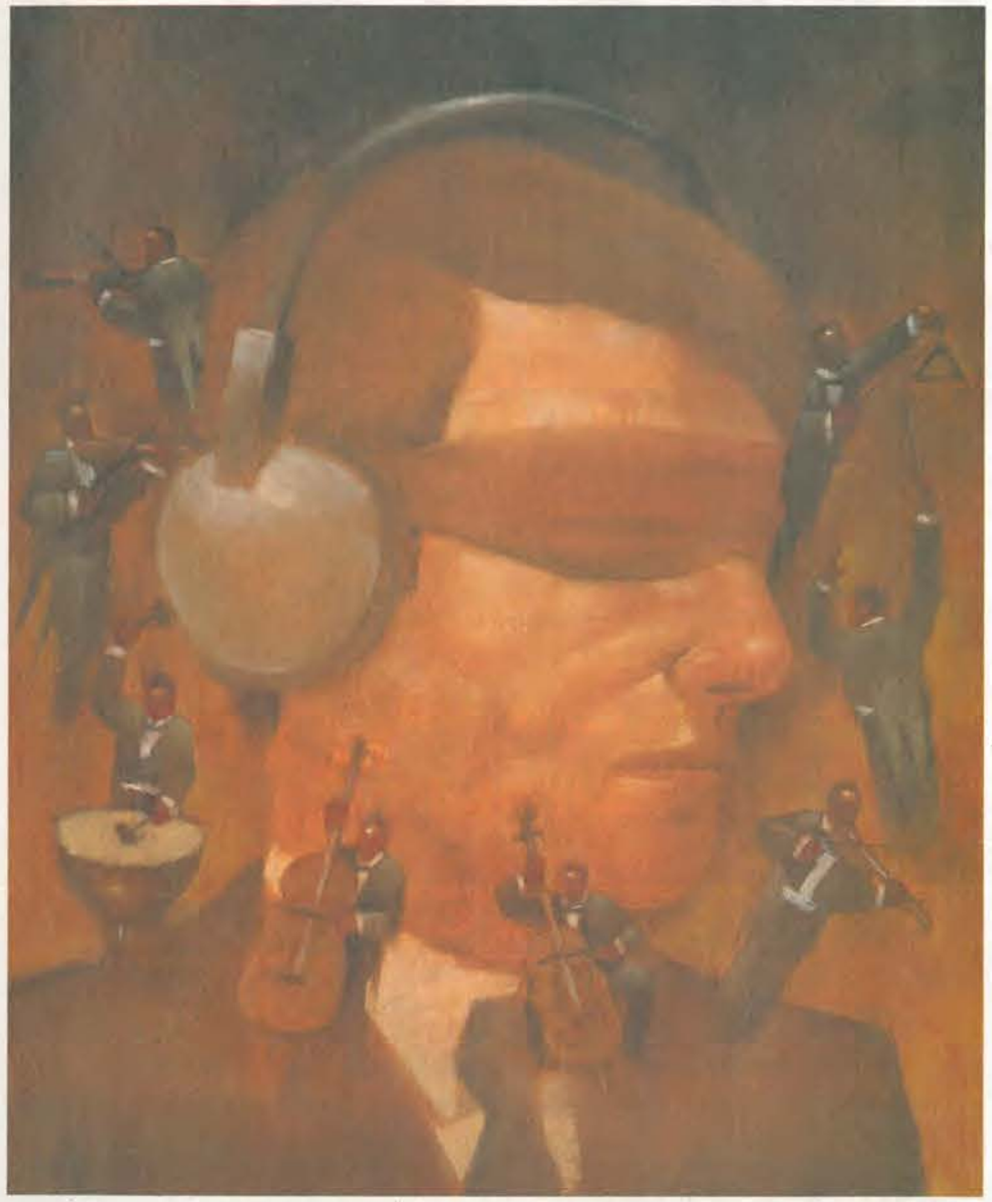
The subroutine 'Command2_Click()' illustrates how to read temperature records and histogram data from the logger. Reading data from the logger does not disturb the current temperature logging mission.

Temperature log data

After downloading data from the logger, the program saves the data into a file. The format of the file is DOS text format. Data can be imported into Microsoft Excel and other spreadsheet programs.

Figure 9 shows typical results of a temperature logging session. Figure **9a)** shows the temperature history as a function of time while Fig. **9b)** shows the temperature histogram. Both were generated by the Excel spreadsheet.

Finally, my thanks to Ian Mcateer from Silver Birch Marketing Limited, who supplied samples and technical support for this project. ■



Stereo from all angles IV

John Watkinson's previous article looked at how loudspeakers create the illusion of stereo. Now John shows how the requirements of that illusion define the role of the stereo microphone.

The creation of a convincing virtual sound source at some point between a pair of stereo loudspeakers is possible when the loudspeakers reproduce a pair of waveforms with sufficient accuracy in the frequency, time and spatial domains.

Prior to the shading compensator described in my last article, which must be considered as part of the loudspeaker system, must be pair of signals for each sound source there that are identical, except for a difference in amplitude which is a function of the direction.

A limitless number of these signals can be linearly added to produce the pair of waveforms fed to the speakers. Some of these signals will be directly from sound sources, others will be due to reflections, reverberation or ambience.

It follows from the above that the job of a stereo microphone is to produce, for each sound source, a pair of audio signals that have no phase or time differences but whose relative levels are a function of the direction of that source.

Again an arbitrary number of sources must be simultaneously handled by linear superimposition. My next article will discuss the additional constraints on distortion performance that linear superimposition in stereo causes.

Crossed-pair microphones

A widely used stereophonic technique involves the use of directional microphones that are coincidentally mounted to avoid timing errors between the signals but with their polar diagrams pointing in different directions. This configuration is known variously as a crossed pair or a coincident pair. **Figure 1a)** shows a stereo microphone constructed by crossing a pair of figure-of-eight microphones at 90°.

I discussed the actual requirement for the variation of relative amplitude with source position in my last article with the help of **Fig. 2**. You will see that the configuration of Fig. 1 approaches this requirement.

Output from the two microphones will be equal for a sound source straight ahead, because that source will be equally off-axis with respect to both microphones, assuming that their polar diagrams are identical. As the source moves left, it will move closer to the axis of the left-facing microphone and so the output will increase. At the same time it moves further from the axis of the right-facing microphone whose output reduces.

Consequently, this configuration appears to have the trend that Fig. 2 requires. Assuming that the microphones are mounted at 90° to one another, when a sound source has moved 45° off-axis, it will be in the response null of one of the microphones and so only one loudspeaker will emit sound. Thus the fully left or fully right reproduction condition is reached at ±45°.

The angle between nulls in L and R is called the acceptance angle, which has some parallels with the field of view of a camera.

Note that in most listening installations the loudspeakers will subtend an angle of more like 60° to the listener, and so a 90° sound stage is being shrunk into only 60°. This is technically incorrect, but in practice the error is relatively benign.

Loudspeaker directivity

In my experience, the better the directivity characteristics of the loudspeakers, the further apart they can be placed without the 'hole-in-the-middle' effect occurring. The closer the speakers approach 90° apart, the more the listener feels enveloped in the ambience of the recording and indeed the more important the inclusion of a shading compensator becomes.

In the crossed figure-of-eight microphone of Fig. 1, sounds between 45° and 135° will result in a pair of signals that are 'out of phase'. The main consequence of this is the cancellation of low frequencies because the woofers are in one another's near field. Recording engineers try to avoid placing dominant sound sources in this region. However, in the rest of the audio spectrum, when listening on speakers with good time domain response and directivity characteristics, virtual images can still be formed between the speakers. According to conventional explanations of intensity stereo this should not be possible and indeed on poor loudspeakers it isn't.

Note that the broadcaster has an understandable abhorrence of the anti-phase condition because the mono listener gets nothing! Sounds between 135° and 225° are back in-phase and are mapped onto the frontal stereo image. Fig. **1b)** shows how the 360° pickup of the microphone is mapped into the stereo image between the speakers. Note that the side zones fold into the centre along the ±45° axes. Some experience of origami is useful here.

The all-round pick-up of the crossed eight makes it particularly useful for classical music recording, where it will capture the ambience of the hall. Listening room reverberation will then provide sound from remaining directions.

Chaotic reverberation

Although this is obviously not accurately portrayed from a spatial standpoint, the chaotic nature of real reverberation means that different but still chaotic reverberation can still be plausible. Clearly what cannot be achieved with an intensity stereo system is the creation of a dominant sound source that is not between the speakers.

Other polar diagrams can be used: the crossed cardioid is popular for example. There is no obvious correct angle at which cardioids should be crossed, and the actual angle will depend on the application.

Commercially available stereo microphones are generally built on the side-fire principle, with one capsule vertically above the other. The two capsules can be independently rotated to any desired angle. Usually the polar diagrams of the two capsules can be changed.

While 'mid and side', or M-S, signal stereo can be obtained by using conventional L, R microphones and a sum and difference network, M-S signals can be obtained directly using a microphone having suitably oriented capsules. 'Mid and side' signals were discussed in my last article.

Figure 3 shows M-S microphones in which the S capsule musts always be a crossed eight. A variety of responses other than omnidirectional can be used for the M capsule. Note that in the M-S microphone, the fully left and fully right conditions are reached where the polar diagrams cross over, i.e where M=S and M=–S.

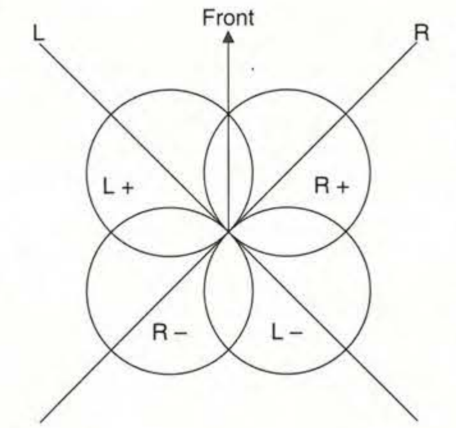
Benefits of the M-S microphone

The M-S microphone has a number of advantages. The narrowing polar diagram at high frequencies due to diffraction is less of a problem because the most prominent sound source will generally be in the centre of the stereo image and this is directly on the axis of the M capsule.

An image-width control can easily be built into an M-S microphone. Adjusting the gain applied to the S-signal changes the acceptance angle of the microphone. A favourite monophonic microphone can be turned into an M-S microphone simply by mounting a side facing crossed eight above it.

The sound-field microphone

In the sound-field microphone, four capsules are fitted in a tetrahedron. By



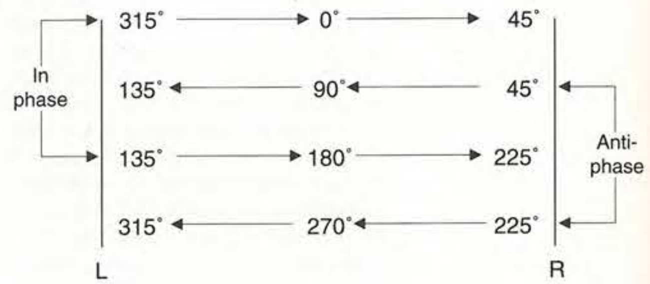


Fig. 1. Crossed-eight stereo microphone has two capsules at right angles, a), and crossed-eight microphone maps 360° sound stage into 90° as in b).

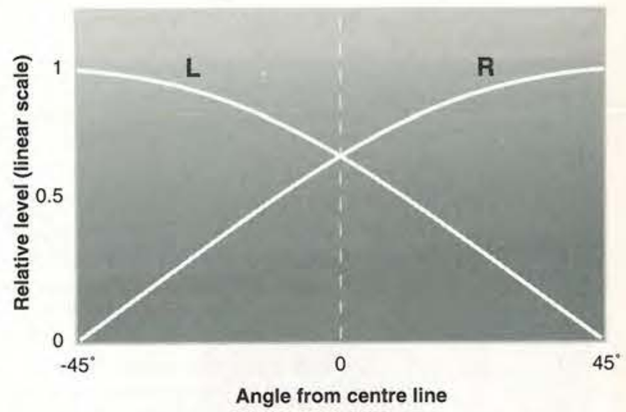


Fig. 2. Apparent position of the virtual sound source is a function of the level difference between the channels. This defines the characteristics that a microphone must have.

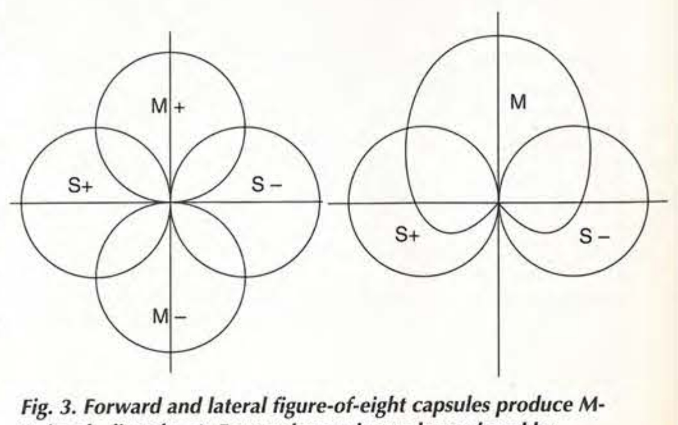


Fig. 3. Forward and lateral figure-of-eight capsules produce M-S signals directly, a). Forward capsule can be replaced by cardioid, as in b).

adding and subtracting proportions of these four signals in various ways it is possible to synthesise a stereo microphone having any acceptance angle and to point it in any direction relative to the body of the microphone. This can be done using the control box supplied with the microphone.

Although complex, the sound-field microphone has the advantage that it can be electrically steered and so no physical access is needed after it is slung. If all four outputs are recorded, the steering process can be performed in post production by connecting the control box to the recorder output.

The coincident microphone and the sound-field microphone both allow stereophonic reproduction in that the direction from which the reproduced sound appears to come parallels that of the original. However, these techniques require directional microphones.

Other microphone techniques will be found in stereo, such as the use of spaced omnidirectional microphones. Spaced microphones result in time of arrival differences and frequency-dependent phase shifts in the output signals. When reproduced by loudspeakers and both signals are heard by both ears the intensity stereo mechanism simply cannot operate.

Spaced omnidirectional microphones

With spaced omnidirectional microphones the result shown in **Fig. 4** is obtained.

Consider an off-axis source well to the left. When the source begins to emit sound, it will clearly be picked up by the left microphone first, and the listener will hear sound from the left speaker first.

However, as the sound continues, it will result in two microphone signals having a small amplitude difference but a large time difference. If initially it is assumed that the source frequency is such that the path length difference is an integer number of wavelengths there will be no phase difference. This will be reproduced by two speakers as a source slightly left of centre because of the small amplitude difference.

If the frequency is reduced slightly, the wavelength becomes longer and the right microphone signal will begin to lead in phase, moving the phantom image right. If the frequency is increased slightly, the wavelength shortens and the right microphone signal lags, moving the image left.

Consequently the apparent location of the initial transient is in disagreement with the location of the sustained sound which is itself highly frequency dependent. The result is that all sources having a harmonic content appear to be very wide.

Where the distance between microphones corresponds to an odd multiple of a half wavelength, the two microphone signals are out of phase. As both ears hear both loudspeakers, the result is a dip in frequency response and no image at all. This comb filtering effect is a serious drawback of spaced microphone techniques and makes monophonic compatibility questionable.

Not stereo, but spacious...

A central source will give identical signals that create a central phantom image. However, the slightest movement of the source off-axis results in time differences that pull the image well to the side. The resulting hole-in-the-middle effect is often counteracted by a third central microphone which is equally fed into the two channels.

This technique cannot be described as stereophonic reproduction because it is impossible for the listener to locate a particular source. Instead a spacious effect is the result.

As there is no scientific basis for the technique it is hardly surprising that there is no agreement on the disposition of the microphones. A great many spacings from a few centimetres to a few metres are used. One rule of thumb that has emerged is that the spacing should be no more than a third of the source width.

There is no science to support the creation of stereophonic images by spaced omni-directional microphones, but there certainly is some science to suggest why recording engineers would want to use omnidirectional microphones. Used singly, rather than in a stereophonic pair, a directional microphone sounds 'drier' than an omnidirectional microphone because it tends to collect only the direct sound to the detriment of the ambient or reverberant sound.

In monophonic working it was then logical to use an omnidirectional microphone wherever possible. This thinking then carried over into stereo, where, unfortunately it doesn't apply in the same way. This is because the microphones point in different directions and one can capture what the other doesn't.

Crossed eight versus omnidirectional

Figure 1b) shows that sounds from 360° are mapped into the output signals when a crossed-eight microphone is used. Consequently in theory and in practice a crossed-eight stereo microphone gives every bit as much ambience as an omnidirectional microphone, but with directional information as well.

It should be obvious that a crossed-cardioid microphone will sound drier than a crossed-eight, but the recording situation which demands a crossed cardioid would preclude the use of a n omnidirectional microphone.

Omnidirectional microphones may be audibly preferable to directional microphones because they can have better time-domain response. An omnidirectional microphone responds once to a pressure step or transient.

However, a directional microphone may obtain its directionality by using an acoustic labyrinth to delay the sound reaching the rear of the diaphragm. This gives the required directivity, but with the penalty that there is now a multiple path for transients to reach the diaphragm. Transients may thus be smeared, impairing sound quality.

Blumlein's shuffler

In his early experiments, Blumlein used omnidirectional microphones out of necessity, but he was aware that the signals could not be used directly. Instead a patented signal processor, known as a shuffler, was used, **Fig. 5**.

The shuffler converted phase differences in the input signals into the level differences required by the intensity stereo format. The two microphone signals fed a sum and difference circuit.

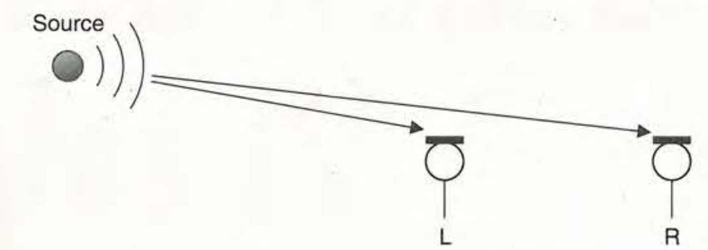


Fig. 4. Spaced microphones. Time of arrival due from initial transient conflicts with apparent position of sustained note.

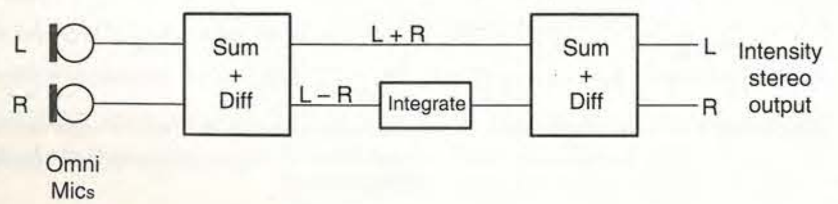


Fig. 5. Blumlein's shuffler allowed spaced omnidirectional microphones to produce intensity stereo outputs.

The difference signal would have an amplitude proportional to the angle of the source and to the frequency. Frequency dependence was removed with an integrator, prior to another sum and difference unit which produced the L–R intensity output.

The integrator gave a dynamic range problem in the signal processor. If the microphones are close enough to give a moderate phase shift at the highest audio frequencies, they will give a minute phase shift at the lowest frequencies.

Figure 6 shows the solution. Here, a linear array of omnidirectional microphones is used. The centre pair is used for the highest frequencies, the outermost pair for the lowest frequencies and the intermediate pair for the mid range. The dynamic range issue in each microphone pair is then solved and the three signal bands are combined by selective filtering.

Because they must allow the overall response to be minimum-phase, the band-combining filters are critical. In many respects the band combining filter is the inverse of a constant voltage loudspeaker crossover.

EMI tried to develop such a microphone after WW II as part of the Stereosonic system, but nothing emerged. In my view, the analogue technology of the day would not have allowed the necessary shuffler to have reached the required performance.

With modern electronics and capsule design, this is no longer the case and this approach deserves to be revisited with today's components. Incidentally the Stereosonic system was the first to identify the need for a shading compensator; a stage which is conspicuously absent from most of today's products. ■

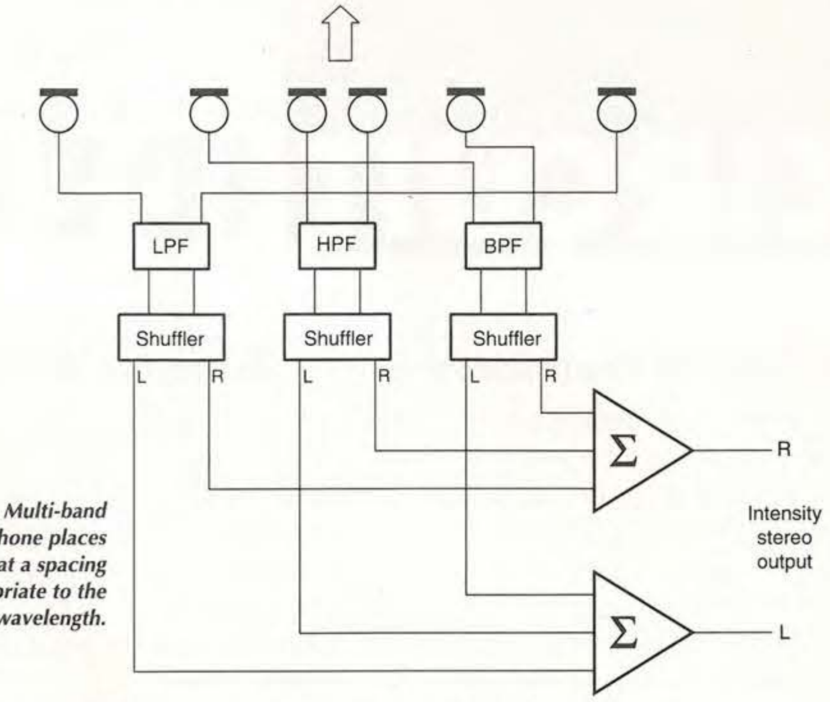
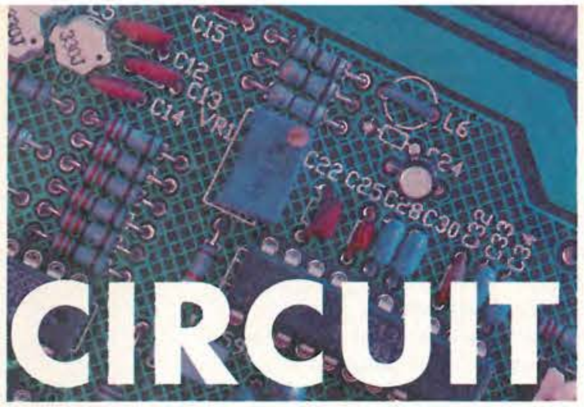


Fig. 6. Multi-band microphone places capsules at a spacing appropriate to the wavelength.

CIRCLE NO. 119 ON REPLY CARD



CIRCUIT IDEAS

Fact: most circuit ideas sent to *Electronics World* get published

The best circuit ideas are ones that save time or money, or stimulate the thought process. This includes the odd solution looking for a problem – provided it has a degree of ingenuity.

Your submissions are judged mainly on their originality and usefulness. Interesting modifications to existing circuits are strong contenders too – provided that you clearly acknowledge the circuit you have modified. Never send us anything that you believe has been published before though.

Don't forget to say why you think your idea is worthy.

Clear hand-written notes on paper are a minimum requirement: disks with separate drawing and text files in a popular form are best – but please label the disk clearly.

Lead-acid battery that can operate *in situ*

£75 winner

This float charger carries out a constant-current, constant-voltage charging cycle and is intended for use where the load is always connected, taking power from the mains via the charger or from the battery when mains are absent. The circuit shown takes a 6-cell battery of 35Ah capacity and peak charging current is 16A (C/2).

The core of the arrangement is a buck regulator with the switching element and freewheel diode arranged to make the positive side common, which enables the use of an n-channel fet operating in common-source mode; no level shifting is needed. It has the drawback that the battery negative is not the circuit negative.

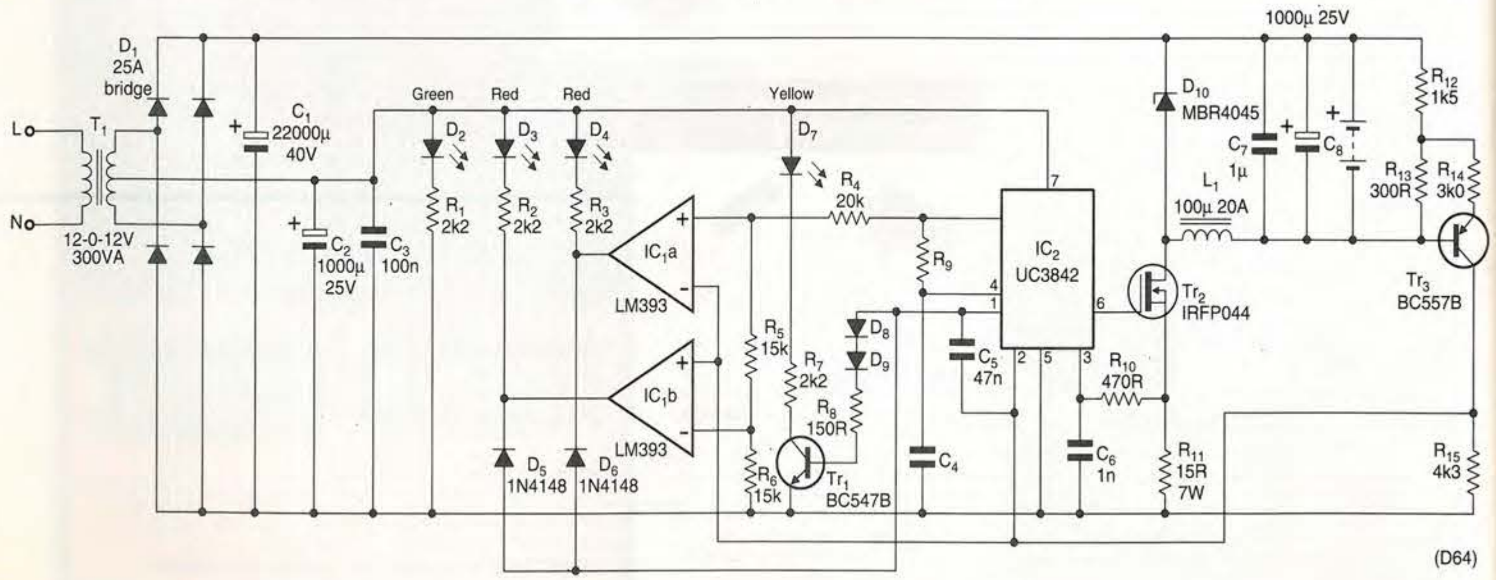
A *UC3842* regulator maintains 14.4V at 20°C across the battery. The voltage dropped by peak current through R_{11} is compared with error amplifier output on pin 1 of the regulator and the on time of the cycle adjusted. The components $D_{8,9}R_8$ and the base-emitter of Tr_1 clamp the error amplifier output so that the peak voltage across R_{11} does not exceed 0.25V to limit the charging current to 16A at switch-on with a discharged battery. In this constant-current part of the cycle, Tr_1 lights the yellow led to indicate the fact. Near full charge, error voltage falls and charge current decreases, so that the clamp circuit is inactive and the yellow led goes out.

Voltage feedback from the battery is divided by the number of cells in $R_{12,13,14}$ to give a mean cell voltage across R_{13}, the voltage across Tr_3 collector load being used as the feedback to the regulator. Advantages are that lead length is not a factor in battery-voltage sensing and the temperature coefficient of the transistor, which must be near the battery and away from power semiconductors, supplies temperature compensation.

The *LM393* op-amps constitute a fault detector, so that if battery voltage is outside the 10.2-20V area, one of the red leds illuminates and charge current is switched off.

Power for the circuit comes from the transformer centre tap to prevent the circuit draining battery current in the absence of mains power.

Ian Benton
Ilkeston
Derbyshire
D64



National Instruments sponsors Circuit Ideas

Over the next 12 months, National Instruments is awarding over £3500 worth of equipment for the best circuit ideas.

Once every two months for the next year, National Instruments is awarding an NI4050 digital multimeter worth over £500 each for the best circuit idea published over each two-month period. At the end of the 12 months, National is awarding a LabVIEW package worth over £700 to the best circuit idea of the year.* The first winner, selected from this issue or the following one, will be announced next month.

About National Instruments

National Instruments offers hundreds of software and hardware products for data acquisition and control, data analysis, and presentation. By utilising industry-standard computers, our virtual instrument products empower users in a wide variety of industries to easily automate their test, measurement, and industrial processes at a fraction of the cost of traditional approaches.

Software

Our company is best known for our innovative software products. The National Instruments charter is to offer a software solution for every application, ranging from very simple to very sophisticated. We also span the needs of users, from advanced research to development, production, and service. Our flagship LabVIEW product, with its revolutionary, patented graphical programming technology, continues to be an industry leader. Additional software products, such as LabWindows/CVI, ComponentWorks, Measure and VirtualBench, are chosen by users who prefer C programming, Visual Basic, Excel spreadsheets, and no programming at all, respectively.

Hardware

Our software products are complemented by our broad selection of hardware to connect computers to real-world signals and devices. We manufacture data acquisition hardware for portable, notebook, desktop, and industrial computers. These products, when combined with our software, can directly replace a wide variety of traditional instruments at a fraction of the cost. In 1996 we expanded our high-performance E Series product line in PCI, ISA and PCMCIA form factors, shipped our first VXI data acquisition products, and added remote (long-distance) capabilities to our SCXI signal conditioning and data acquisition product line.

Our virtual instrumentation vision keeps us at the forefront of computer and instrumentation technology. National Instruments staff works actively with industry to promote international technological standards such as IEEE 488, PCMCIA, PCI, VXI plug&play, Windows 95/NT, and the Internet. More importantly, we integrate these technologies into innovative new products for our users.

NI4050

The NI 4050 is a full-feature digital multimeter (DMM) for hand-held and notebook computers with a Type II PC Card (PCMCIA) slot. The NI 4050 features accurate $5^1/_2$ digit DC voltage, true-rms AC voltage, and resistance (ohms) measurements. Its size, weight, and low-power consumption make it ideal for portable measurements and data logging with hand-held and notebook computers.

- DC Measurements: 20mV to 250V DC; 20mA to 10A
- AC Measurements: 20mV rms to 250V rms; 20mA rms to 10Arms;
- True rms, 20Hz to 25kHz
- Up to 60 readings/s
- UL Listed
- $5^1/_2$ Digit Multimeter for PCMCIA

LabVIEW

LabVIEW is a highly productive graphical programming environment that combines easy-to-use graphical development with the flexibility of a powerful programming language. It offers an intuitive environment, tightly integrated with measurement hardware, for engineers and scientists to quickly produce solutions for data acquisition, data analysis, and data presentation.

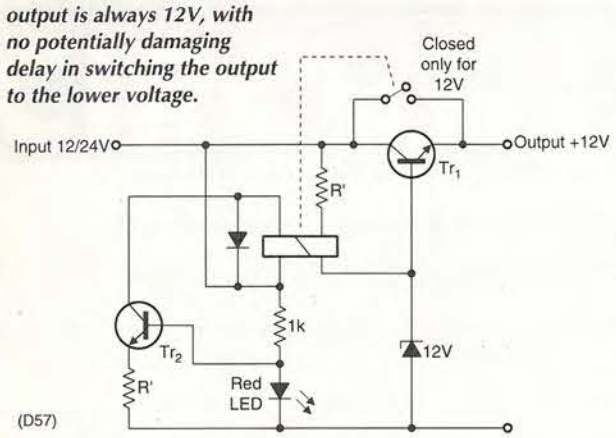
- Graphical programming development environment
- Rapid application development
- Seamless integration with DAQ, GPIB, RS-232, and VXI
- Full, open network connectivity
- Built-in display and file I/O

*All published circuit ideas that are not eligible for the prizes detailed here will earn their authors a minimum of £35 and up to £100. The first NI4050 will be awarded next month for the best idea from the December or January issue.

National Instruments – computer-based measurement and automation

National Instruments, 21 Kingfisher Court, Hambridge Road, Newbury, Berkshire, RG14 5SJ. Tel (01635 523545), Fax (01635) 524395 info.uk@ni.com www.ni.com.

Whether the input to this circuit is 12V or 24V, the output is always 12V, with no potentially damaging delay in switching the output to the lower voltage.



12V or 24V in – 12V out *all* the time

I detected a drawback in D M Bridgen's circuit in April 1999 on page 646. For a few milliseconds when the input reaches 24V, the output is 24V, regardless of whether the input is 12V or 24V. This delay until the relay opens could be long enough to damage some components connected to the output.

In my alternative, a double-coil relay prevents that possibility, at the expense of a few extra parts. The two coils are connected in opposition, so that with 24V input, the fields cancel and the relay opens to provide a 12V output.

With a 12V input, only one coil is energised via the Tr_2 constant-voltage source; the relay closes, connecting the 12V input to the output. The relay contact is therefore always open unless the input is 12V. The resistor R_x is equal to $1.3R/12$, where R is the coil resistance.

Jean-Marc Brassart
Saint-Laurent-du-Var
France
D57

Robust 5-30V, 10A power supply

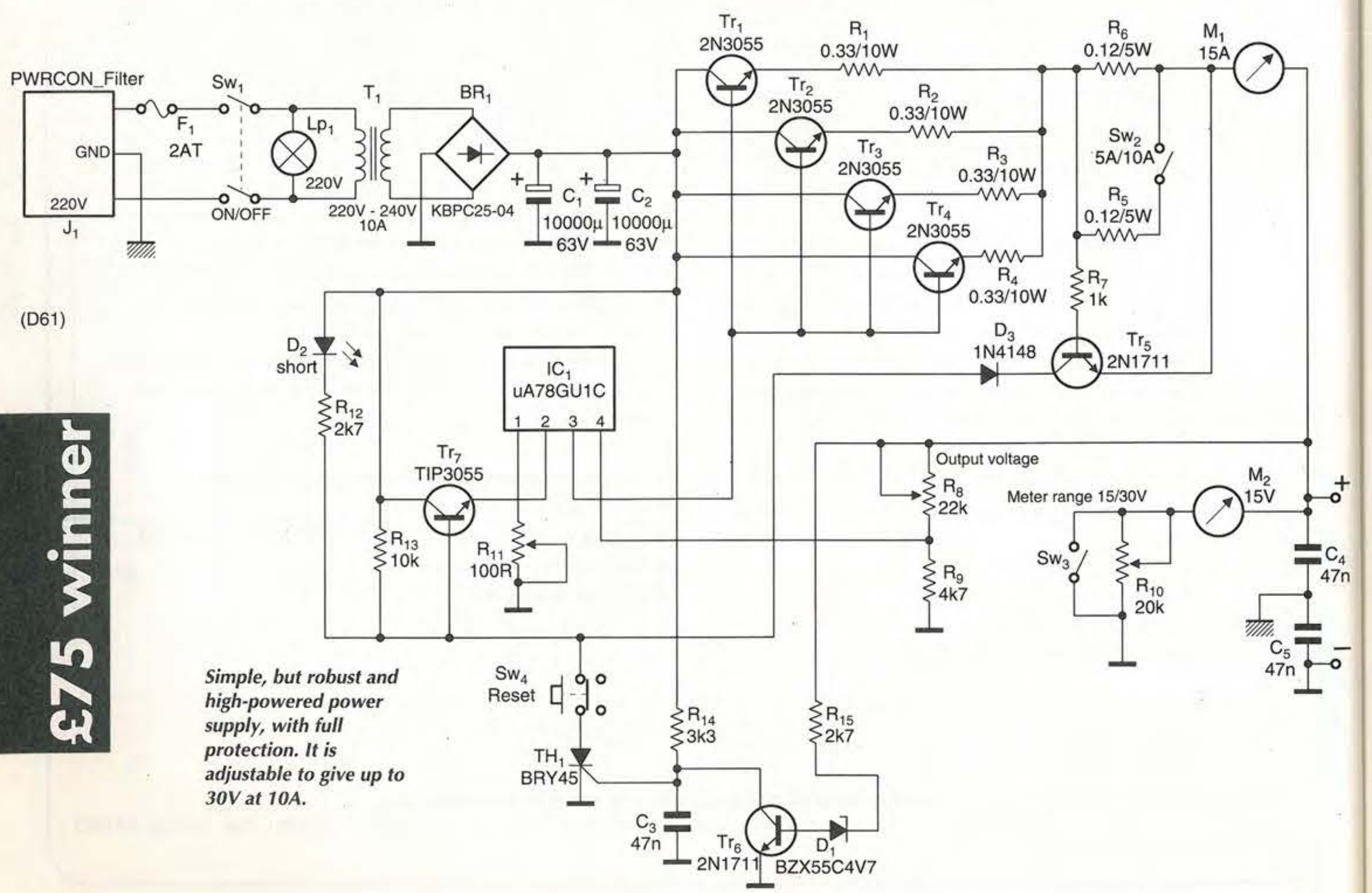
Although this design has been around for some time, it has given good service and never failed. It is protected against overload and short-circuits.

After the rectifier and reservoir capacitors, the four *3055s* are in parallel to take a load of up to 10A, the series resistors equalising the currents. Regulation is by the *μA78GU1C* feedback coming via $R_{8,9}$, R_8 being used to adjust the output and the minimum output set by R_{11}.

As regards protection, $Tr_{5,7}$ take care of current limiting. Resistors $R_{5,6}$ drop a voltage dependent on output current and cause Tr_5 to conduct at high current output, cutting off Tr_7 and therefore the regulator. Switch Sw_2 sets the current limit.

Short-circuit protection is by means of the thyristor, a drop in output voltage cutting off Tr_6 when it is lower than the zener voltage and triggering the thyristor to shut down the regulator. Reset is by Sw_4, which must also be pressed at switch on.

Jef Collin
Turnhout
Belgium
D61



Simple, but robust and high-powered power supply, with full protection. It is adjustable to give up to 30V at 10A.

£75 winner

Stepper motor drive has appeal for analogue *and* digital designers

A Microchip *PIC12C508* microcontroller generates pulses in the correct sequences and widths to control the speed and direction of a four-coil stepper motor.

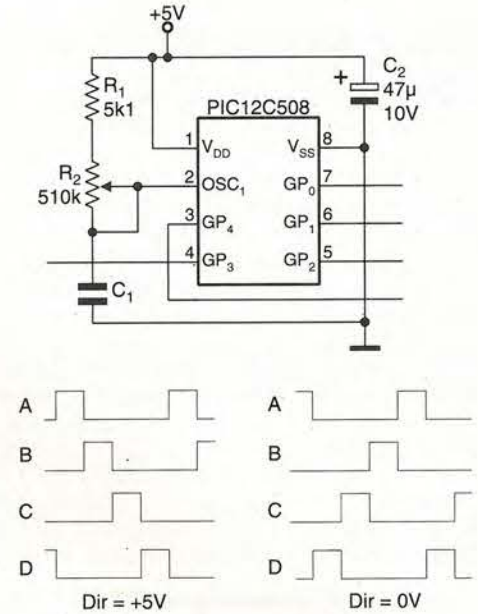
The controller outputs the four phases, the sequence being determined by the status of GP3. The built-in *RC* oscillator sets the pulse width and therefore the speed, which is controlled by $R_{1,2}C_1$, the table indicating the result of varying R when C_1 is 100pF and 0.01μF.

Yongping Xia
Torrance
California
USA
D56

£75 winner

Control software, for generating phase A to phase D pulses via a the PIC12C508.

```
STATUS   equ     0x03
GPIO     equ     0x06
Z        equ     0x02
phase    equ     0x07
         org     0x0
main     clrf    GPIO         ;
         movlw   0x28         ;
         tris    GPIO         ;
         movlw   0xc0         ;
         option               ;
         clrf    phase        ;
                              ;
start                         ;
         btfsc   GPIO, 3      ;     1/2   1/2
         goto    forward      ;     2     3
         movf    phase, 0     ;     1     3
         btfss   STATUS, Z    ;     1/2   4/5
         goto    next_1       ;     2     6
         movlw   0x03         ;     1     6
         movwf   phase        ;     1     7
         nop                  ;     1     8
         nop                  ;     1     9
         goto    send_pulse   ;     2     11
next_1                        ;           6
         decf    phase, 1     ;     1     7
         nop                  ;     1     8
         nop                  ;     1     9
         goto    send_pulse   ;     2     11
forward                       ;           3
         movlw   0x03         ;     1     4
         subwf   phase, 0     ;     1     5
         btfsc   STATUS, Z    ;     1/2   6/7
         goto    next_2       ;     2     8
         incf    phase, 1     ;     1     8
         nop                  ;     1     9
         goto    send_pulse   ;     2     11
next_2                        ;           8
         clrf    phase        ;     1     9
         goto    send_pulse   ;     2     11
send_pulse                    ;           11
         movf    phase, 0     ;     1     12
         call    find_phase   ;     5     17
         movwf   GPIO         ;     1     18
         goto    start        ;     2     20
find_phase                    ;
         addwf   PCL, 1       ;
         retlw   0x01         ;     phase A
         retlw   0x02         ;     phase B
         retlw   0x04         ;     phase C
         retlw   0x10         ;     phase D
                              ;
         end                  ;
```



Driving a stepper motor from the outputs of a PIC microcontroller, which varies speed and direction.

The first National Instruments digital multimeter winner, to be selected from the December 1999 and January 2000 issues, will be announced next month.

These values determine the stepper motor microcontroller's pulse widths.

R_1+R_2	Pulse width C_1=100pF	C_1=10nF
5.1k	72μs	5.1ms
10k	130μs	9.4ms
20k	260μs	19ms
51k	630μs	47ms
100k	1.25ms	90ms
200k	2.5ms	180ms
510k	6.3ms	470ms

£100 winner

Track your products with an electronic serial number – easily and at negligible cost
Program a chip using this exceedingly simple PC-based one-chip programmer then embed the chip into your design. The 32-digit serial number within the chip can then be used to, say, track the procuct through manufacture, or to identify it later in its life.

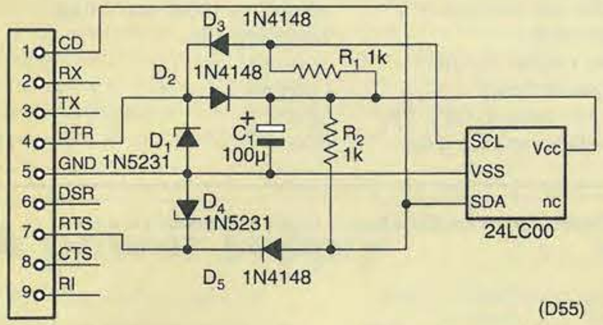
Easy programmer for 16byte eeproms

A 16byte eeprom by Microchip, the *24LC00*, is packaged in a 5-pin SOT-23 and provides up to 32 digits, used, for example, to generate electrically readable serial numbers.

Shown in the diagram is the interface between the pc and the eeprom, in which the two outputs on pins 4 and 7 generate an I^{2}C sequence to the 24LC00, pin 4 also providing power to the ic. Diodes $D_{1\text{-}4}$ clamp these outputs to 5V and $R_{1,2}$ are pull-ups for SCL/SDA lines.

The C program initially writes the 32-bit serial number at the start, the number increasing by one for each device written to.

Yongping Xia
Torrance
California
USA
D55



In devices needing electrically readable serial numbers, the 24LC00 16-byte eeprom is tiny and provides a 32-bit number. The circuit shown programs the eeprom from a pc's serial port.

Software for programming an EEPROM through a PC's COM port. This C program writes a serial number into the 24LC00. A 32-digit initial number is asked when the program is started. This number is programmed into the 24LC00 by hitting the return key. The serial number is incremented automatically for next device.

```
#include <stdio.h>
#include <conio.h>
#include <dos.h>
#include <bios.h>
#define MCR      4
#define MSR      6
#define SCL_LOW     0xfe
#define SCL_HIGH   0x01
#define SDA_LOW    0xfd
#define SDA_HIGH   0x02
int      i, j, id_num[32], out_data, port_add,
key_num, temp_data, read_ack,
         in_put,  read_data, find_err=0,
base_add1=0x3f8, base_add2=0x2f8;
void get_start_num(void)            /* get an initial
ID number */
{
  printf("Start ID Number:\n");
  for (i=0; i<32; i++)
   {
     in_put=getche();
     if (in_put>0x39 || in_put<0x30)
       i=32;
     else
      {
        for (j=i; j>0; j-)
          id_num[j]=id_num[j-1];
        id_num[0]=in_put & 0x0f;
      }
   }
}
void send_control(int control_data)      /* send 1
byte to device */
{
  for (j=0; j<8; j++)
   {
     if ((control_data&0x80)!=0)
        out_data|=SDA_HIGH;
     else
        out_data&=SDA_LOW;
        outportb(base_add1+MCR, out_data);
        delay(1);
        out_data|=SCL_HIGH;
        outportb(base_add1+MCR, out_data);
        delay(5);
        control_data*=2;
        out_data&=SCL_LOW;
        outportb(base_add1+MCR, out_data);
        delay(1);
   }
  out_data|=SDA_HIGH;
  outportb(base_add1+MCR, out_data);
  delay(1);
}
int get_ack(void)                              /* get
acknowledge */
{
  out_data|=SCL_HIGH;
  outportb(base_add1+MCR, out_data);
  delay(5);
  read_ack=inportb(base_add1+MSR)&0x80;
  out_data&=SCL_LOW;
  outportb(base_add1+MCR, out_data);
  delay(1);
  if (read_ack!=0)
    return 1;
  else
    return 0;
}
int write_ic(void)
{
  out_data=0x03;
  for (i=0; i<16; i++)
   {
      out_data&=SDA_LOW;
      outportb(base_add1+MCR, out_data);
      delay(1);
      out_data&=SCL_LOW;
      outportb(base_add1+MCR, out_data);/* start */
      delay(1);
      send_control(0xa0);
      if (get_ack()!=0)
         return 1;
      send_control(i);                    /* send address
*/
      read_ack=inportb(base_add1+MSR)&0x80;
      if (get_ack()!=0)
         return 1;
      temp_data=id_num[i*2+1]*16+id_num[i*2];
      send_control(temp_data);               /* send data
*/
      if (get_ack()!=0)
         return 1;
      out_data&=SCL_LOW;
      outportb(base_add1+MCR, out_data);
      delay(1);
      out_data&=SDA_LOW;
      outportb(base_add1+MCR, out_data);
      delay(1);
      out_data|=SCL_HIGH;
      outportb(base_add1+MCR, out_data);
      delay(5);
      out_data|=SDA_HIGH;
      outportb(base_add1+MCR, out_data);/* stop */
      delay(1);
   }
  return 0;
}
int verify_ic(void)
{
  for (i=0; i<16; i++)
   {
      out_data&=SDA_LOW;
      outportb(base_add1+MCR, out_data);
      delay(1);
      out_data&=SCL_LOW;
      outportb(base_add1+MCR, out_data);/* start */
      delay(1);
      send_control(0xa0);
      if (get_ack()!=0)
         return 1;
      send_control(i);                    /* send address
*/
      read_ack=inportb(base_add1+MSR)&0x80;
      if (get_ack()!=0)
         return 1;
      out_data&=SDA_LOW;
      outportb(base_add1+MCR, out_data);
      delay(1);
      out_data|=SCL_HIGH;
      outportb(base_add1+MCR, out_data);
      delay(5);
      out_data|=SDA_HIGH;
      outportb(base_add1+MCR, out_data);
      delay(1);
      out_data&=SDA_LOW;
      outportb(base_add1+MCR, out_data);
      delay(1);
      out_data&=SCL_LOW;
      outportb(base_add1+MCR, out_data);/* start */
      delay(1);
      send_control(0xa1);          /* send control */
      if (get_ack()!=0)
         return 1;
      read_data=0;
      for (j=0; j<8; j++)          /* read device */
       {
         out_data|=SCL_HIGH;
         outportb(base_add1+MCR, out_data);
         delay(5);
         read_data*=2;
         if ((inportb(base_add1+MSR)&0x80)!=0)
           read_data++;
         out_data&=SCL_LOW;
         outportb(base_add1+MCR, out_data);
         delay(1);
       }
      out_data|=SCL_HIGH;
      outportb(base_add1+MCR, out_data);
      delay(5);
      out_data&=SCL_LOW;
      outportb(base_add1+MCR, out_data);
      delay(1);
      if (read_data!=id_num[i*2+1]*16+id_num[i*2])
         return 1;
      out_data&=SDA_LOW;
      outportb(base_add1+MCR, out_data);
      delay(1);
      out_data|=SCL_HIGH;
      outportb(base_add1+MCR, out_data);
      delay(5);
      out_data|=SDA_HIGH;
      outportb(base_add1+MCR, out_data);/* stop */
      delay(1);
   }
  return 0;
}
void init(void)
{
  clrscr();
  for (i=0; i<32; i++)
    id_num[i]=0;
  outportb(base_add1+MCR, 0x00);
  delay(100);
}
void increase_id_num(void) /* increase ID number by 1
*/
{
  i=0;
  do{
      id_num[i]++;
      if (id_num[i]==10)
       {
          id_num[i]=0;
          i++;
       }
      else
          i=32;
    }while(i<32);
}
int main(void)
{
  init();
  get_start_num();
  clrscr();
  printf("Hit enter to program device, hit q to
quit\n");
  for (;;)
   {
     clrscr();
     printf("\n\n\n\nCurrent ID = ");
     for (i=0; i<32; i++)
       printf("%d",id_num[31-i]);
     if (find_err==1)
      {
        printf("\nThere is an error");
        find_err=0;
      }
     printf("\n\n\n\n\n\n\n\n\n\n\n\n\n\n\n\n\n");
     printf(" Push Enter to program the device.  Push Q
to quit");
     in_put=getch();
     if (in_put==0x51 || in_put==0x71)
      {
        clrscr();
        exit(1);
      }
     else
      {
        clrscr();
        printf("\n\n\n\nCurrent ID = ");
        for (i=0; i<32; i++)
          printf("%d",id_num[31-i]);
        printf("\nProgramming...");
        outportb(base_add1+MCR, 0x03);
        delay(400);
        if (write_ic()!=0)
           find_err=1;
        else
         {
           delay(1);
           if (verify_ic()==0)
             increase_id_num();
           else
             find_err=1;
         }
        outportb(base_add1+MCR, 0x00);
        delay(10);
      }
   }
}
```

£75 winner

Ssb transmission checker fits in a microphone housing

Two-tone generator for rapid checking of ssb amateur transmissions, small enough for mounting inside a microphone.

This two-tone generator assists in the adjustment of amateur ssb transmissions and is compact enough to allow its inclusion in the base of a table microphone.

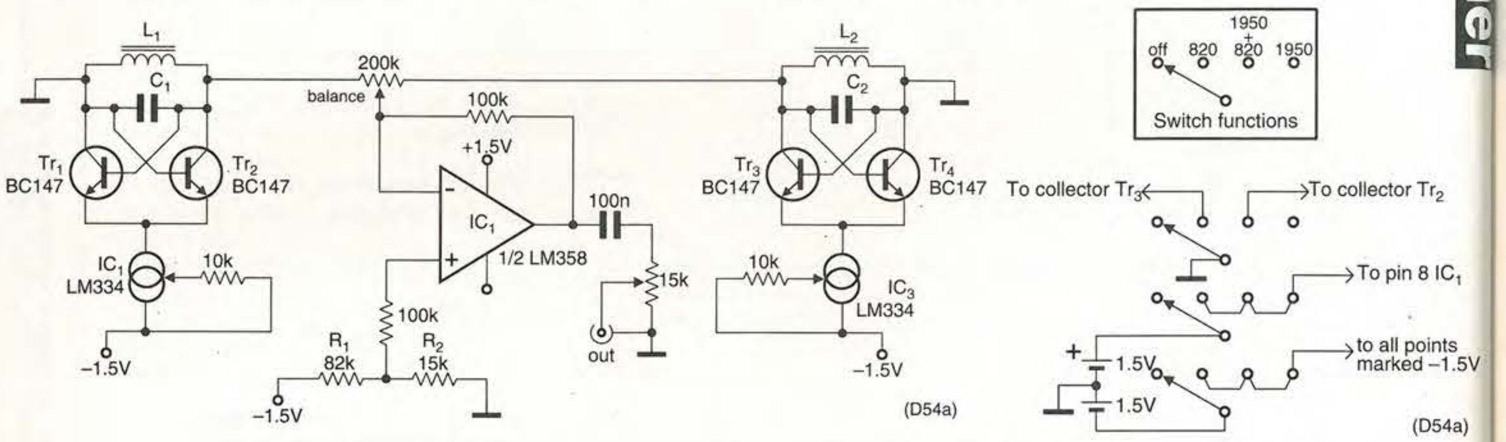
Two similar *LC* oscillators operate on unrelated frequencies: in this example, 820Hz and 1950Hz, the tuned circuits determining the frequencies. Balance is set to obtain equal amplitudes at the output of the *LM358*. The voltage rails are ±1.5V and $R_{1,2}$ set the bias.

Switching, shown separately for clarity, allows each or both oscillator outputs by shorting the unused oscillator, this allowing dc coupling between them.

Amplitude is within 1% with supply variation of 1.25-1.5V or, with a higher voltage rail, within 0.05% for a variation of 1.5-6V.

Transistors are *BC147* and the 'long tails' are *LM334* current sources. To guard against rfi, a shielded enclosure should be used.

P J Anderson
Sydney
Australia
D54



Efficient thermostat uses 50W bulbs to achieve ±1° temperature environment

Proportional thermostat controls a dc load to within 1°C.

This design is capable of running four 50W, though greatly underrun, lamps used as heaters, maintaining a temperature setting to within 1°.

The *ICM7555*, IC_2, is used as an oscillator running at around 30kHz with a 50% duty cycle and driving the complementary emitter follower. In turn, the follower drives the power fet to switch the load; one of the *555* devices that has a decent amount of drive capability might well manage without the emitter follower. The diode is only needed if the load is inductive.

An operational transconductance amplifier IC_1, a *CA3080E*, has a maximum current output dependent on the current into pin 5. Resistor R_5 has just under three times the resistance of the oscillator feedback resistor to ensure that it will be able to hold the 555 oscillator high or low to turn the output permanently on or off.

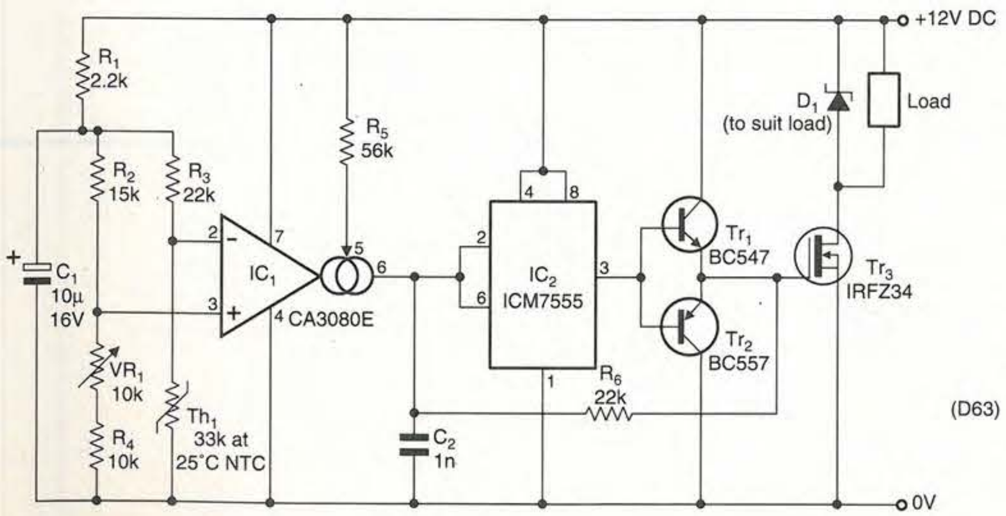
The variable resistor controls the set point between 25°C and 40°C.

When the temperature is low, the *CA3080E* output is low, the *555* output high and the fet on. High temperatures reverse this and the fet goes off. Between the two extremes, the *3080* input voltages are about equal and the *555* oscillates, its duty cycle being controlled by the *3080* output current. There is therefore a proportional control band to minimise overshoot.

Gain of the *3080* is quite low – an advantage here, since too high a gain would give too narrow a proportional band. As shown, the band is about 1°C, though varying the 2.2kΩ resistor varies the bandwidth; it also prevents the differential input voltages exceeding 5V – not a good idea with the *3080*.

The only drawback to the circuit is that when the circuit leaves the proportional band the frequency drops and becomes audible as a whistle if the load vibrates.

Ian Benton
Ilkeston
Derbyshire
D63



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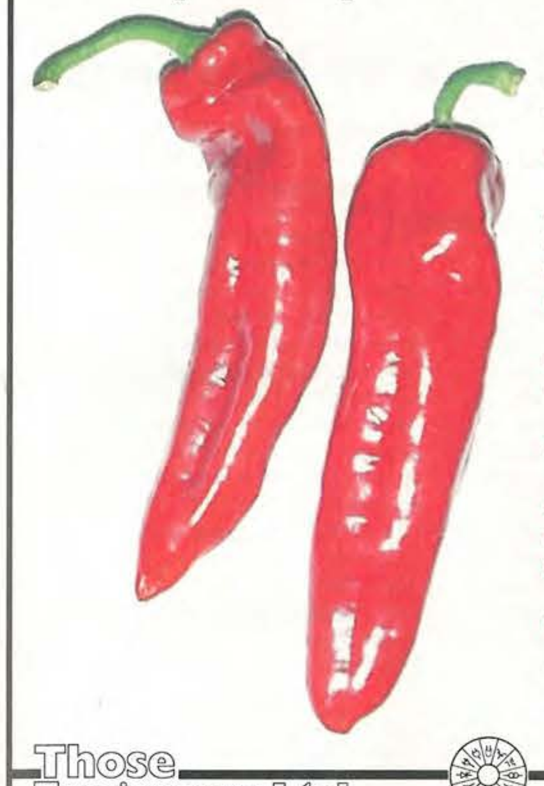

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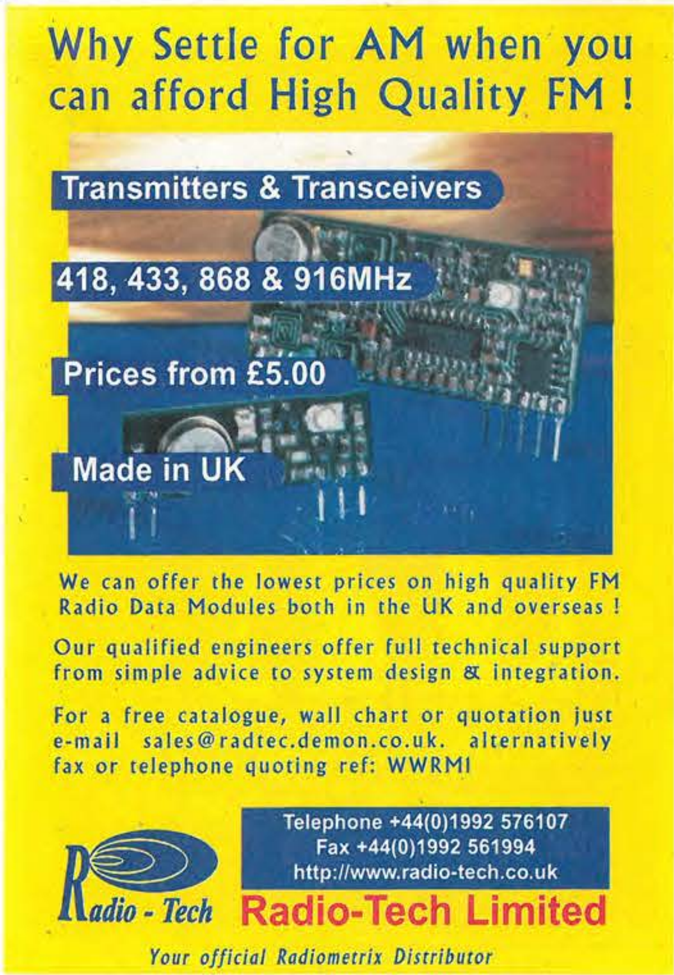

CIRCLE NO.121 ON REPLY CARD

CIRCLE NO.122 ON REPLY CARD

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How do I bring this signal back to a 0V base line?

Accepting a ±10V input, this circuit corrects a zero offset of up to ±5V, so that the output is an inverted replica of the input but referred to zero volts.

Comparator IC_1 detects input polarity and determines the polarity of the voltage applied to the digital potentiometer IC_2. If the comparator IC_{1a} output goes high, the resistor R_1 is connected to zero volts through the pot, so that the potentiometer input is around +5V; when the comparator is low, the pot input is taken to –5V by the comparator pull-down. Voltage at the potentiometer wiper is buffered and taken to IC_{3b}, where it counters the voltage from the input, its output being an inverted form of the circuit input. Since the comparators are 5V devices, attenuators reduce the voltage at their inputs, and positive feedback is applied to provide hysteresis.

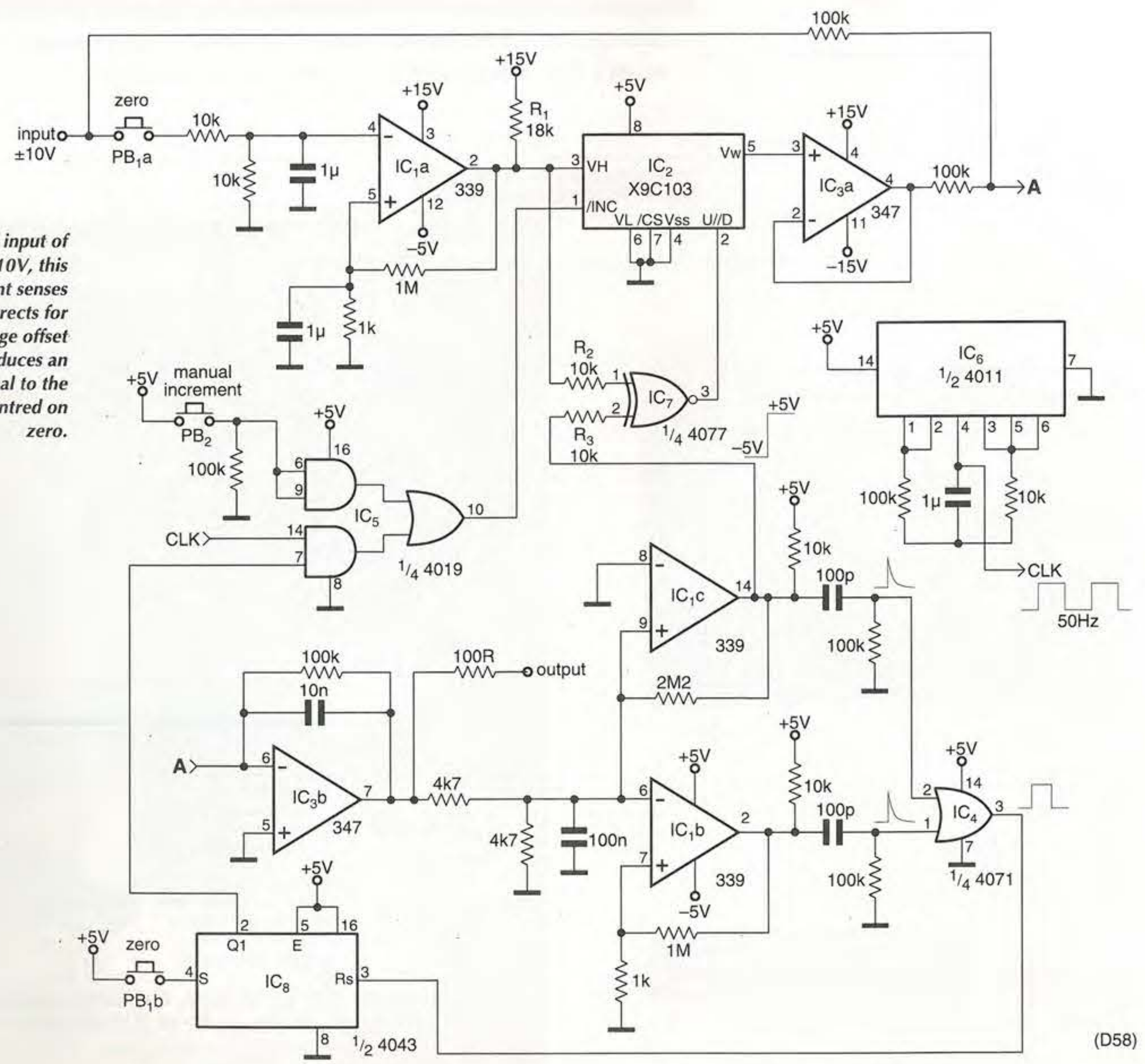
Momentarily pressing PB_1 applied the input to IC_{1a}, which senses the polarity and Q_1 of IC_8 goes high to allow clock pulses from IC_5 to increment the potentiometer IC_2, which ramps up or down depending on the control voltage from IC_7. The initial error, output of IC_{1c}, is compared with the polarity of the potentiometer input in IC_7, which selects the direction in which the potentiometer ramps to reduce the error. As the error passes zero, inputs 1 or 2 of IC_4, depending on the polarity of the initial error, cause positive-going spikes at its output. These are cleaned up in IC_4 and used to reset IC_8 to close IC_5 and stop clock pulses going to IC_2. The output is now zero ±50mV, the potentiometer having 100 steps maximum.

IC_6 is the 50Hz clock, which allows a zeroing time of 2s; a faster process could be arranged, but various noise-reducing capacitors have been selected for that purpose and would need to be reduced for a faster operation.

Switch PB_2 provides a manual increment. Dallas Semiconductors have the *DS1666* variable-sensitivity pot., which is compatible with the one shown here, the Xicor *X9C103*, provides higher resolution. There are also the *X9C303* logarithmic type, needing a variation in R_3, and the 128-step *AD5220*. No debouncing is needed on either switch.

Ian E Shepherd
H R Wallingford
Wallingford
Oxfordshire

Taking an input of up to ±10V, this arrangement senses polarity, corrects for a zero voltage offset and produces an output equal to the input, but centred on zero.



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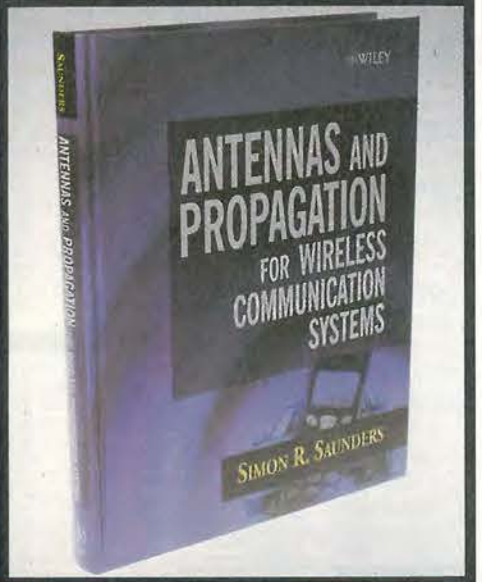

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NEW PRODUCTS

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Audio distortion meter

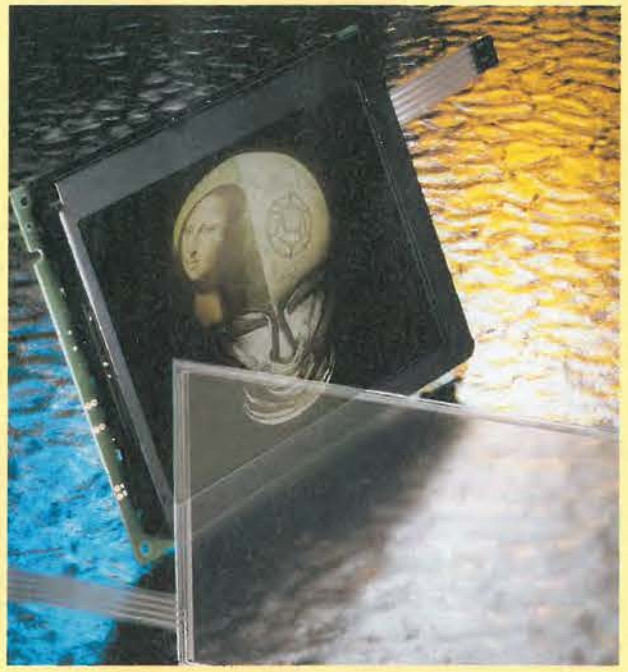
Vann Draper has introduced a Grundig Digimess millivolt meter. The MV100 is an audio frequency meter for measuring distortion, harmonic distortion and levels including rectification, RMS, peak, quasi peak, noise and interference voltages with a range of 5Hz to 88kHz. It does not provide a signal source. An adjustable time constant and bar graph is included and psophometric filters allow selective or wideband measurement. Weighting to linear CCIR 468-3 is standard.
Vann Draper Electronics
Tel: 01162 771400
Enquiry No 501

Prototyping board for FPGAs

Lockes Digital Developments has released the Protoflex-6K prototyping board based upon the Altera Flex 6000 family of programmable logic devices. The board provides a Flex 6000 quad-flat-pack device, onboard power conversion and two prototyping areas. One is a 0.254cm matrix area for conventional through hole devices and the other an uncommitted SOJ area for surface mount devices. Access to all device pins is possible via banks of header pins. VME, CompactPCI and stand-alone versions of this single Eurocard form-factor board can be supplied. The 1960 logic elements on the Flex EPF6024 give capacity for complex logic designs. Designs can be downloaded from a PC or blown into an onboard EPROM for automatic configuration at power-up. A customisation service is available. Work has begun on a card that uses Altera's Apex 20K devices.
Lockes Digital Developments
Tel: 01258 821222
Enquiry No 503

Touchscreen technique sputters onto glass

A technique for sputtering Indian Tin Oxide directly onto glass has let Densitron produce touchscreens that only require one layer of polyester and one layer of glass. Called DensiTouch, they use a single chip controller. They can be custom built and, using standard LCDs with integral touchscreens, parallax has been eliminated. They suit applications requiring 0.25 VGA, 128 x 240 and 64 x 240 displays.
Densitron
Tel: 01959 700100
Enquiry No 502



25/15W DC-DC converter

With UL1950 recognition to operational level, the NPH 25 and 15W DC/DC converters from Newport are for telecoms and industrial distributed power systems. Efficiency levels are up to 90 per cent and power densities up to 1.42W/cm³. The 25W units have footprints of 50 by 35mm and the 15W models 50 x 25mm. Height for both is 10mm.
Newport Components
Tel: 01908 615232
Enquiry No 504

Wireless test subsystem

Blue Wave Systems has combined the Comstruct PCI/C6600 DSP board, PMC/SB3410 and development tools to produce a subsystem for testing wireless equipment.
Newport Components
Tel: 01908 615232
Enquiry No 518

Transient voltage suppressors

AVX has introduced a low capacitance version of its Transguard bidirectional transient voltage suppressor. Called Antennaguard, it has capacitances below 12pF in an 0603 case size or below 3pF in 0402 or 0603 case sizes. Working voltage is 18V and they operate from –55 to +125°C. They can protect high gain FETs on output stages of RF sections in mobile phones, pagers and WLANs, reducing a 15kV air discharge ESD strike to a level that can be handled by most FETs and preamplifiers.
AVX
Tel: 01252 770000
Enquiry No 505

418MHz radio

Low Power Radio Solutions has released two radio receiver modules based on Micrel MICRF ICs. The LPRx2 and 3 are complete AM superhet receiver modules for upgrading car or domestic alarms. They have a saw-based front end filter to improve selectivity. The LPRx2 is available at either 418MHz as per MPT1340 or 433MHz to ETS300-220. It uses an eight-pin SIL outline.
Low Power Radio Solutions
Tel: 01993 709418
Enquiry No 506

Custom graphics LEDs

A custom graphics LED capability has been announced by Anders Electronics for manufacturers of white and brown goods. Complex icons, different colours and alphanumerics can be combined in one module. Applications include dishwashers, washing machines, microwave ovens and air conditioners.
Anders Electronics
Tel: 0207 388 7171
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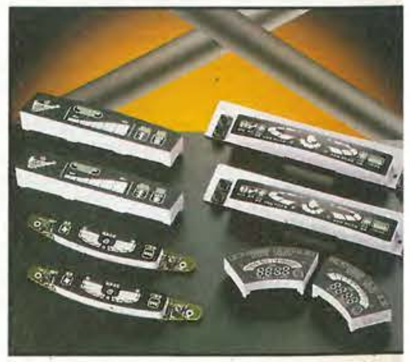


Please quote *Electronics World* when seeking further information

Power amplifier

SiGe Microsystems has introduced a version of its Bluetooth power amplifier. The silicon germanium amp boosts class II Bluetooth radios for class I, 100m range applications. The PA2423M has a current rating of 160mA including the bias network current, and better than 45 per cent power added efficiency. It produces +23.5dBm with 20dB gain and includes power down plus output power control pins. Supply voltage is 2.7 to 3.6V typical and it operates with up to 6.0V applied during battery charging. It comes on tape and reel in a 3 by 3mm MSOP eight-pin package.
SiGe Microsystems
Tel: 001 613 748 1334
Enquiry No 508

Surface mount thermistors

Beyschlag Centralab Components has introduced negative temperature coefficient 0603 surface mount thermistors. Resistance values are 4.7, 10, 47, 68 and 100kΩ. E6 will be added soon. Electrical tolerance is up to 3 per cent on resistance values at 25°C and up to 1 per cent tolerance on B25 and B85 values. Devices are packed on a reel of punched paper tape suitable for pick-and-place mounting. They are stable at temperatures up to 150°C. Applications include battery packs and temperature compensation of LCDs and TCXOs.
BC Components
Tel: 00 31 40 259 0724
Enquiry No 510

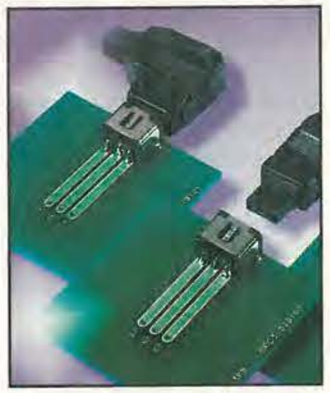
Power amplifier

A dual-stacked smartcard connector has been introduced by AMP for use in set-top boxes. Distance between slots is either 11.4 or 16.2mm and it accepts single-thickness cards. The

product can be supplied with a standard stack height or with a raised floor that lets components be mounted on the PCB beneath the connector. It is possible to use a variable thickness connector as the lower part. The connector is assembled using clip-on techniques and the module's spring contacts.
AMP
Tel: 0181 420 8250
Enquiry No 511

5A power connector

Hirose's RP34L connector is for use in portable equipment. It can handle



up to 5A, 30V AC and 42V DC. The connector can be fully polarised with two or three contacts, with the three-pin version having one leading contact for grounding. The PCB mounted parts are 7mm deep with a metal cover that provides a snap lock mechanism. The two-way plug comes as a prepared cable assembly with straight cable exit. The three-way is available in straight or right angled versions and is wired using crimp contacts and snap together hoods.
Hirose
Tel: 01908 260616
Enquiry No 512

GaAs amplifier

Link Microtek has introduced a monolithic power GaAs amplifier that achieves a typical third-order intercept point of 41dBm. Made by Stanford Microdevices, the SXH-189 operates from 50MHz to 2GHz and is for use as a driver device in infrastructure equipment for cellular, ISM, narrow band PCS and multi-carrier systems. The MMIC delivers a typical gain of 22.0 at 50MHz and 13.0 at 2GHz, with noise figures of 3.5 and 4.2dB, respectively.
Link Microtek
Tel: 01256 355771
Enquiry No 513

Graphic VFD module

The GU800 module from Itron provides a blue-green display in 128 x 32 and 128 x 64 versions and comprises the VFD glass, driver, control Asic with integral refresh twin screen graphic RAM, and interface logic. Models at 160 x 16, 160 x 32 and 256 x 32 will be released later this year. Features include operational temperature between −40 and +85°C, 8-bit parallel interface that is 5V CMOS compatible and suitable for connection to a host CPU bus, and a 5V supply. There is a 16-level brightness control function with a power down facility.
Itron
Tel: 0845 603 9052
Enquiry No 514

Dual-channel analogue chipset

Infineon has introduced the Duslic dual-channel analogue termination for access network and customer premises applications. For decentralised nodes, the software programmable codec SLIC chipset integrates DTMF generation, DTMF decoding, FSK generation (caller ID) and three party conferencing. It also provides rate adaptive echo cancellation complying with ITU G.165. This function eliminates the echo where voice is digitised, so there is no need for network based echo cancellation. This suits wireless local loop or voice over IP systems.
Infineon Technologies
Tel: 00 49 89 234 22767
Enquiry No 509

XDSL network processor

Infineon has introduced the Harrier-XT intelligent protocol handling and controller IC for xDSL customer-premises equipment, including integrated access devices supporting combined data and voice services. It is based on the firm's 32-bit 50MHz Tricore unified processor core architecture and is for ATM and HDLC network termination, Internet connectivity and constant-bit-rate services, such as voice and video delivery. Full-duplex xDSL sustained data throughput rates of 13Mbit/s are supported. There is 48kbyte of on-chip SRAM. The embedded memory supports combinations of instruction, data and scratchpad. Network access protocols can be downloaded into the device for upgrades, modifications to service levels, and design-in to similar network access devices. The IC includes an autosensing Ethernet media access controller and a two-channel HDLC protocol controller. Other onchip features include T1 and E1 circuit emulation, AAL1 and AAL5 ATM adaptation layer support, up to 128 virtual circuits with a priority and WFQ scheduling mechanism.
Infineon technologies
Tel: 00 49 89 234 22767
Enquiry No 515

DC-DC converter

Micro Call has introduced the Linear Technology LT1949 step-up PWM DC-DC converter capable of generating bias voltages for TFT LCD panels. It includes a 1A, 30V internal switch that produces output voltages up to 28V using an inductor. Fixed frequency operation is at 600kHz. An external loop compensation pin lets a

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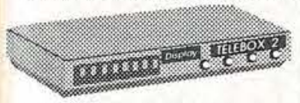

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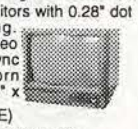

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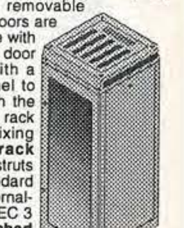

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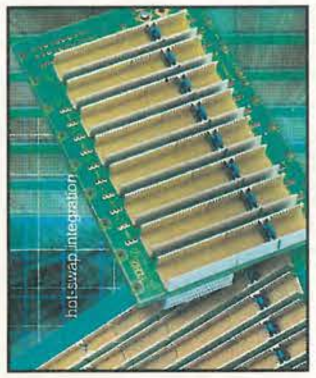
user improve feedback loop performance and use low ESR ceramic capacitors. It comes in an MSOP-8 package.
Micro Call
Tel: 01296 330061
Enquiry No 517

CPCI backplanes

APW Electronics has introduced PICMG CPCI 2.1 revision 3.0 specification left and right hand system slot backplanes. Revision 3.0 resolves the conflicts between the original core specification and the interim hot swap amendment and introduces 66MHz operation, constrained to five slots in a 3V3 signalling environment. Mixed 66 and 33MHz systems are not allowed. On the backplane, to reduce signal skew and improve clock signal integrity, all clock lines must be matched to within ±1mm, run on inside layers and be without vias. For 33MHz eight-slot systems, system boards must generate seven independent clocks and the backplane must route all clocks separately. Mixed 32 and 64-bit operation is allowed.
APW Electronics
Tel: 01489 780078
Enquiry No 519



55/85W supplies

Available from Digital Power, the HPS 85 universal input switched mode power supplies are for power network, computer peripheral, datacoms and telecoms system. They deliver up to 85W output with forced air cooling or 55W with natural convection. Single, dual, triple or quad output versions are available, and each incorporates

Chip inductors for EMI shielding

ELJ chip inductors from Panasonic measuring 1.6 x 0.8mm (0603) are for use in EMI and RFI shielding designs, high-frequency circuitry and power supplies. They suit mobile phones, hand-held instruments and other portable equipment. There are 23 devices with values between 1.5 and 100nH at 100MHz. Maximum typical Q factor at 100MHz is 47, and maximum DC current is from 200mA for the 100nH device to 500mA for the 1.5nH inductors. Operating temperature is between –40 and +85°C. They come in tape-and-reel format for use with automated manufacturing systems. Tolerances are ±0.3nH for devices with inductances up to 3.3nH and ±5 per cent for the rest.
Panasonic Industrial
Tel: 01344 853349
Enquiry No 518

PCI-based digital radio

Spectrum has introduced a multiprocessor, TMS320C6202-based, digital radio. The Ingliston is a quad PCI DSP system using four Texas Instruments' 2000Mips, 250MHz, fixed-point C6202 processors. It is suitable for multi-channel, multi-function applications such as signal intelligence, medical and commercial imaging, and third-generation wireless base stations. Each processor contains 3Mbit memory. Also available are complementary DSP and I/O hardware, software development kit, system management software, and digital radio specific DSP algorithms. The I/O system has one PMC site, two PEM sites and DSP-Link3, which provides access to more than 150 Industrypack modules. Ingliston can also support the 300MHz TMS320C6203 with 7Mbit memory.
Spectrum Signal Processing
Tel: 001 604 421 5422
Enquiry No 534

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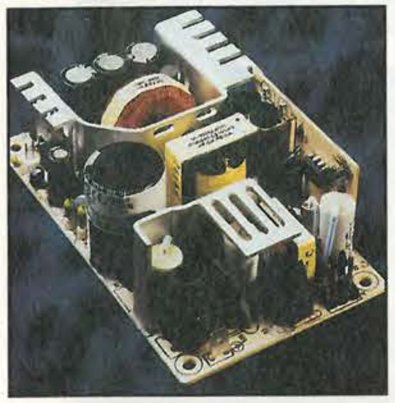
a +12V fan output. Based on a 7.6 by 12.7cm footprint, they can be specified with an optional cover and chassis. They meet requirements for conducted emissions including EN55022 class B. Where power factor correction is required, models are available that meet EN61000-3-2 for input line harmonics.
Digital Power
Tel: 01722 413060
Enquiry No 520



Dual CAN micro

Dallas Semiconductor has announced the DS80C390 8051 microprocessor that integrates two Can bus controllers with peripherals. It addresses 4Mbyte of external data memory and 4Mbyte of external program memory. Applications include factory automation, marine control and navigation, industrial control systems, and HVAC control.
Dallas Semiconductor
Tel: 001 972 371 3832
Enquiry No 521

PC plug-in card

A plug-in ISA card for a PC provides functions to analyse, decode and process communications transmissions in the HF, VHF, UHF and SHF bands. It is available from Sight Systems. The Wavecom W41PC provides government bodies, telecoms authorities and the military with a signals intelligence gathering system for radio communications, datacomms, fax and pager signals. The system may be configured for stationary monitoring of one transmission with one system or as an automated broadband monitoring system using a network of cooperating systems.
Sight Systems
Tel: 01903 504494
Enquiry No 522

MPEG-2 decoder cards

Imagine Graphics has introduced MPEG-2 digital Pal and NTSC video and audio decoder cards from Visual Circuits. As well as composite and S video outputs, the Reeltime Plus cards have RGB video and balanced audio outputs on each channel for professional video engineers needing to interface to video projection and PA systems. There are four and two channel versions. Various cable assemblies and breakout panels are available as options. The MPEG Plus card, a half-length PCI card with RGB, S video, composite video and balanced audio outputs, is also available.
Imagine Graphics
Tel: 01727 844744
Enquiry No 523

3MHz precision LCR bridge

Wayne Kerr's latest range of precision component analysers extend the capabilities of previous LCR measurement bridges by measuring parameters up to 3MHz frequencies and supporting for the first time real-time swept frequency measurements. The company claims an industry-leading basic accuracy of 0.02 per cent for the 6430A (20Hz - 500kHz) and 6440A (20Hz - 3MHz) analysers. Significantly, pricing for the instruments is less than for previous bridges despite higher accuracy and frequency specifications. For capacitor testing, when dissipation factor measurements are carried out accuracy is 200ppm. The company has also added extra measurements, such as admittance, conductance and resonant frequency, to the standard list of L, C, R, Z, Q, phase and dissipation. A real-time swept frequency measurement provided by the 6440A provides either linear or logarithmic axes and measurement functions such as impedance and phase.
Wayne Kerr
Tel: 01243 825111
Enquiry No 524

PCB adapter

A PCB adaptor has been introduced by Harting for the Han 05/0 motor interface connector, letting it be directly mounted on a PCB. Part of the company's Han Drive family, it is claimed to help set up decentralised motor control with less wiring. The ability to mount the connector directly on a PCB eliminates the need for manual wiring and lets the production of the electronic equipment and the mechanical components, including the cabinet, be separated and automated.
Harting
Tel: 01604 766686
Enquiry No 525

Switching relay

Teledyne's RF300 to 303 DPDT and SPDT switching relays are for RF

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switching circuits. To operate from DC to 3GHz, with either a 450 or 200mW coil, the relays provide insertion loss repeatability of ±0.1dB from 0.3MHz to 3GHz. Its fully welded, hermetically sealed TO5 style metal can ensures that the switching path is not subject to RF noise disturbances and transient impact.
Teledyne Relays
Tel: 01634 863494
Enquiry No 526

SM antennas

Philips has introduced two SM ceramic antennas that can be assembled directly onto a mobile phone's PCB using pick-and-place equipment and reflow soldering. The narrow-band dielectric resonator antenna (DRA) suits Bluetooth, Dect, PHS and W-LAN, and the wideband printed wire antenna (PWA) is for GSM, DCS, Amps and CDMA. The DRA is shaped like a ceramic block. It can also be used as a second diversity-antenna to improve the quality of voice or data reception. The PWA can be used for dual-band applications, such as 900 and 1800MHz GSM phones, and in PDAs.
Philips Components
Tel: 00 31 40 272 20 91
Enquiry No 528

Thin-substrate assemblies

Autosplice has available a knurled pin design for use in assemblies requiring thin PCB substrates. Knurl pins are claimed to provide the improved thermal characteristics of a thin PCB and high pin-retention strength for small DC to DC converters. For use with the firm's star-pin technology, the knurl pin has raised bumps on its surface rather than uniform star-shaped star-pins. The irregular spacing of the bumps provides a mechanically stable, interference contact even when inserted into PCBs as thin as 0.08cm. The knurl pin suits pin-through-paste and bottom side paste operations typical in double-sided assembly operations on thin PCBs because the pin protrusion can be held to no more than 0.013cm on the bottom of the board. The pin is inserted into a thin PCB on the top side of an SMT assembly where SMT components will be applied on the bottom side. The pin is not to protrude more than 0.013cm out of the bottom. The PCB is flipped, the bottom side is stencilled with paste, the surface mount components placed in the paste and then reflowed. During reflow the paste applied at the areas over the holes where the pins were inserted reflows and creates a fillet throughout the hole and on bottom and top side. Designs are available in 0.046, 0.063, 0.079 and 0.1cm round configurations, with plans to add 0.15 and 0.21cm round configurations soon.
Autosplice
Tel: 001 619 535 0077
Enquiry No 532

Dot matrix yellow LED displays at 45°

A 16 x 32 dot matrix LED display from Rohm combines yellow output with a 45° viewing angle and 10mm profile. The LUM-512HY350 has a 6mm dot pitch and can be stacked vertically and horizontally. The 512 dot modules can be used as building blocks for high resolution custom applications such as large passenger information displays and messaging boards. Each unit integrates driving and control circuitry into a module measuring 96 by 192mm and weighing 100g.
Rohm Electronics
Tel: 00 49 89 234 22767
Enquiry No 516

8-bit micro

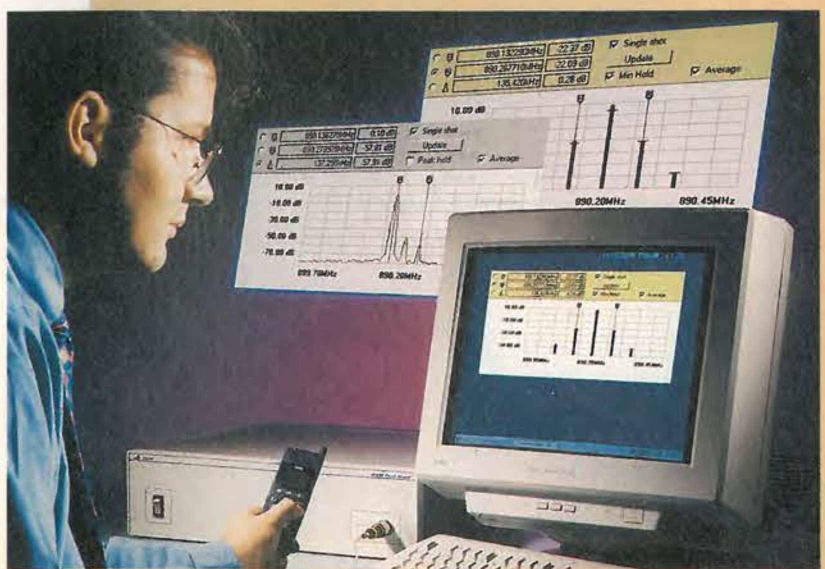
Epson has added the EOC88409 to its 8-bit family of microcontrollers. The single chip product is based on the firm's EOC88 CMOS 8-bit core CPU. A built in A/D converter operates at 2MHz with 10-bit resolution and a minimum conversion time of 11µs. The device can control up to 4M x 3byte of memory with the 22-bit external address bus and 3-bit chip enable signals. It comes in a QFP15-100 pin or die.
Epson Electronics
Tel: +49 89 1 400 5363
Enquiry No 529

Resistors for voltage balancing

Meggitt has released YP resistors to fit on the screw terminals of PEH200 or similar aluminium electrolytic capacitors. They are for voltage balancing of series connected capacitors or capacitor discharge in high voltage circuits. Initial values are 18, 27 and 47kΩ but other values can be supplied to order. Hole centres are 32mm (M5) and 13W can be dissipated with a hot spot temperature of 40°C. Operating temperature is –55 to +105°C. By offsetting resistors, parallel connections can double capability.
Meggitt Electronic Components
Tel: 01793 611666
Enquiry No 530

Load driver supports 40V operations

Temic Semiconductors has launched a general-purpose BCDMOS load driver. The U6815BM can drive actuators and lamps in automotive and industrial applications. It is made on a 0.8µm process to support operations up to 40V. It can be used in most microcontroller subsystems that control up to six high and six low-side loads. Applications include DC motor control for flaps in air conditioning systems, lamp and LED drivers, and drivers for valves, solenoids and relays. The device has a bidirectional serial interface to the microcontroller for control and diagnostic functions. Each driver stage can directly support 600mA typically. Several parallel combinations are possible. The high and low-side outputs are at adjacent pins. The intelligent load driver is protected against open loads, short circuits, over and undervoltage and over temperature. The diagnostic block reports malfunctions to the microcontroller so the system can take counter measures.
Temic
Tel: 00 33 240 181987
Enquiry No 531



GSM test with I/Q adjustment

Two capabilities have been added to the IFR 2935 Phonetest GSM test set – an in-channel spectrum analyser and an I/Q (a complex modulation technique) adjustment display. There have also been improvements in level accuracy, VSWR and power burst analysis. Applications include go or no-go test, mobile alignment, repair and final QA audit assessment. Optional software provides maintenance and production facilities.
IFR International
Tel: 01438 742200
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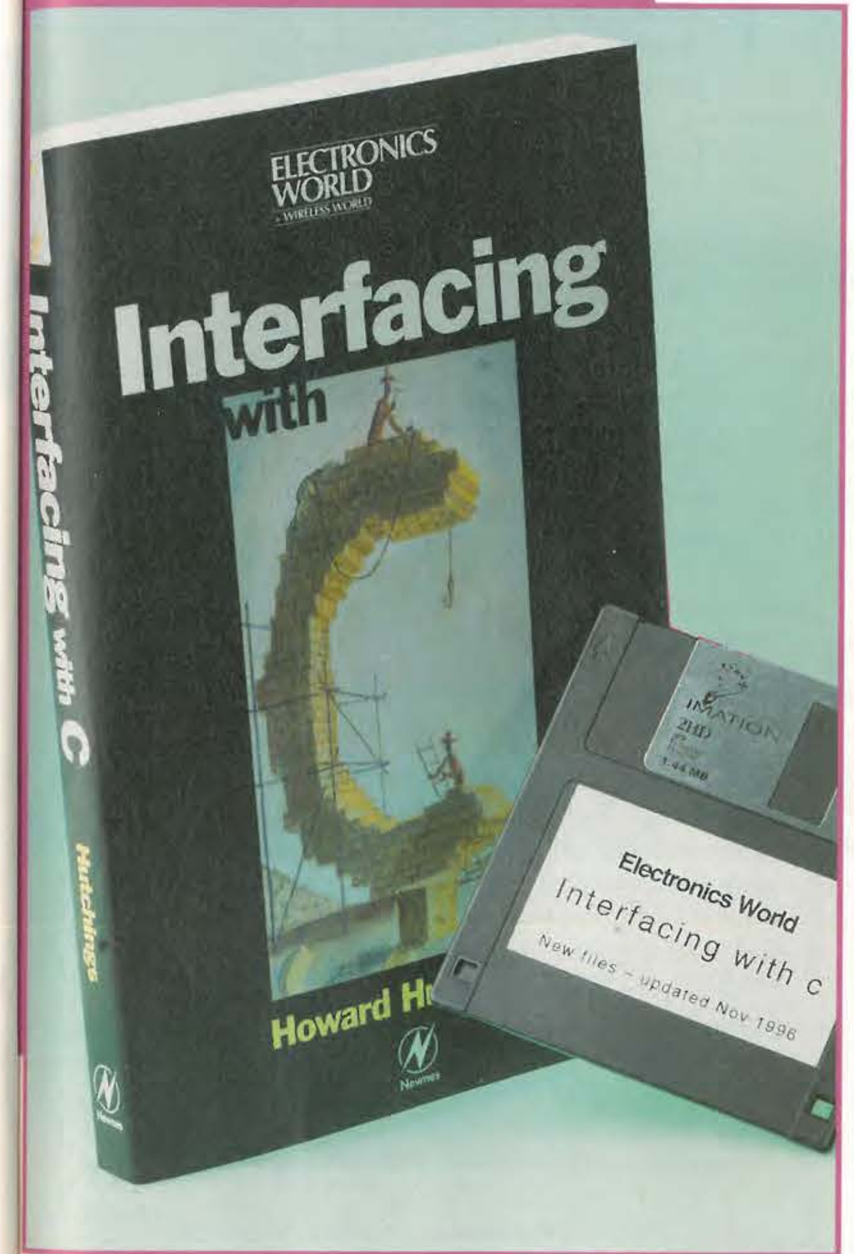

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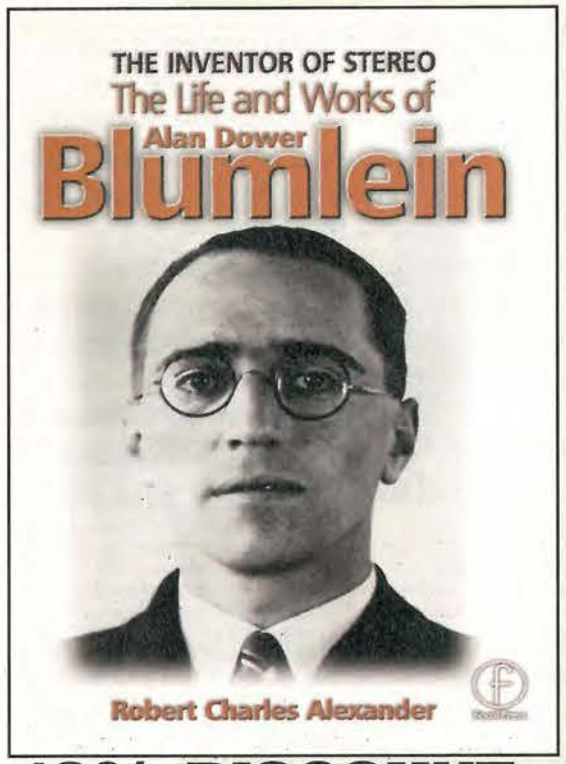

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This book is the definitive study of the life and works of one of Britain's most important inventors who, due to a cruel set of circumstances, has all but been overlooked by history.

Alan Dower Blumlein led an extraordinary life in which his inventive output rate easily surpassed that of Edison, but whose early death during the darkest days of World War Two led to a shroud of secrecy which has covered his life and achievements ever since.

His 1931 Patent for a Binaural Recording System was so revolutionary that most of his contemporaries regarded it as more than 20 years ahead of its time. Even years after his death, the full magnitude of its detail had not been fully utilized. Among his 128 patents are the principal electronic circuits critical to the development of the world's first electronic television system. During his short working life, Blumlein produced patent after patent breaking entirely new ground in electronic and audio engineering.

During the Second World War, Alan Blumlein was deeply engaged in the very secret work of radar development and contributed enormously to the system eventually to become 'H2S' - blind-bombing radar. Tragically, during an experimental H2S flight in June 1942, the Halifax bomber in which Blumlein and several colleagues were flying, crashed and all aboard were killed. He was just days short of his thirty-ninth birthday.

For many years there have been rumours about a biography of Alan Blumlein, yet none has been forthcoming. This is the world's first study of a man whose achievements should rank among those of the greatest Britain has produced. This book provides detailed knowledge of every one of his patents and the process behind them, while giving an in-depth study of the life and times of this quite extraordinary man.

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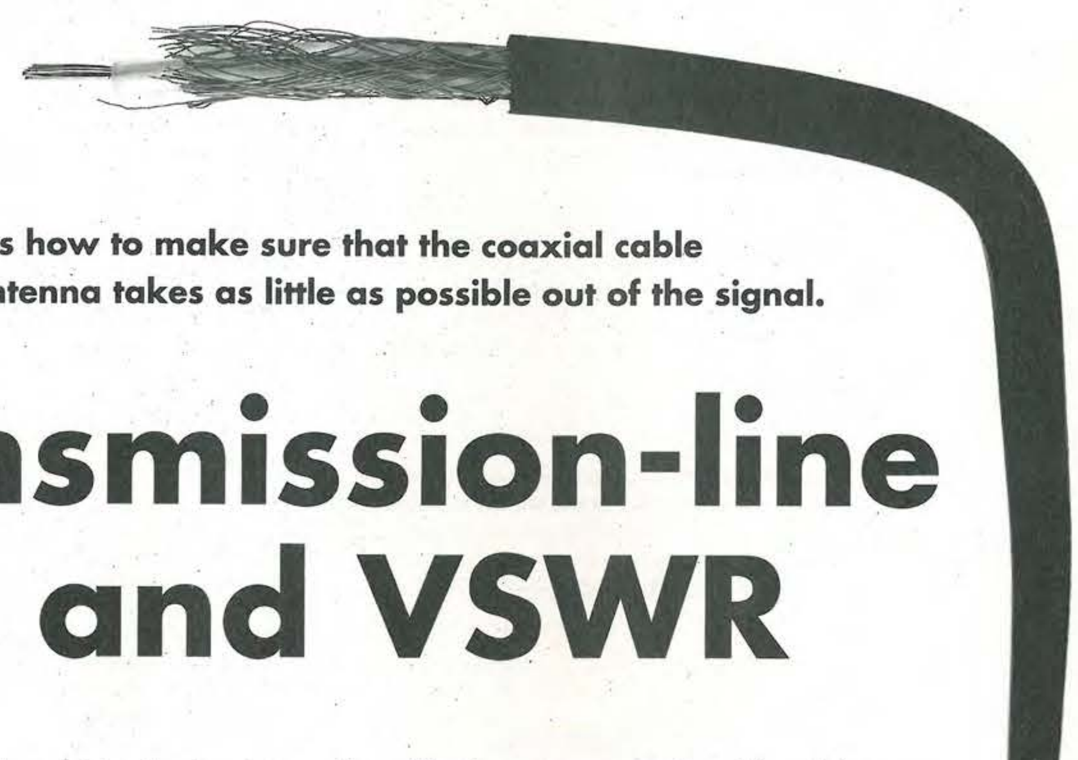

Joe Carr explains how to make sure that the coaxial cable connecting an antenna takes as little as possible out of the signal.

Transmission-line loss and VSWR

All electrical sources have an internal resistance or impedance: batteries, the AC power mains, radio transmitters, and signal generators all have some value of internal resistance, **Fig. 1**.

In transmitters and signal generators that value is usually stated as 'output impedance.' The impedance could be any value $Z_S=R_S\pm jX$, where R_S is the resistive component and X is either a capacitive or inductive reactance.

In RF circuits other than television the standard system impedance is 50+j0, or simply 50Ω resistive – i.e. not reactive.

A load will also have some value of impedance: $Z_L=R_L\pm jX$. Although transmitters and signal generators will control the output impedance so that it is purely resistive, and matches the accepted value of standard impedance – i.e. 50Ω, except for TV, which is usually 75Ω – the load may vary over quite a range of values of R and X.

It is rare to find a load connected directly to the output of a transmitter or signal generator. Most commonly one expects to see a transmission line, **Fig. 2**, between the source and load. The transmission line also has a characteristic impedance, which is denoted by Z_O.

The beauty of matching

One the fundamental facts about connecting source to load is that maximum power transfer occurs when the load and source are matched, i.e. when $Z_S=Z_L$. If this is not the case, then not all of the power is delivered to the load.

In a transmitter-antenna system, mismatch means not all of the available power is radiated as a radio signal. In a signal generator test set-up, it means a possibly erroneous measurement – and, rarely, damage to the measuring equipment.

The problem becomes more complicated in real situations because of the increased number of possible mismatches. It is necessary to match Z_S to Z_O, and Z_O to Z_L.

If a load is not matched to a source, then some of the RF power supplied by the source will not be absorbed in the load. It will be reflected back towards the source. Thus, we must contend with both the applied, or forward, power P_F supplied by the source, and the reflected power, P_R, rejected by the load.

Standing waves

When the forward and reflected waves of an RF signal combine in the transmission line, they algebraically add, and set up a pattern of standing waves.

In the case where $Z_O=Z_L$, there are no standing waves, and such a line is said to be 'flat.' But if Z_O is not equal to Z_L, then the standing waves emerge, producing a non-zero reflected power, and voltage nodes – and current nodes, incidentally – along the line.

The voltage will vary from a maximum, V_{MAX} to a minimum, V_{MIN}. The maxima and minima are quarter wavelength apart, and repeated maxima and repeated minima are half wavelength apart. One implication of this situation is that an impedance connected to the end of a transmission line is repeated every half wavelength.

The reflected wave can be defined in terms of a *reflection coefficient* Γ.

$$\Gamma = \frac{Z_L - Z_O}{Z_L + Z_O} = \frac{V_S}{V_F} = \sqrt{\frac{P_R}{P_F}} \qquad (1)$$

The standing waves are defined in terms of the *standing-*

wave ratio, which can be calculated from the reflection coefficient,

$$SWR = \frac{1 + ABS(\Gamma)}{1 - ABS(\Gamma)} \qquad (2)$$

The SWR can also be defined in terms of impedances,

$$SWR = Z_L/Z_O \text{ if } Z_L > Z_O \qquad (3)$$

Or,

$$SWR = Z_O/Z_L \text{ if } Z_O > Z_L \qquad (4)$$

Finally, SWR can be defined in terms of power,

$$SWR = \frac{1 + \sqrt{\frac{P_R}{P_F}}}{1 - \sqrt{\frac{P_R}{P_F}}} \qquad (5)$$

Note: Although SWR can be measured using any of these properties, it is common practice to use the term voltage standing wave ratio, or VSWR as synonymous with SWR. That usage will be observed herein.

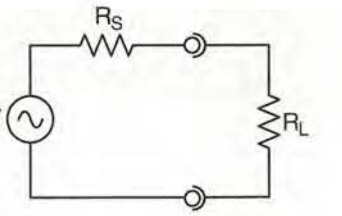

Fig. 1. In practice, all signal generators have an internal resistance, R_S, that affects how much power can be delivered to the load, R_L.

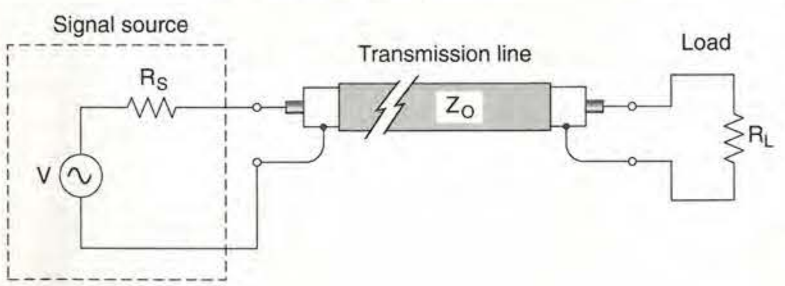

Fig. 2. Since it is not usually possible to site a transmitter right next to its load, power is normally passed to the load via a transmission line with a characteristic impedance, Z_O.

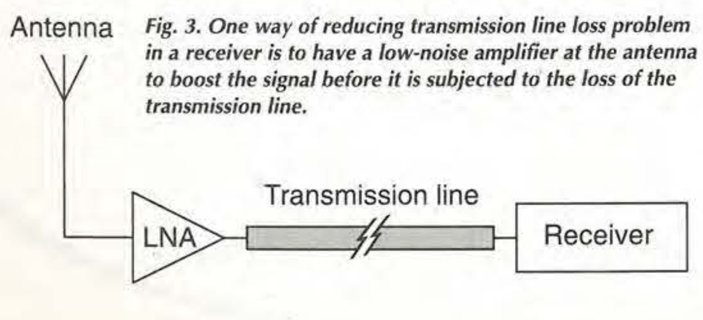

Fig. 3. One way of reducing transmission line loss problem in a receiver is to have a low-noise amplifier at the antenna to boost the signal before it is subjected to the loss of the transmission line.

A scenario

A young apprentice technician was sent to check out a newly installed low-band VHF – i.e. 30-50MHz – communications antenna. He returned and told his boss that everything was in order because the VSWR was about 1.66:1, which was less than 2:1, so it met specifications. The wizened old boss was sceptical. He agreed that a VSWR of <2:1 was within the specification. But the boss knew that low VSWR sometimes conceals deeper problems.

The antenna was mounted on a tower, and the tower was up on the crest of a hill. Altogether, there was 250ft (76.2m) of RG-8/U 52Ω transmission line. On further questioning, the boss discovered that the particular model antenna used on the system had a feed-point impedance of 300Ω resistive. That would infer a VSWR of 300/52=5.77:1, not 1.66:1 that the technician measured.

So what happened? The VSWR should be almost 6:1, but it only read <2:1. Why? The solution is in the loss of the transmission line. There are two basic forms of loss in coaxial cable: *copper loss* and *dielectric loss*. The copper losses result from the fact that the copper used to make the inner conductor and shield has resistance.

The picture is further compounded by the fact that RF suffers skin effect, so the apparent RF resistance is higher for any given wire than the DC resistance. Further, the skin effect gets worse with frequency, so higher frequencies see more loss than lower frequencies.

The copper loss is due to the I^2R when current flows in the conductors. When there is a high VSWR at the load end, part of the power is reflected back down the line towards the transmitter. The effect of the reflected RF current is to increase the average current in the conductor. Thus, I^2R loss increases when VSWR increases.

Similarly with dielectric loss. This loss occurs because voltage fields of the RF signal cause problems. The simplistic explanation is that the voltage fields tend to distort electron orbits, and when those orbits return to their normal state some energy is lost. These losses are related to E^2/R.

As with copper loss, the dielectric loss is frequency sensitive. The *loss factor* of coaxial cable increases with frequency. Let's look at two examples.

Cable characteristics

Table 1 shows a popular, quality brand of two 52Ω coaxial cables: RG-8/U and RG-58/U. Note that the loss for the RG-8/U varies from 1.8dB/100ft (0.059dB/m) at 100MHz to 7.10dB/100ft (0.233dB/metre) at 1000MHz. The RG-8/U cable is larger than the RG-58/U, and has less loss.

Take a look at the same end points for the smaller cable: 4.9dB/100ft (0.161dB/m) at 100MHz and 21.50dB/100ft (0.705dB/m) at 1000MHz.

Clearly, the selection of the cable type is significant. At the FM broadcast band of 100MHz, the smaller cable would show 12.25dB of loss. If a 100W signal is applied at the transmitter end, the ratio of loss is $10^{(12.25/10)}$=16.79, so only about 6W is available to the antenna.

The problem is even worse if the antenna is used for 900MHz cellular telephones. In that band, the RG-58/U cable loss is 20dB/100ft (0.656dB/m), so the overall loss is a whopping 50dB. The power ratio is $10^{(50/10)}$= 10^5:1, so only about 10μW makes it to the antenna.

Table 1. Loss in decibels per 100ft for two popular coaxial cable types.

Type	100MHz	200MHz	300MHz	400MHz	500MHz	600MHz	700MHz	800MHz	900MHz	1GHz
RG-8/U	1.80	2.70	3.45	4.20	4.73	5.27	5.80	6.25	6.70	7.10
RG-58/U	4.90	7.30	9.40	11.50	13.33	15.17	17.00	18.50	20.00	21.50

If RG-8/U cable were used instead of RG-58/U, then the losses would be 4.5dB at 100MHz and 16.75dB at 900MHz. At the cellular frequencies in the 900MHz band the loss factor still be high, about 47:1, so only 2W would make it through the loss. The rest is used to heat up the coaxial cable.

Fortunately, certain specialty cables are available with losses around 2.5dB/100ft (0.082dB/metre) at 900MHz. Such cable would produce about 6.25dB of overall loss, or a ratio of 4.2:1. That cable would deliver nearly 24W of the original 100W.

And on the receiving side...

The problem is also seen on receiver systems. Suppose that a 900MHz receiver is at the end of a 250ft transmission line. Further suppose that the signal is a respectable 1000μV, which in a 50Ω load is –47dBm. A loss of 6.25dB would make the power level at the antenna terminals of the receiver –47dBm–6.25dB=–53.25dBm, or about 485μV, which is still a reasonable signal. But if the lesser coaxial cable is used (RG-58/U instead of the specialty grade), then the loss is 50dB, and the signal at the receiver would be –47dBm–50dB=–97dBm. This level is getting close to the sensitivity limits of some receivers, or about 3.2μV.

The above example was from telecommunications, but it can apply equally whenever RF is sent over coaxial cable. Cable-tv operators, local-area networkers, and other users of strictly land-line RF also see the same loss effect. The correction is shown in **Fig. 3**. Here, a low-noise amplifier, LNA, is placed at the head end of the transmission line to boost the signal before it suffers loss.

Positioning the booster

On first blush it might seem easier to put the amplifier at the receiver end. But that doesn't work out so well because of two factors. First, there is an inherent noise factor in any amplifier. If the signal is attenuated before it is applied to the amplifier, then the ratio of the signal to the internal noise of the amplifier is a lot lower than if the signal had been applied before attenuation.

So, while the signal would still be at the same level regardless of where the amplifier is placed, the all-important signal-to-noise ratio is deteriorated if the amplifier is at the receiver end. The second reason is that any lossy device, including coaxial cable, produces a noise level of its own,

$$F_{N(coax)} = 1 + \frac{(L-1)T}{290} \qquad (6)$$

Here, $F_{N(coax)}$ is the noise factor of the coax, L is the loss express as a linear quantity, and T is the physical temperature of the cable in kelvin.

The linear noise factor due to loss can be converted to noise figure, which can be added to the system noise decibel for decibel.

Now let's return to the problem of the technician and his boss. The tables from the coaxial cable maker told the boss that the loss at 40MHz is 1.2dB/100ft, so the overall loss is 3dB – halving the power. This loss is called the *matched line loss*, designated L_M. But, you also have to consider the total line loss, *TLL*, which is,

$$TLL = 10\log\left[\frac{B^2 - C^2}{B(1 - C^2)}\right] \qquad (7)$$

Here, B=antilogL_M, $C=(SWR_{LOAD}-1)/(SWR_{LOAD}+1)$ and SWR_{LOAD} is the VSWR at the load end of the line.

The boss knew that the VSWR measured at the load end might be considerably higher than that measured at the transmitter end of the line. Given that L_M=3dB,

$$B = LOG^{-1}(3) = 1.995.$$

If $VSWR_{LOAD}$=5.77, as is the case, then C=(5.77–1)/(5.77+1)=4.77/6.77 = 0.705. Thus, the TLL is,

$$TLL = 10\log\left[\frac{1.995^2 - 0.705^2}{1.995(1 - 0.705^2)}\right]$$

$$= 10\log\left[\frac{3.98 - 0.497}{1.995(0.497)}\right]$$

$$= 10\log\left[\frac{3.483}{0.992}\right] = 10\log 3.51 = 10 \times 0.545 = 5.45\text{dB}$$

The VSWR at the input end of the line, down the hill by the transmitter, then is,

$$VSWR_{in} = \frac{B + C}{B - C} = \frac{1.995 + 0.497}{1.995 - 0.497}$$

$$= \frac{2.492}{1.498} = 1.66:1$$

When he had made the calculations, the boss turned to the younger technician, and told him that he had just improved his experience level. The lad was on his way to the wisdom that real experience provides.

As the boss headed out to the local honky-tonk* to have his pint and a bowl of grits†, he told the boy: "Now you shag your lazy butt back up that hill and impedance match that damn antenna!"

■

* "Honky-tonk" is from the dialect of the American Appalachian section, and translates into real English as "Pub of low repute, with country music". Given that the Appalachian dialect is ancestrally related to Border Scots English, one wonders where this word originated.

† You don't even want to know what Americans call 'grits' tastes like. One noted Englishman likened it to wallpaper paste.

Transformers from coax

Chris Hancock offers easy solutions to making baluns and combined impedance/voltage transformers from standard co-axial cable and microstrip-line structures made using commercially available board.

Normally, antennas used to transmit RF or microwave power are fed from an unbalanced transmission line with an impedance of either 50Ω or 75Ω.

Commonly used transmission line structures are microstrip and co-axial lines. Microstrip lines are often used for hand held transmitters, such as those in mobile phones. Co-axial lines are used to feed antennas in large portable, and permanent transmitters, as found in wireless local area networks, base-stations, broadcast TV, radar and local microwave links.

In most cases, an antenna requires a balanced feed with respect to ground. This makes it necessary to use a device that converts the unbalanced output of the feed cable to a balanced output required by the antenna.

Depending on the structure used, which is governed by the application, the input impedance of the antenna can vary drastically. Ideally, the antenna is designed to present a purely resistive load to the feed line, **Table 1**. In order to get maximum energy into the antenna, it is necessary to match the impedance of the feed line to that of the antenna input. If there is a reactive element present, then this should be matched at the source.

*Chris Hancock MSc PhD

Table 1. Typical antenna input impedances[1,2].

Antenna type	Typical input impedance (Ω)
Marconi monopole	36
Turnstile	38
Hertz – end-fed half-λ dipole	73
Helical beam	100 to 200
Log-periodic array	200 to 800
Folded dipole/conical	300

At the same time, it is necessary to provide an unbalanced to balanced condition in order to keep the overall system in a balanced state, as mentioned above. This article presents a simple, practical, method of achieving both of the above, along with some interesting extensions to the solution that may be used to give other desirable effects.

How it works

Figure 1 presents the basic idea of an unbalanced-to-balanced system. You can see that the coaxial feed line has an outer conductor that is grounded. The two conductors – outer braid and centre inner – therefore, do not carry signals with the same relationship to ground potential. Consequently, this line is termed an unbalanced line.

It is necessary to change from this unbalanced condition to a balanced condition in order to drive a balanced load, such as a dipole antenna. If this was not done, the antenna lead connected to the ground of the feed cable would ground that point of the antenna and impair operation.

Standard transformers are of limited use in this type of application due to excessive losses at high frequencies. This limitation can be overcome by using either co-axial, or microstrip, half wavelength, $\lambda_g/2$, transmission line structures.

You can see from Fig. 1b) that the inner conductor of the line is tapped at $\lambda_g/2$ from the end. The tap (far end) and the near end of the inner conductor provide two signals of equal amplitude but 180° out-of-phase. Neither of the signals is grounded, hence providing the required condition for balance.

This unbalanced-to-balanced arrangement is known as a *balun*. It is reversible in that it functions equally as a balanced to unbalanced converter.

As well as acting as a balun, this arrangement also provides impedance and voltage transformation. The principle of operation is shown in **Fig. 2**.

If the load (antenna, etc) is balanced, i.e. $R_{L1}=R_{L2}$, then R_{L2} is transformed by the half wave line to a similar value at the junction with the main feed cable, where it is paralleled with R_{L1}.

Current delivered from the feed line, i_t, must be double that required in each of the two resistors, i.e. $i_t=i_1+i_2$. Therefore, the sum, $R_{L1}+R_{L2}$, is four times the impedance seen at the input, i.e. $R_{L1}+R_{L2}=R_{out}=4R_{in}$.

Voltages V_1 and V_2 are equal in amplitude and opposite in phase. If it is assumed that the $\lambda_g/2$ transmission line is loss-less, i.e. $V_1=V_2=V_{in}$, then the difference voltage is twice the amplitude of the input voltage, i.e. $V_{out}=V_{in}-(-V_{in})=2V_{in}$.

The relationship between the current, voltage and impedance seen at the inputs and outputs of the transmission line transformer is thus the same as those for a conventional, low frequency, transformer, as given by,

$$\frac{R_{out}}{R_{in}}=\left(\frac{V_{out}}{V_{in}}\right)^2=\left(\frac{I_{in}}{I_{out}}\right)^2 \qquad (1)$$

Two practical applications of this arrangement are given shown in **Fig. 3**. The first shows a 50Ω transmitter source matched to a 200Ω helical-beam or a log-periodic antenna load. The second shows a 75Ω transmitter matched to a folded dipole, or conical antenna.

Note that the characteristic impedance of the cable used in these two arrangements is the same at that of the source in order to prevent any mismatch between the transmitter source and the unbalanced feed line.

The connection between the unbalanced line and the $\lambda_g/2$ transformer should be always such that a co-axial environment is maintained with an impedance equal to the characteristic impedance of the line and free from discontinuities. If this is not the case you could get a mismatch at the junction and consequently a VSWR of greater than unity on the feed line.

Benefits of multi-section transformers

It is possible to extend the network to comprise multiple $\lambda_g/2$ sections in order to get some interesting and desirable results.

Assume, for example, three sections cascaded together as in **Fig. 4**. The source impedance is 50Ω. Provided that the characteristic impedance of the third section matches the unbalanced output impedance of the second section, the output impedance will be matched to a load of 800Ω, i.e. $R_L=16\times R_{in}$. The voltage will be four times the input voltage.

Such a transformer provides a method of driving a balanced, high-impedance load from an unbalanced, low-impedance source at VHF and UHF. It does so without material limitations experienced using standard iron dust and ferrite transformer cores.

Note that in order to drive the third section, an unbalanced condition is required, which is achieved by using a separate $\lambda_g/2$ section. Transitions starting from the source are: unbalanced-balanced-balanced-unbalanced-balanced.

It is sometimes required to solely go from the balanced to unbalanced state. This might be the case in the push-pull output stage of an amplifier, where it may be required to drive a single ended unbalanced load.

This method can be used to produce high voltages at very high frequencies where only a low-voltage generator is available.

With this method, it is vital to have each of the sections contained in the network perfectly matched, as in Fig. 4. This is because zero mismatch between junctions, where $\Gamma=0 \Rightarrow VSWR=1$, means that all of the transmitted energy will reach the load.

If there are reflections at the junctions then voltage maximums along the line may exceed the dielectric strength of the cable, causing breakdown.

Also, at frequencies greater than 1GHz, dielectric losses become large. A high VSWR can lead to dielectric heating at voltage maximums along the line. This could lead to overheating at these points.

The existence of reflected waves means greater power (I^2R) losses. This becomes an important issue with long feed lengths.

Finally, a high VSWR leads to increased noise and can cause the transmission of 'ghosts' when transmitting video and data signals[3].

Limitations

Taking this theory to the limit, **Fig. 5** shows a hypothetical situation where it is required to transform a 50Ω, 50V source to 12.8kΩ, 1600V.

If a co-axial construction that uses a low density PTFE dielectric with a typical permittivity value of 2.2 is considered, then characteristic impedance Z_o is given by.

$$Z_o=\frac{138}{\sqrt{2.2}}\log_{10}\frac{D}{d}$$

Where D is the inner diameter of the outer conductor and d is the outer diameter of the inner conductor.

The transformation can be achieved, theoretically, using nine $\lambda_g/2$ sections. The problem is that in order to attain the desired characteristic impedance for sections seven/eight and nine, i.e. 3.2kΩ and 6.4kΩ respectively, the ratios for D/d are 2.45×10^{34} and 6×10^{68} respectively!

A further issue that must be considered in this case is that of the breakdown voltage at the final stage; if standard co-axial cable was used then it is likely that it would break down at the voltage maximum, and would most certainly not tolerate a high VSWR.

And in practice?

I have made a number of such structures from flexible and semi-rigid co-axial cable and from microstrip lines. The main advantage of using microstrip-lines is that the characteristic impedance can easily be altered by changing the width of the line.

I should point out though that the normal line-impedance range for microstrip is between 20Ω and 125Ω[4].

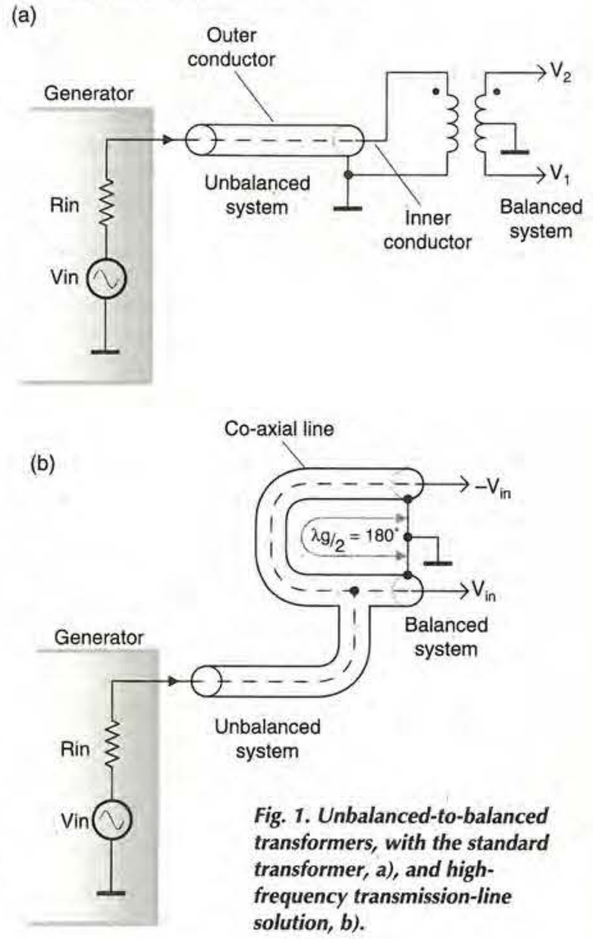

Fig. 1. Unbalanced-to-balanced transformers, with the standard transformer, a), and high-frequency transmission-line solution, b).

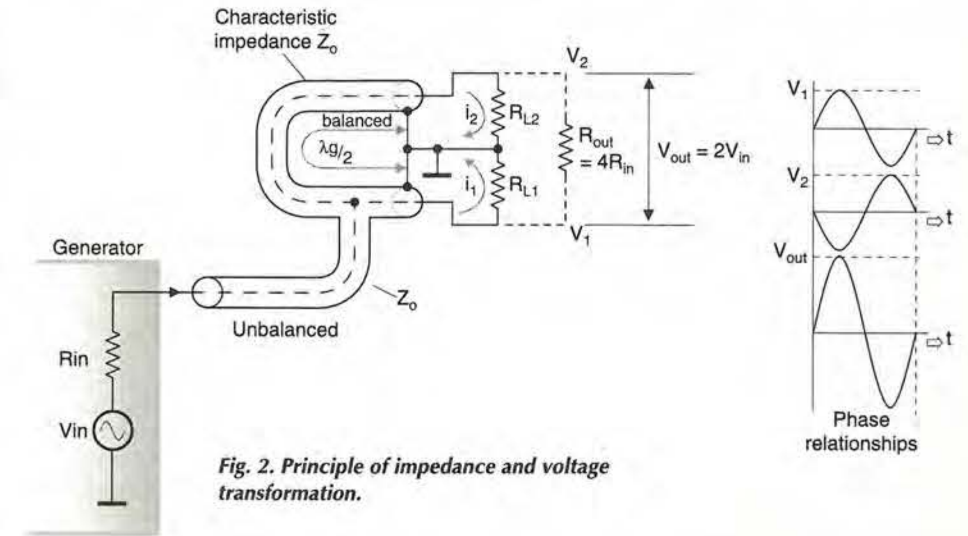

Fig. 2. Principle of impedance and voltage transformation.

Problem one. I needed to match a low impedance 50Ω, 100V source to a load of around 800Ω, 200V. Common-mode current returning back to the source had to be minimised so that there was no live RF voltage on the outer sheath of the co-axial feed cable. The frequency involved was 500MHz.

A 50cm long length of LFH-316 co-axial cable was used for the first two sections. The $\lambda_g/2$ length was accurately measured using a Hewlett Packard HP4396A, 100kHz to 1.8GHz network/spectrum analyser.

A length of cable greater than $\lambda_g/2$ was connected to the analyser's port one. With the Smith chart option selected, the S_{11} scattering parameter was measured. The cable length was shortened until one complete circumference of the Smith chart was traversed, i.e. 180Ω rotation from open circuit, towards the generator, and back again.

The line did not exactly follow the outer perimeter of the chart due to loss in the cable at 500MHz; also the open-circuit line radiates power from the end, but is still far more convenient for trimming purposes than using a shorted line.approach

In this first section, the input from the generator was connected to a SMA connector at one end. The outer sheath was connected locally together, hence the balanced output from the first section consists of the input SMA centre pin and the same centre, but $\lambda_g/2$ away: compare this with Fig. 1.

The third section was connected directly to the second section in the same manner and consisted of a length of a $\lambda_g/2$, 200Ω cable especially made for the application. Again, the outer sheaths were connected together and the two balanced outputs from the first section drove the balanced input of the second section, providing an unbalanced output required to drive the third section: compare this with Fig. 3.

Problem two. In instance two, I needed to match an unbalanced 50Ω source to a 200Ω balanced load at 2.45GHz and provide good isolation between the source and the load.

To achieve this, three pieces of semi-rigid, 50Ω RG402/U with an outside diameter of 3.58mm and breakdown voltage 1.9kV rms were used for the three $\lambda_g/2$ sections.

A $\lambda_g/4$ microstrip transformer, i.e. impedance inverter, with a characteristic impedance of 100Ω, was used between the second and third $\lambda_g/2$ sections to match the 200Ω output impedance from the first section to the 50Ω of the third section, **Fig. 6**.

The 100Ω microstrip section was made using RT Duroid 5880[5] substrate. This is a non-woven glass microfibre reinforced PTFE composite structure whose permittivity, ε_r, at 10GHz is 2.2 ±0.02. Substrate thickness chosen was 3.175mm (125mil), giving a line width of 2.81mm (111mil) for a 100Ω line impedance.

All lines were cut using a sharp scalpel blade, and the widths were adjusted to the desired impedance by removing small slivers of copper from the, initially oversized, lines. The $\lambda_g/2$ and $\lambda_g/4$ line lengths were measured, adopting the same method described in the previous example, but this time using an HP 8753D, 30kHz to 6GHz network analyser.

For the $\lambda_g/4$ line, a 50Ω surface mount load resistor was put on the end of the line to make sure that the line gave a 90° phase shift to traverse half way around the Smith chart. It ended on the centre line at the point: 200Ω±j0Ω.

Once the 90° $\lambda_g/4$ length was known, the desired characteristic impedance was easily found by placing a surface mount resistor at the distal end of the line and then adjusting the width to give the desired impedance at the generator end.

Problem Three. In the final instance I needed to match a 50Ω source to a 160Ω load at an operating frequency of 2.45GHz. This was achieved by using three low-impedance $\lambda_g/2$ sections, made from microstrip lines, to provide good isolation between the source and load.

There must be minimal standing waves on the lines, hence the VSWR must be kept as close as possible to unity. The solution adopted is shown in **Fig. 7**. You can see that a $\lambda_g/4$ impedance transformer first transforms the 50Ω source to match a load of 10Ω. This is followed by two 10Ω $\lambda_g/2$ sections, feeding a single 40Ω $\lambda_g/2$ section to transform the output impedance to the desired 160Ω.

For the impedance inverter, a line impedance equal to the geometric mean of the source and load impedance, i.e. √(50×10)=22.4Ω, was required.

The material used for this application was an RT/Duroid 6010, non-woven glass ceramic filled PTFE composite, whose dielectric constant at 10GHz is 10.2 ±0.25.

The substrate height found to be most suitable was 0.64mm (25 mil). The combination of high permittivity and small thickness meant that it was possible to fabricate low impedance lines using practically realisable line widths. The line lengths and widths required to achieve the necessary $\lambda_g/2$ sections and the $\lambda_g/4$ transformer are given in **Table 2**.

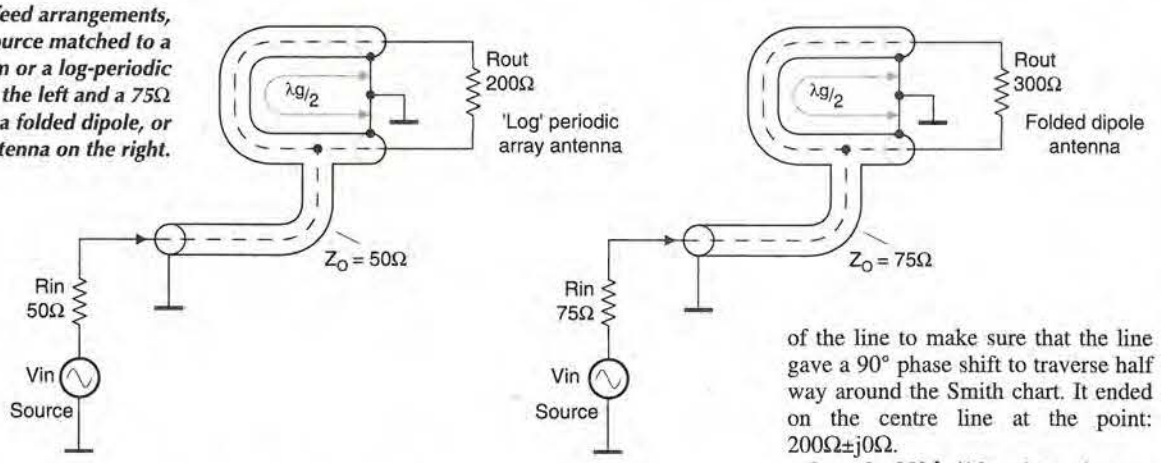

Fig. 3. Practical antenna feed arrangements, with a 50Ω transmitter source matched to a 200Ω helical-beam or a log-periodic antenna load on the left and a 75Ω transmitter matched to a folded dipole, or conical antenna on the right.

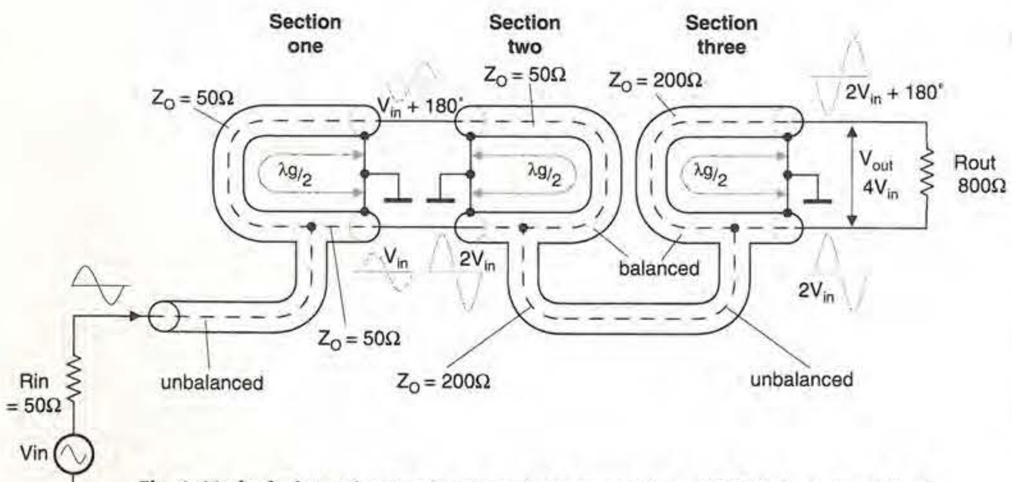

Fig. 4. Method of transforming low-impedance source to match high impedance load.

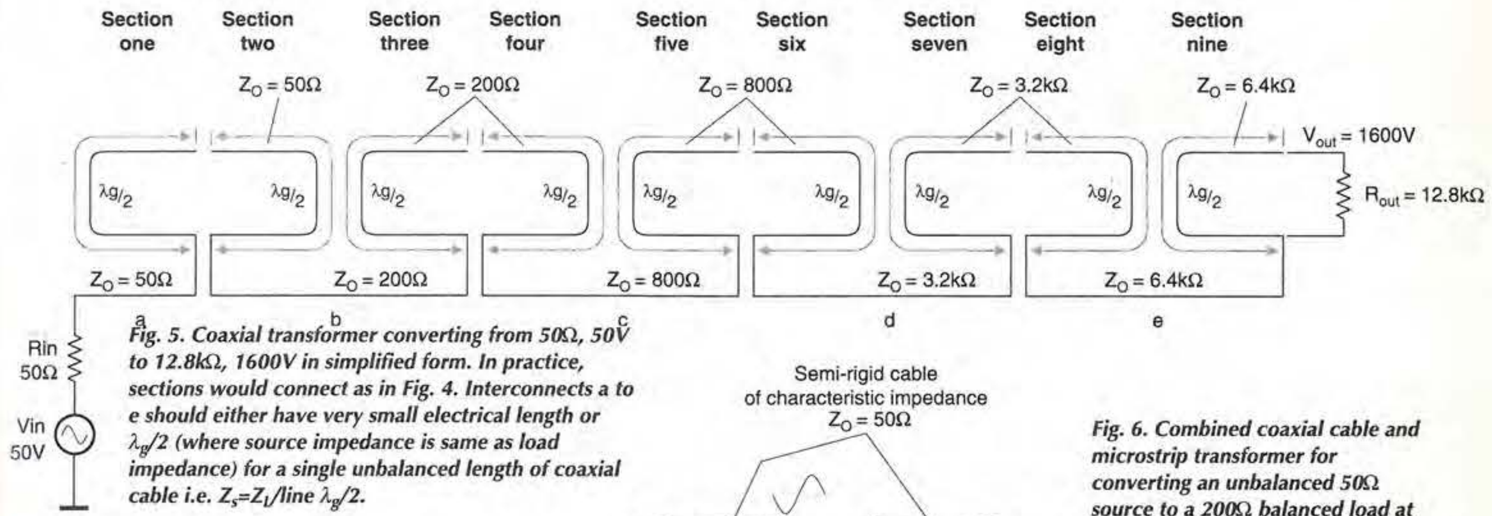

Fig. 5. Coaxial transformer converting from 50Ω, 50V to 12.8kΩ, 1600V in simplified form. In practice, sections would connect as in Fig. 4. Interconnects a to e should either have very small electrical length or $\lambda_g/2$ (where source impedance is same as load impedance) for a single unbalanced length of coaxial cable i.e. Z_s=Z_L/line $\lambda_g/2$.

Table 2. Microstrip line dimensions for low impedance matching/isolation circuit.

Line impedance (Ω)	22.35	10	40
Line width (mm)	2.25 88.5mil	6.17 243mil	0.89 35mil
Line length (mm)	λ_g=38	$\lambda_g/2$=19	$\lambda_g/4$=8

In summary

This article has shown that simple $\lambda_g/2$ co-axial and microstrip-line sections can be used to provide both voltage and impedance transformation at the VH and UH frequencies commonly used today in advanced communication systems.

The technique has been successfully adopted to provide the unbalanced to balanced condition required when connecting feed cables to various antenna structures. I have shown that the technique prevents the wave contained within the antenna feed cable from tending to 'spill over' the end and travel back over the surface of the cable, causing the outer sheath to have an RF voltage on it.

Finally, I have shown that these networks can be used to provide effective isolation between high power sources and their loads. ■

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2. Schoenbeck, R J, 'Electronic Communications – Modulation and Transmission,' Merrill, 1988, ISBN: 0-675-20473.
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5. Rogers Corporation Data sheet on RT/Duroid Microwave Laminates, Microwave Materials Division, 100 S. Roosevelt Avenue, Chandler, AZ, USA.

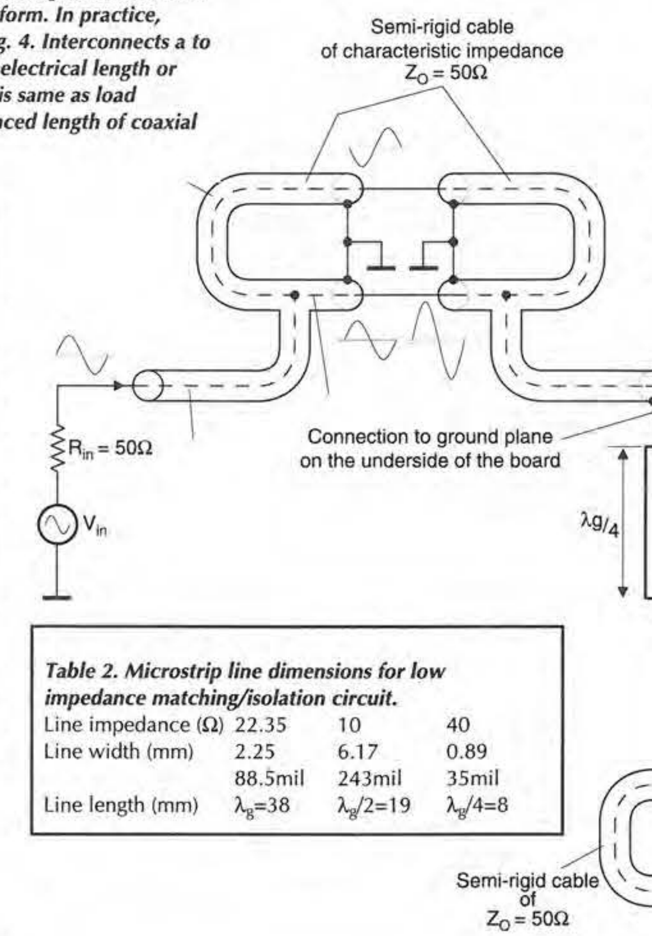

Fig. 6. Combined coaxial cable and microstrip transformer for converting an unbalanced 50Ω source to a 200Ω balanced load at 2.45GHz with good source to load isolation. It comprises three 50Ω $\lambda_g/2$ sections and a 100Ω $\lambda g/4$ impedance inverter.

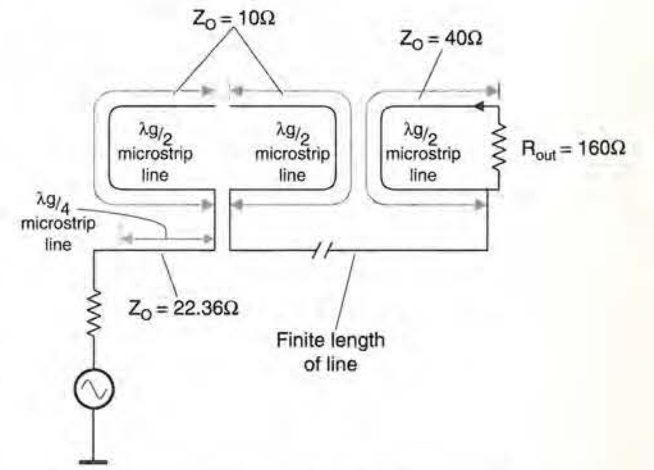

Fig. 7. Low-impedance three section transformer to provide good input/output isolation.

Make better Wien oscillators

On building two Wien-bridge oscillators operating at different frequencies but otherwise identical, a reader asked, "Why are the amplitudes different?" Ian Hickman answers the question, and gives tips on how to get the best out of the circuit.

I recently received a letter from a reader who was puzzled by an unexplained discrepancy. He had designed a two-tone generator for himself, the two audio tones being summed at the virtual earth input of an inverting op-amp stage. This is of course the correct way to do it, as it avoids the production of any intermodulation products due to interaction between the two sources, and switching either tone off has no effect on the level of the other.

The basic circuit

Each tone generator consisted of a Wien bridge oscillator, the basic circuit of which is shown in **Fig. 1**.

Assuming $R_1=R_2=R$ and $C_1=C_2=C$, the frequency is given by $f=1/2\pi RC$ and the attenuation from op-amp output back to the non-inverting input is a factor of 3, or 9.54dB. My correspondent said that he had seen many treatments deriving this figure, and with it the operating frequency, but none dealing with how to predict the amplitude of the output.

Figure 2 is a simplified version of the circuit of each of the two oscillators he used. Each could be set for one of two pre-set frequencies – switching arrangements not shown here – the output level being stabilised as usual with an R53 thermistor.

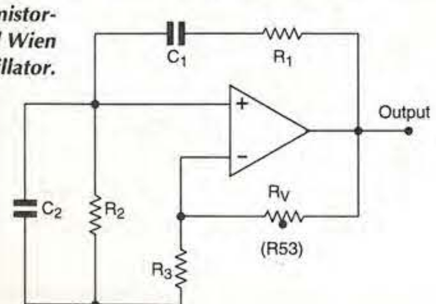

Fig. 1. Basic thermistor-stabilised Wien bridge oscillator.

The problem was that the output levels of the two oscillators differed, whereas given that the circuitry of each was identical, the designer had expected them to be the same.

Loop gain

The open-loop gain of the op-amp he was using is 50dB, i.e. around 300× at 1kHz, and greater still at the lowish audio frequencies used in his design.

So the differential input to the op-amp would be 0.3% or so of the output. Thus the input via the Wien network at the non-inverting input will be virtually identical to that at the inverting input via the thermistor branch. This amounts to one third of the output amplitude, i.e. the closed loop gain is ×3. As noted above, the frequency is (ideally) given by $f=1/2\pi RC$ so, for example, with components in the Wien network arm having exact values of 10kΩ, 0.1µ, the frequency would be 159Hz $(1/(2\pi)=0.159)$.

But note that a more exact expression for the operating frequency is,

$$f=\frac{1}{2\pi\sqrt{R_1R_2C_1C_2}}$$

So if C_1 is 10% low but C_2 is 10% high, say, the frequency will still be right. However, the attenuation in the Wien network is no longer 3:1, in fact it increases by 0.62dB. So the attenuation in the negative feedback arm via the thermistor will have to change also. This will require a different current through the thermistor, changing its dissipation and hence its temperature, to achieve the appropriate resistance. The change in current is brought about by a corresponding change in output voltage.

Setting the output frequency

Note that the adjust-frequency potentiometers in Fig. 2 affect only the series *CR* arm. For a Wien bridge oscillator, the frequency is given by $f=1/(2\pi RC)$ only if the two resistors R_1 and R_2 of Fig. 1 are equal, and the two capacitors likewise.

In all other cases, the full equation,

$$f=\frac{1}{2\pi\sqrt{R_1R_2C_1C_2}}$$

must be used.

With a potentiometer adjustment in the series arm only, as in the circuit shown, any adjustment to the frequency must necessarily alter the 3:1 nominal attenuation through the network, and hence also the output level. There are several ways round this, as discussed next.

Figure 3 shows an improved frequency adjustment circuit, having several advantages. If the track resistance of the potentiometer R_{tune} does not exceed 10% of the reactance of *C* at the operating frequency, adjusting it from end to end will result in negligible variation in output amplitude, while providing a ±9% tuning range.

This arrangement has a further advantage compared with Fig. 2. For a pre-set potentiometer, the sum of the resistance from one end of the track to the wiper, plus that from the wiper to the other end of the track, slightly exceeds the track end-to-end resistance. This is due to the wiper contact resistance, which is not negligible – except in the case of wire-wound potentiometers.

The wiper contact resistance is also slightly variable, both in the long term, contributing frequency drift, and in the short term, contributing noise. In the circuit of **Fig. 3**, the wiper contact resistance is in series with the very high input impedance of the op-amp, and therefore has virtually no effect.

Setting output amplitude

The drive to the thermistor will adjust itself to whatever is necessary, namely to set the attenuation in the negative feedback path to just marginally less than that in the positive feedback path. So including a potentiometer as part of the 270Ω at the inverting input of the op-amp, Fig. 2, will provide an amplitude adjustment, which can be used to set the desired output, after first setting the frequency.

The output amplitude of a circuit such as that shown is determined by the characteristics of the particular R53 thermistor used. The nominal room temperature resistance of this device is 5000Ω – or 5E3, hence the type number – with a ±20% tolerance.

The maximum recommended operating temperature of the thermistor pellet, in its evacuated glass envelope, is 220°C, and the device can be used, with suitable derating, at ambient temperatures up to 175°C. Power sensitivity is nominally 62.5°C/mW.

What value for R_3?

Maximum permitted dissipation at 20°C is 3mW. This is sufficient to drive the device's resistance typically down to 63Ω. I generally make R_3 about 120Ω, driving the thermistor resistance down to about 240Ω, and avoiding any possibility of over-dissipation. This results in an op-amp output amplitude of about 2.5 to 3.0V peak to peak.

Heavier drive to the thermistor can be beneficial in some respects, as discussed later. But with R_3 at 120Ω, the op-amp sees a load at its output of 360Ω. Op-amps are generally characterised with 10kΩ and/or 2kΩ loads; some types also handle a 600Ω load. Most op-amps will work happily with a 360Ω load in this application, as the required peak to peak output voltage and current swing is quite modest.

Current through the thermistor is of course the same as that through R_3. Thus, knowing the alternating voltage across the thermistor – circuit design should always be such as to avoid any DC in it – its dissipation is readily determined.

Don't forget that there is a considerable tolerance on R53 thermistors. Two different R53s taken at random from stock, in a circuit as Fig. 1 running at 110Hz with R_3=120Ω, gave outputs of 2.53V and 2.74V peak-to-peak.

Running the thermistor harder has some advantages. Most obviously, the hotter the thermistor, the less variation in output amplitude there will be, with change in ambient temperature. However, another advantage accrues at low frequencies.

In an audio oscillator covering 20Hz to 20kHz in the usual three ranges, operation at the lowest frequency may result in 'amplitude bounce'. The thermal time constant of the thermistor combines with that of the circuit to form an underdamped level-control loop. Consequently, any small adjustment in the frequency results in the amplitude oscillating up and down at about one cycle per second, gradually settling to a steady value.

With lower drive levels the effect is more pronounced, and in extreme cases the level control loop can become unstable. At higher output frequencies, the loop is well damped, and there is no problem.

Distortion

For a thermistor-stabilised Wien bridge audio oscillator, distortion is worst at the lowest frequency. For at 20Hz, the half period of the sine wave is no longer negligibly short compared with the thermal time constant of the thermistor pellet. Consequently the temperature of the latter tends to rise slightly at each peak of the waveform and fall at the zero crossings. This results in distortion, which is almost purely third harmonic.

Running the thermistor cooler – subject to the level loop stability constraints mentioned above – can reduce the effect. Thus the temperature variations are less, and so is the distortion. It is also possible, with a different thermistor-stabilised oscillator design, to outphase the third harmonic component, providing a very low distortion output right down to 20Hz.[1]

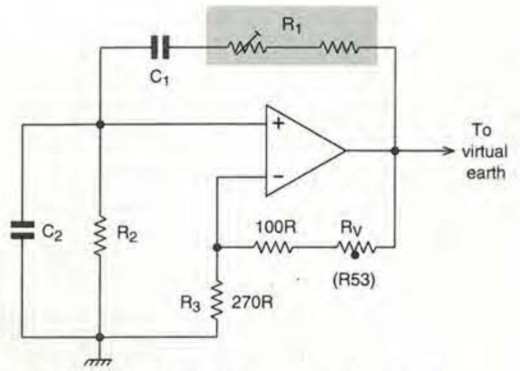

Fig. 2. Two circuits like this produced different output levels.

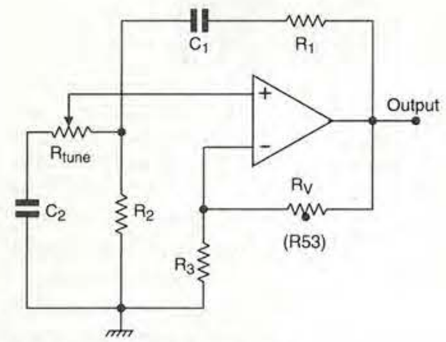

Fig. 3. A modification permitting ±9% tuning range, with negligible effect on output level.

In summary

Thermistor-stabilised Wien-bridge audio oscillators offer good, though not excellent, performance, from a simple and economical design. As always with circuit design, the theory will tell you what you should get, but what you actually get is down to those all-important tolerances.

The various measures outlined above permit the designer to cope with these, finishing up with a circuit tailored to the particular requirement. ■

Reference

1. Rosens, R, 'Phase shifting oscillator', *Wireless World*, Feb. 1992 pp. 38-41.

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LETTERS

Letters to "Electronics World" Quadrant House,
The Quadrant, Sutton, Surrey, SM2 5AS
e-mail jackie.lowe@rbi.co.uk using the subject heading 'Letters'.

Audiotherapy

May I congratulate you on your comment *'Aromatherapy and Audiophools.'* in the October 1999 issue. It needed saying.

I remember my good friend, the late James Moir, saying to me that people seem to prefer belief in magic rather than a solid scientific reality. Aromatherapy I can just about accept, as trace elements and compounds can interact with our body chemistry. But fancy talk I can only accept if our hi-fi amplifiers generate copious supersonic 'whiskers' along with their audio output.

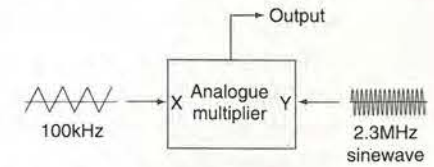
After all, our electrical engineers who manipulate thousands of kilowatts have found that Ohm's law satisfies their design problems; dielectric losses and skin effect only need to be considered at radio frequencies, and possibly in circuits carrying thousands of amps. Of course, if and when all our radio and audio circuits are going to use exclusively very short pulses at RF repetition rates we will have to redesign pretty well everything. Maybe the cable manufacturers are getting ready.

Meanwhile, as Andrew Emmerson says, it is all good for business!

Ralph West
Devillac
France

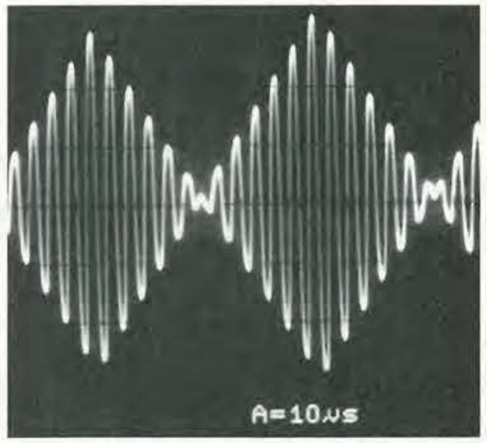
Mystery waveform

In reply to Nick Wheeler's letter about his mystery waveform in the letters pages of November's *Electronics World*, the waveform can be produced by multiplying a low-frequency triangular wave with a high frequency sine wave.



The diagram shows 'A=10μs' but does not show where A is, so I cannot be precise about frequencies. Looking at the diagram, the high frequency component is about 23 times the frequency of the low frequency component. If the low frequency triangular wave fundamental component was, say, 100kHz then the high frequency sine wave would need to be 2.3MHz. The waveform shown can be obtained then by multiplying the two in a suitable multiplier stage such as an analogue multiplier.

Peter Goodson
Bracknell
Berkshire

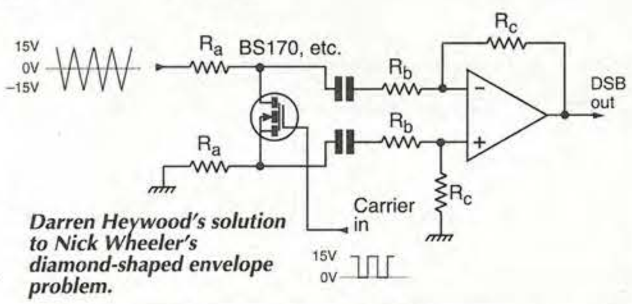


The block diagram shown is one answer to Nick Wheeler's question, "How did I produce the waveform shown in the photo?"

Nick Wheeler's wave is a double-sideband suppressed carrier, modulated by a sawtooth wave. It may be mistaken as an AM wave but if you home in on its zero crossing, it may be noticed that the wave undergoes a phase change (as far as I can see from the photograph) hence DSB.

I know of two ways of generating such a wave, the first involves the use of either one of two mixers. Alternatively, DSB may be generated using a discrete chopper amplifier but the limiting factor with this method is bandwidth restrictions of the op-amp and gain requirements. DSB is closely related to PAM, while DSB utilises both phases, PAM utilises just one phase.

Darren Heywood
Buckley
Flintshire



Darren Heywood's solution to Nick Wheeler's diamond-shaped envelope problem.

Musical numbers

None of the critics of the several telephone number changes seems to realise that any single step alternative would have produced immense chaos.

Any change had to consider that humans are very much creatures of habit and also have a limited ability to remember long numbers. In the UK and elsewhere, telephone companies have minimised the memory problem by using as few digits as possible and then systematically divided these necessary digits into smaller, logical separate, more easily remembered, parts.

The original national number system used '0' plus nine digits. Its potential maximum of up to 800 million numbers must have seemed more than adequate for the long-term needs of a country with only 60 million people. At that time even experts had a limited view of the potential of computer technology.

The new number range of '0' plus ten provides a tenfold increase in capacity. Placing the extra digit immediately after the initial '0' gives the simplest, easily remembered, standard number change combined with maximum future flexibility. Unfortunately the existing system had no unused first digits.

Any simple reuse of a digit automatically produces many new numbers, which more or less duplicate the first digits of different existing national numbers. Those who overlooked the change would produce, over several months, millions of no tones 'faults' and wrong numbers. There would be no simple way of interpreting and advising misdialled calls. The multiple step method avoided this problem. The first step, changing London to 071 and 081 freed the 01 code, which was given a recorded error message.

After a few years it was safe to use 01 instead of 0 to precede existing area codes. The old area codes were then given recorded error messages.

After a further few years people had forgotten previous area codes. Old codes commencing with 02, 03 etc. became available for other purposes. London could again have a single national code, 020, and its local numbers finally increased to the eight digits planned to several years previously.

Why has BT not explained the underlying reasons to the public? In my experience, when people know all the facts they are much less likely to complain.

Overall, apart from poor quality publicity, BT seems to have found an optimum solution to the problem. It would be interesting to hear from any critic who, considering all aspects, could suggest a better one.

RG Silson
Tring
Hertfordshire

Contrary to Andrew Emmerson's editorial of December 1999, the change of numbers is not due to competing operators. It is due to the need to correct a mistake made in 1995 when an extra digit was added to the codes of all UK numbers.

The extra digit should have been added to the end of the subscriber's telephone numbers rather than the beginning as it was.

The extra digit at the front did increase the possible numbers by a factor of ten, however it did not increase the capacity for subscribers numbers as next-door neighbours cannot have different codes – except at area boundaries of course.

The mistake is now being corrected by reducing the area codes to three numbers and moving the digit along to the individual subscribers number. ■

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With battery-operated and power-efficient circuits in mind, International Rectifier has recently launched a new Schottky power rectifier diode – the 10MQ100N – that exhibits a typical voltage drop of only 0.68V when carrying 1.5A peak.

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The SMA-packaged 10MQ100N is primarily intended for high-density surface-mount applications, but it serves equally well working between two tracks on the reverse side of a through-hole board.

For a full data sheet and applications ideas, visit International Rectifier's web site, www.irf.com.

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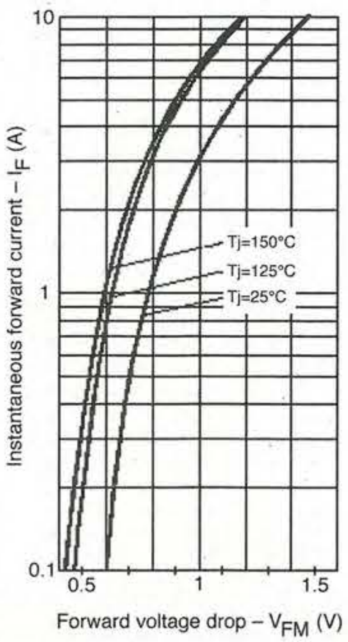

Maximum forward voltage drop of the 10MQ100N.

Designers with no Internet access, contact International Rectifier European Response Centre DBI, 31 Chelsea Wharf, 15 Lots Road, London, SW10 0QJ

Major ratings and characteristics

Characteristics	Rating
Forward current, $I_{F(DC)}$	2.1A
Reverse voltage, V_{RRM}	100V
Peak forward current, I_{FSM}	120A
Forward voltage at 1.5Apk, Tj=125°C	0.68V
Junction temperature	–40 to 150°C
Typical junction capacitance, C_T	38pF
Maximum thermal resistance	160°C/W

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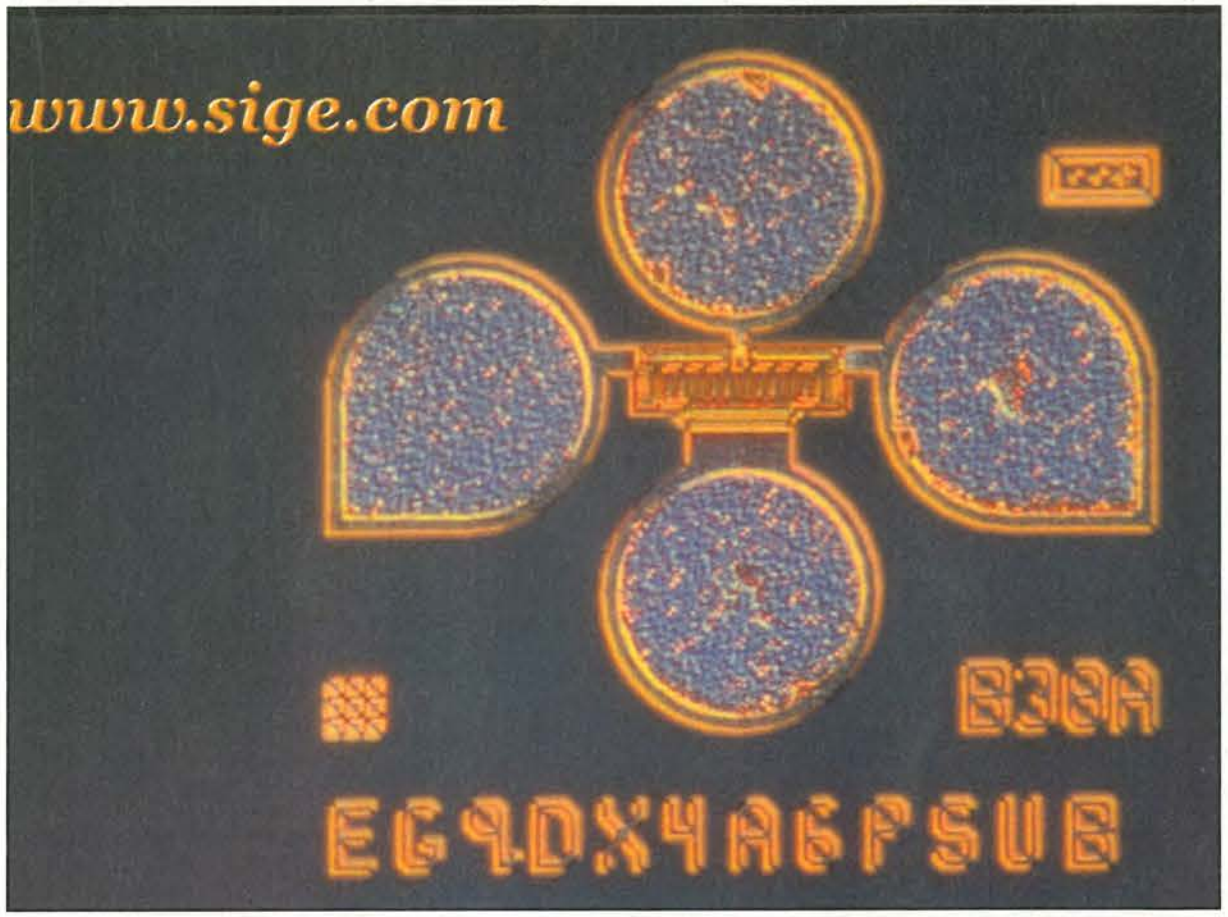


SiGe Microsystems claims that over 12 companies have bought its low phase noise transistors for local multi-point distribution services (LMDS), used for broadband access to the home such as cable. Lower phase noise reduces bit errors and allows for increased data rates, the firm claims. The transistors can be used to build oscillators running at up to 20GHz.

GaAs takes on SiGe

The once out of favour GaAs technology has given semiconductor makers plenty to laugh about lately, but the time has come when the major design houses are looking elsewhere. So does SiGe fit the bill to be the key high frequency technology for the Year 2000? Richard Ball examines the facts.

The explosion in the use of wireless products such as mobile phones in the last couple of years has left gallium arsenide, or GaAs, semiconductor manufacturers laughing all the way to the bank.

At last it seems that GaAs – a much maligned semiconductor technology – has found favour. However, manufacturers and design houses realise its lifetime and opportunity are finite, and that silicon germanium, SiGe – and perhaps even bulk CMOS – will ultimately provide a better fit in large volume markets.

Major GaAs design houses and manufacturers, including Anadigics, M/A-COM, RF Micro Devices, TriQuint and Vitesse are starting to use SiGe.

During September two firms, M/A-COM and Anadigics signed deals with Temic, the Atmel division that manufacturers SiGe chips.

Using SiGe, Temic introduced a DECT RF chipset last year and GSM power amplifiers this year. It started shipping in high volume from its 6in wafer fab in Germany early this year.

What is significant about both Temic deals is they bring together established RF component experts with Temic's proven SiGe process technology. As Dr Charles Huang, chief technical officer for Anadigics describes the move: "Having access to Temic Semiconductors' SiGe facility and technology provides us with an opportunity to complement our existing gallium-arsenide and silicon programmes."

US firm Anadigics has been a long time proponent of GaAs, but introduced its first silicon part, a dual frequency synthesiser, in June. The Temic deal continues the shift away from pure GaAs development.

Ron Michels, v-p of the cable and broadcast segment at Anadigics says: "Anadigics is committed to identifying, developing, and exploiting new process technologies to provide leading RF solutions for our customers."

M/A-COM, another GaAs stalwart, will use Temic's process for devices targeting both wired and wireless infrastructure especially gigabit data rate LANs and the local loop.

"Both companies agree this joint partnership is an excellent opportunity to produce silicon germanium products for the rapidly changing demands of the telecoms marketplace," says Rick Hess, a v-p at M/A-COM.

RF Micro Devices, a spin off from Analog Devices, is another experienced high-frequency design firm choosing to use SiGe. It has chosen to link up with IBM Microelectronics which has one of the most advanced SiGe research and manufacturing programmes. IBM was the first firm to introduce high volume SiGe chips this time last year.

Already using silicon, GaAs and heterojunction bipolar transistors, RF Micro Devices' access to SiGe will fill out its capabilities nicely, allowing it to replace GaAs in many areas.

Why SiGe, not GaAs?

SiGe offers the economies of scale that you get with conventional silicon processing. Producing die on 8in silicon wafers gives much cheaper parts than from smaller GaAs wafers.

Weighed up against this is the claim by GaAs makers that their technology is cheaper in terms of mask generation.

"In semiconductor manufacturing the cost is primarily capital cost, and the amount of equipment you need relates to the number of masking levels. Our four-level metal process is implemented with 15 masks; bipolar SiGe requires 25 to 30 masks," claims Vitesse's Chris Gardner.

But, when it comes to the crunch, GaAs just uses too much power to be used for anything beyond the low integration parts such as power amplifiers – and SiGe can now even be used for that function. With the trend towards further integration, power consumption – particularly that due to leakage current – is critical. And this is where SiGe triumphs, even over silicon.

The firms plan to develop third-generation mobile phone chips. The belief is that using SiGe in the front end will reduce power consumption and device count.

"IBM's SiGe technology will enable us to design products that deliver high performance and low power consumption in wireless communications applications," says William Pratt, chief technical officer for RF Micro Devices.

Vitesse is one of the biggest names in GaAs – the first to offer commercial parts – but even it has succumbed to the reality that SiGe is the way forward.

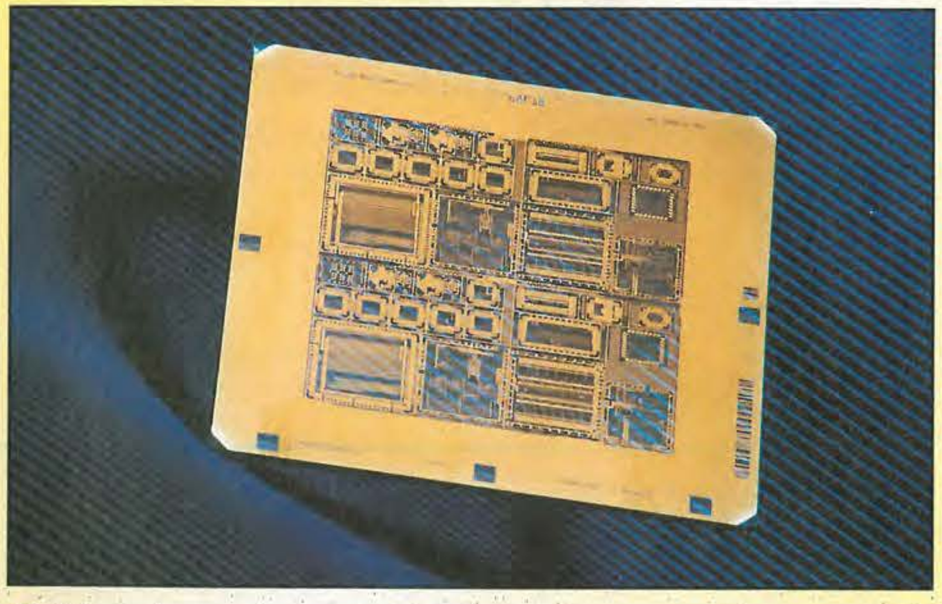
"We no longer consider ourselves technology snobs; we can also write cheques," says Chris Gardner, v-p and general manager of the telecoms division at Vitesse. "If we need SiGe we know the guys at IBM and we can write them a cheque."

Moreover, during the past year, Vitesse has bought three firms that deal in silicon. These include designers of Sonet/SDH, ATM, fibre channel and network processor chips. SiGe would offer excellent power savings in many of these devices.

TriQuint is another name heavily associated with GaAs, producing RF and microwave modules for mobile phones, base-stations and satellite systems. The company is reported to be looking at silicon development, where higher integration makes GaAs too expensive.

Whether GaAs is left to a few niche applications, or whether new high volume applications come along that use frequencies that SiGe cannot reach, remains to be seen.

But for now, the shift towards SiGe, and away from GaAs, is well and truly under way. GaAs suppliers will be hoping that SiGe is the joker in the pack, and keeps the laughter flowing during trips to the bank. ■



Multi-product wafer service for SiGe

Austria Mikro Systeme, a long time developer of silicon germanium technology, is offering a multi-product wafer (MPW) service for SiGe. Starting in February, AMS will run seven lots of wafers through its fab, the final run in December next year. Untested packaged samples are provided after 12 weeks for the SiGe process. The 0.8μm process is based on the firm's mixed signal CMOS process and adds a heterojunction bipolar transistor module.

Tanδ capacitor tester

...continued from page 20

So just what level of charged capacitor voltage can be absorbed by the meter without damage resulting?

As with a fuse, the fusible resistors can become open circuit either by a large peak current pulse of short duration or a much smaller current for a longer time. The peak current that flows depends on the voltage on the charged capacitor while its duration depends on capacitance value. The maximum charged capacitor voltage that the meter can withstand varies according to the capacitance value.

It is best to assume the maximum safe voltage is zero, using the meter to measure only discharged capacitors.

If sufficient charged capacitor voltage is applied, the ensuing peak current can open circuit the 22Ω and 2.2Ω fusible current sense resistors, $R_{7,8}$, and permanently weld the range relay contacts closed.

Damage to these three components has been confirmed by practical experiments using a prototype meter. Replacing these components restored normal operation.

Using all the above protection measures, a capacitor charged to a very high voltage might store sufficient charge that those integrated circuits most directly connected to the test capacitor are damaged.

To facilitate test and repair of a damaged meter, integrated circuits $U_{1,3\&8}$ are mounted in low profile, turned pin sockets.

Battery power supply

Generating a +5V stabilised supply from four AA batteries poses the difficulty that with fresh batteries, the circuit must reduce the battery voltage. As the batteries discharge, the circuit must automatically changeover to boosting the battery voltage.

For economy of batteries it is desirable to support an auto-off timer to shut down the power supply, and disconnect the battery load. A low battery indication as the batteries become exhausted is essential.

For ease of assembly, I decided to avoid surface mounted components, using only DIL integrated circuit packages. This was not a problem with the measurement circuits or the 100Hz generator. Unfortunately almost all the latest power supply circuits are only available in minute surface-mount packages.

A secondary problem lay in the 'shutdown' terminology. Most of the power supply circuits I looked at claimed a shutdown facility. In many instances, while the power supply chip itself shut down, it did not disconnect the battery load. The battery continued to supply voltage and current into the tanδ meter.

Having tried and rejected a number of power supply circuits, I found a Linear Technology application note titled 'Micropower Buck/Boost Circuits'.[5] This application note discussed the above four cells to 5V power supply problems and suggested a solution.

This used the LT1303CN8 eight-pin DIL integrated circuit as a boost converter, followed by a discrete linear post regulator. I had previously rejected this IC because when used conventionally, its shut-down circuits do not disconnect the battery load. Battery disconnection is achieved using a post regulator circuit.

This application also claimed a near 80% conversion efficiency. It used surface-mounted diodes, inductors and capacitors, but I wanted to use leaded components.

With a 100Ω resistive load to simulate the tanδ meter, bread-boarded, this prototype supply looked good. It provided a stable output voltage with inputs from 3 to 6 volts, using my bench power supply. When shut down, no power flowed in this load resistor.

I etched a trial PCB and used the circuit with four AA batteries to power a prototype meter. Disaster; switching noise from the supply interfered with the accurate zero crossing detection of the 100Hz test current waveform. Power conversion efficiency was less than 65%.

My original six AA battery prototype power supply used a MAX883

Recognising a damaged tanδ meter

A functioning tanδ meter produces around 150mV at 100Hz across its test probes. If the test probes of the meter are touched together, some 50mA of test current flows, the meter displays over-range and the high range LED lights.

Illumination of the high range LED indicates that more than 8mA is flowing in the circuit.

Note that the current actually measured at the test leads depends on the burden of the DMM used. One of my meters measured less than 25mA because of its high AC current burden. A better meter measured more than 35mA.

With R_7 or R_8 open circuited, only a small test current can flow, the high range LED cannot light and the above voltages and currents will not be measured.

An ohmmeter measurement of resistors R_7 and R_8 will confirm they have been damaged.

If the PM128 meter in a tanδ meter displays abnormal behaviour, the above simple tests should confirm normal operation or that the meter has been damaged.

If damage is indicated, R_7, R_8 and the range relay should all be replaced.

Repeat the above test also check visually for other damage, before re-calibrating the meter.

low drop-out linear regulator. This produced little noise and no notable switching transients.

The MAX665CPA on-board –5V converter also produces switching transients. These had not troubled my linear regulator but interfered with the LT1303CN8 control circuits.

The LT1303CN8 performed much better using my bench supply than when using four AA batteries, whether NiCd or alkaline. Efficiency was improved and with less noise.

Apart from using leaded and not surface mounted parts, I had followed the suggested component values. After initial unsuccessful attempts to add filter capacitance, I decided I needed to more closely replicate the performance of the specified surface mounted parts.

Comparative high-frequency measurements suggested two problems. I needed more and lower ESR capacitors, together with a larger value switching inductor, and a better power supply PCB layout. The 5V power supply board was redrawn to minimise ground impedances. Capacitor decoupling was improved by using four terminal connection paths, **Fig. 3**.

Of the capacitors I measured, Oscon types performed well but were expensive and physically much larger than the PCB space I had available. Certain low-ESR capacitors failed to meet their claims. Others were too tall. In the end I chose the lowest cost of those measured, the Rubycon YXF. These performed far better than claimed and fitted into space available.

Using these capacitors in the new printed board layout, I wound and tested various inductors. Doubling the original 22μH value to 47μH, improved conversion efficiency and reduced switching noise.

To further reduce +5V supply noise and isolate the –5V converter's noise from the LT1303CN8 control circuit, I added an extra capacitor, decoupled using a small VHF inductor. Connection of this +5V supply to the main boards is via a bucking mode, bifilar-wound, twelve-turn inductor, wound on a small high permeability toroidal core, **Fig. 4**.

The –5V supply uses a MAX665 eight pin DIL, DC-to-DC converter in high-frequency mode, to minimise its switching noise. To further reduce high-frequency noise from this –5V supply, I added two extra capacitors, decoupled using a second small 4.7μH inductor.

In total four 220μF 10V, low-ESR decoupling capacitors are used for the +5V supply and three for the –5V converter. These changes ensure very low supply line noise levels on both the analogue and logic boards.

In addition to reducing noise from the power supply, these changes much improved the conversion efficiency, achieving around 80% with 4.5 to 5V input. This is the median voltage when using four AA alkaline or Ni-Cad cells.

Capacitor quality – tanδ

The quality of many high-frequency components, such as RF inductors and very-low-loss capacitors, is often defined by their 'Q' factor. Q is the result of dividing a components measured AC reactance by its AC resistive losses.

The reciprocal of 'Q' is tanδ, which is defined as the capacitors AC resistive losses, i.e. ESR, divided by its capacitive reactance, X_C, at that frequency.[6]

TanΩ is used to describe the quality of almost all general purpose capacitors.

$$\tan\delta = \frac{ESR}{X_c}$$

where,

$$X_c = \frac{1}{2\pi fC}$$

Alternatively $ESR = X_c \tan\delta$ and $\tan\delta = ESR \times 2\pi fC$.

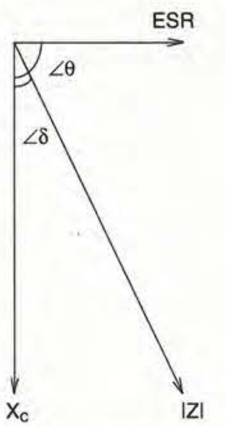
As you can see, tanδ has no upper limit. It can – and frequently does – exceed unity. Particularly at high frequency, *ESR* can exceed the capacitors reactance, X_C.

Tanδ is sometimes confused with the older term 'power factor'. Power factor is defined as the cos(90–δ), which is the same as sinδ or the cosθ.

Hence power factor is,

$$\cos\theta = \frac{ESR}{|Z|}$$

where |*Z*| is the capacitor's measured impedance magnitude. For aluminium electrolytic capacitors, |*Z*| is always much greater than X_C.



For very small angles, tanδ and cosθ have similar values, but a quick pocket calculator check will confirm that they diverge quickly. As the loss angle increases cosθ can never exceed unity, but tanδ has no upper limit.

All capacitors exhibit a variety of losses including a small DC leakage current, resistance of their metallic conducting surfaces, together with resistive dielectric losses. Combined, these dissipate some of the applied energy as heat.

Such losses reduce the theoretical 90° phase difference between the applied current and the capacitors voltage. At 1kHz for example, the phase angle of a typical 1000μF 25V radial-lead electrolytic capacitor measured 67° – substantially less than the theoretical 90° of phase.

This phase angle could be reproduced in a circuit by using a high value resistor in parallel with the capacitor, or a much smaller value resistor in series with the capacitor. Aluminium electrolytic capacitor specifications use this equivalent resistor in series with the capacitor.

The series resistance for the above 1000μF 25V capacitor was measured at 1kHz and found to be 71mΩ, X_C was 169mΩ and tanδ was 0.42.

Measured at 100Hz, the series resistance of this capacitor was 104mΩ, X_C was 1.62Ω and tanδ was 0.064.

This series resistance is the equivalent-series resistance, or *ESR*, of the capacitor.[7]

A capacitor's reactance reduces in proportion to its actual capacitance value and not its specified or low frequency value and measurement frequency.

With increasing frequency, the actual measured capacitance value of an aluminium electrolytic capacitor is much reduced. I tested a 100μF capacitor that measured less than 50μF at 100kHz.

Being a combination of fixed and variable losses, ESR also reduces with frequency, but to a lesser extent. Having reached a minimum value, ESR then increases at higher frequency.

The measured tanδ of an aluminium electrolytic capacitor depends on measurement frequency. Tanδ for aluminium electrolytic capacitors increases with increase of frequency. Tanδ has no upper limit and can exceed unity, especially for a worn out capacitor, **Table 1**.

Aluminium electrolytic capacitor tanδ increases rapidly as the capacitor wears out. It provides a sensitive, easily interpreted measurement. At 100Hz, tanδ for typical good board-mounted capacitors should be less than 0.1. Devices with a tanδ of more than 0.2 should be replaced to ensure ongoing and reliable equipment operation. Capacitors having a tanδ or more than 0.4 are worn out.[1]

Table 1a). Typical tanδ values of new stock capacitors measured at 100Hz – low capacitance values.

Capacitor	1μF	2.2μF	4.7μF	10μF	22μF	47μF	100μF
50V bipolar Al.	0.05	0.05	0.05	0.05	0.05	0.05	0.06
63V polar Al.	0.04	0.04	0.035	0.035	0.035	0.045	0.04
450V polar Al.	0.1	0.1	0.08	0.05	0.05	0.05	

Table 1b). Typical tanδ values of new stock capacitors measured at 100Hz – high capacitance values.

Capacitor	1000μF	2200μF	4700μF	10 000μF
25V polar Al.	0.06	0.075	0.09	0.1
63V polar Al.	0.03	0.05	0.06	0.07

Auto-off timer

This device that turns the meter off automatically after a delay is an ICM7242. It is essentially a 7555 timer combined with a divide-by-2 and divide-by-128 counters. This arrangement permits very long time periods using only modest timing component values. A small sub-board is configured as a universal stand-alone timer.

One gate of an HEF4093B ensures that the timer counters are reset, and hence trigger correctly, when power is

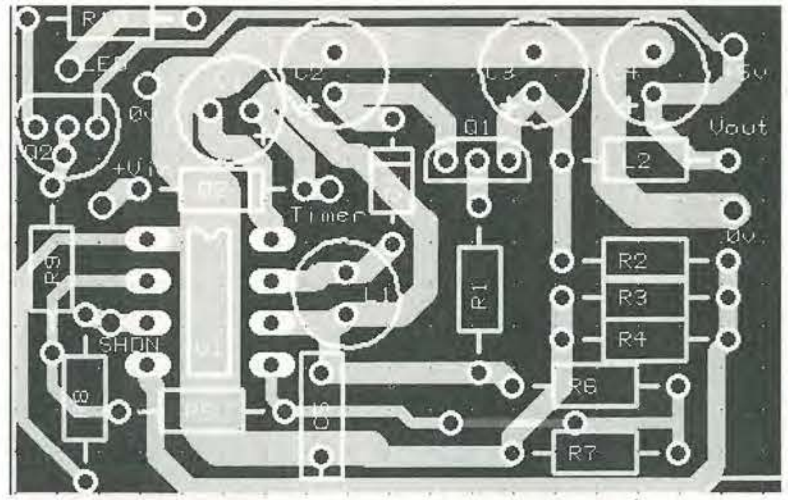


Fig. 3. Final design of the small sub-board used for the four AA batteries to +5V power supply. Note the four-terminal capacitor connections now used. The 220μF capacitors are placed along one board edge, to fit between the battery box and the PM128 display meter.

Parts and assembled meters

Cyril can supply kits or fully-assembled meters together with a bound 24 page user manual. This includes schematic drawings, board layouts, calibration instructions and parts list. This manual is supplied with each pre-assembled tanδ meter. In addition, for those who purchase printed boards or kits of parts, a further 22-page assembly manual with full instructions and drawings is included.

A tanδ reference capacitor, namely a 10μF ±5% 63V metallised PET type manufactured by Evox-Rifa and designated MMK22.5 106K63L4 is also available. E-mail Cyril at cyrilb@ibm.net or write to him via the editorial offices, marking 'Tan-delta meter' clearly on your envelope.

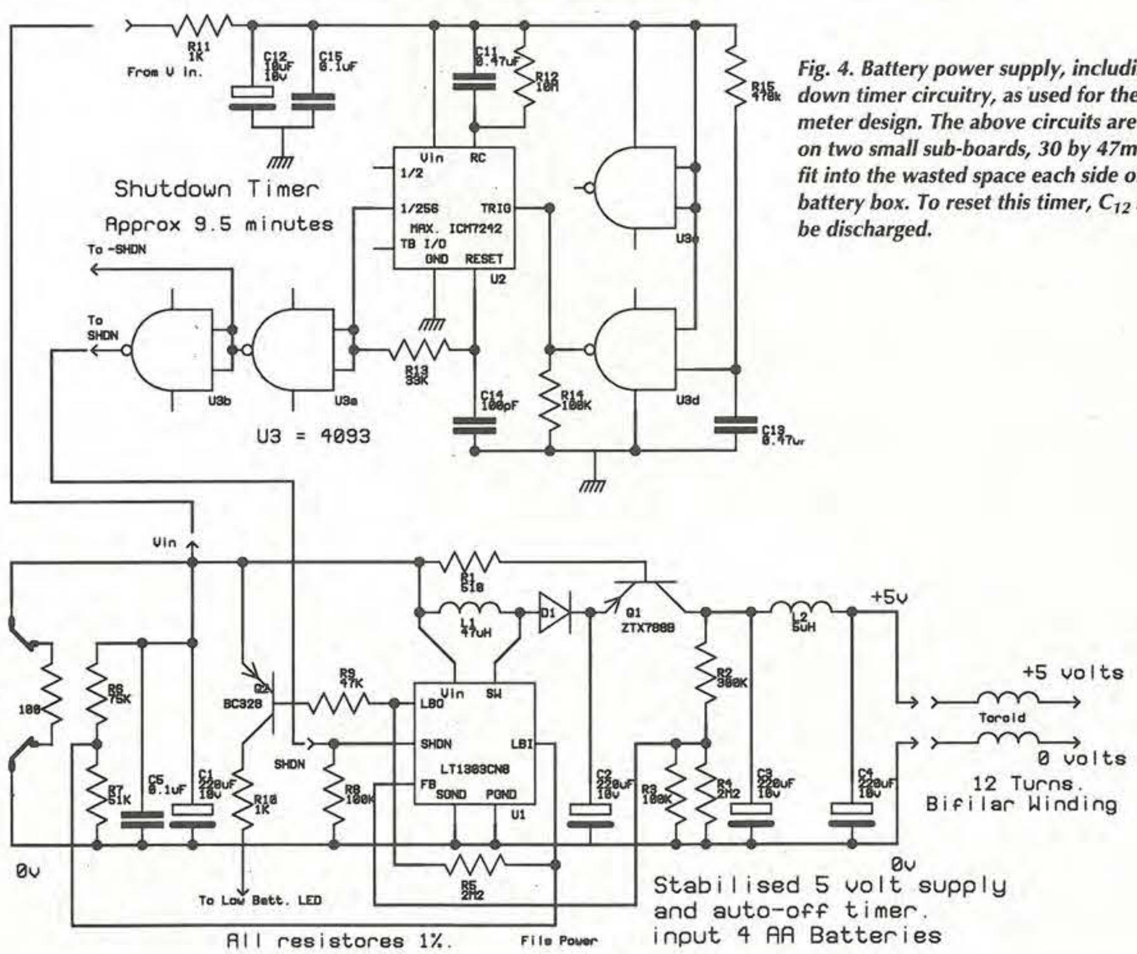


Fig. 4. Battery power supply, including shut-down timer circuitry, as used for the final tanδ meter design. The above circuits are contained on two small sub-boards, 30 by 47mm, which fit into the wasted space each side of the OKW battery box. To reset this timer, C_{12} must first be discharged.

Tanδ meter measuring tanδ of a capacitor in-situ on a pcb.

applied. The LT1303CN8 requires its shut-down pin be pulled high to turn the power supply off. Since other power-supply circuits shut down when pulled low, two gates of the HEF4093B are used. One buffers the other to invert the ICM7242 timed output.

Circuit design revisions

The final measurement circuit schematic is little changed from that published in *Electronics World* June 1999.[3] Apart from the additional INA118 in-amp already mentioned, the most significant change concerned the 1nF and 10nF capacitors. These were originally specified as 1% polystyrene or polypropylene to avoid all possibility of dielectric absorption affecting performance, **Fig. 2a**).

Following careful comparison tests, I found that with the small circuit voltages involved, 250V or 400V rated 5%-tolerance Philips 470 series metallised polyester capacitors, used with 1% metal film resistors performed equally well.

The gains for the two INA118 instrumentation amplifiers and the meter attenuator resistors have been revised. The logic drive level to the sample and hold stage has been adjusted. These changes optimise the linearity of the sample and hold circuits, **Fig. 2b**).

Some component values have been revised in the 100Hz generator to ensure repeatability and assist selection of the five 0.5% capacitance values needed. I find I can usually select these to better than 0.5% from the quantity of 0.1μF and 10nF capacitors required to build a meter. Used with 0.5% E48 value resistors, these capacitance values ensure the specified 100Hz generator output voltage and 100Hz frequency.

Main PCB design

Accommodating the various circuit updates has required a total of six printed board iterations. The bottom board that houses all the added protection components is now quite densely populated. The range relay had to be changed from a DIL to a SIL version to make more space available.

I have now assembled a good number of tanδ meters and all have performed to specification using the calibration adjustments. ■

How the tanδ meter works

The circuit works by sampling the voltage developed across the capacitor terminals at two discrete time intervals which are separated by exactly 90° of the test waveform.

The voltages developed across the capacitor terminals and the range resistors $R_{7,8}$ are first amplified to a usable level using two Burr-Brown INA118 instrument amplifiers.

The 'R' channel is sampled coincident with the peak of the capacitor's current waveform. The 'X' channel is sampled exactly 90° later.

Each channel voltage is sampled using one of two sample-and-hold integrated circuits. The voltage output from each sample and hold is then ratioed by the modified display meter and the resultant tanδ value is displayed, **Fig. 5**.[3]

Both sampling times are controlled by the circuits 'Logic' channel. This channel's trigger voltage is initiated by sensing the phase of the capacitor current waveform using a comparator to detect the zero crossing of this current. Comparator output is frequency doubled, then halved using a CMOS 4046 phase-locked loop and 4018 divider. This results in two equal mark:space ratio square waves. One is at 100Hz and phase locked to the test capacitor current waveform. The second at 200Hz having rising edges coincident with the 100Hz waveform transitions.

Application of some decoding logic provides the triggering needed, to generate the two 80μs-wide sampling pulses used to control the sampling of the 'R' and 'X' channels.

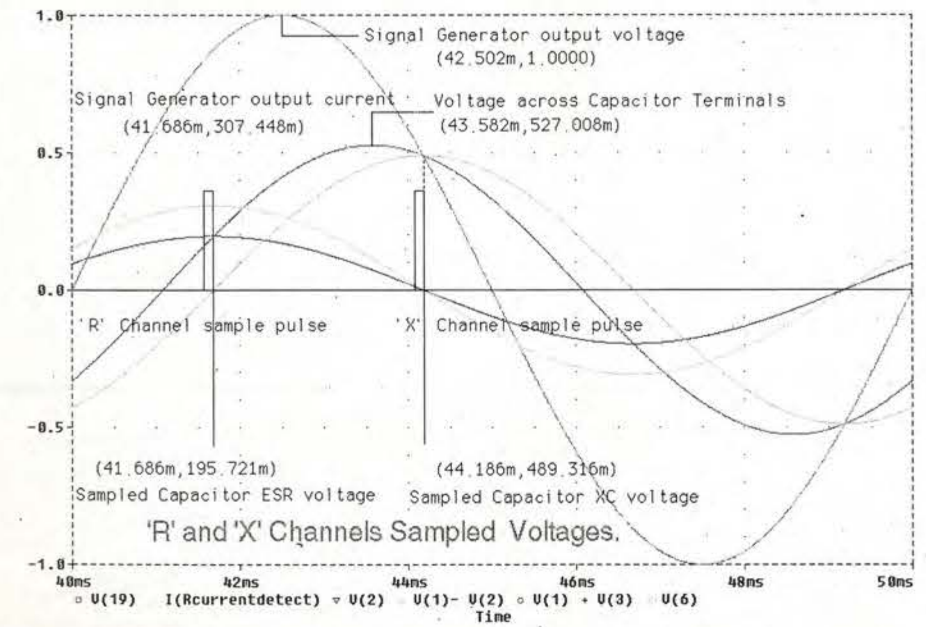


Fig. 5. Sample-and-hold control pulses from the logic circuit, superimposed onto the test capacitor's current and voltage waveforms. The phase angle between the capacitor current and the voltage developed across its reactance remains precisely at 90°, or 2.5ms, at 100Hz. The meter design relies on this relationship.

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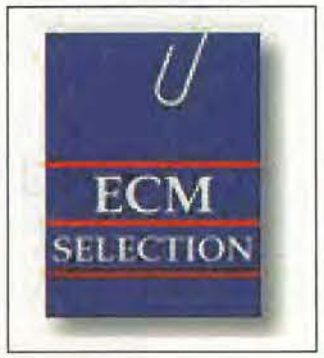

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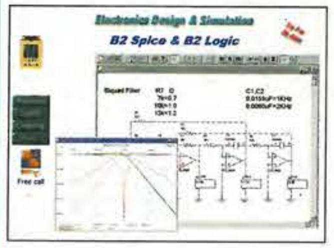

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Hands-on Internet

Using OTAs

Transconductance op-amps have properties that make them particularly attractive for applications involving high frequencies and bandwidths. So why aren't they used more often? From Cyril Bateman's searches on the Internet, it seems that we use them more often than we think.

In the October and December issues, I discussed circuits involving operational transconductance amplifiers, or OTAs. The particular schematics I highlighted demonstrated particular design functions that would be more difficult to implement using traditional voltage-output op-amps.

While similar to a conventional op-amp, a transconductance amplifier accepts a voltage input to produce a current output. Its gain can be set by altering its transconductance g_m or its load resistance.

An OTA can perform most of the circuit functions that are possible with normal op-amps, and makes some functions easier to implement. For example, a differential input instrumentation amplifier with good common-mode rejection can be made without needing the carefully matched resistor networks found in the conventional three op-amp circuit. It was this property that first prompted my investigation.

A transconductance amplifier offers other interesting properties. Used to drive rectifying diodes or optocouplers, it can provide a much increased working bandwidth, compared to that using a voltage-output amplifier. Using a MAX435, the *Universitat Politecnica de Catalunya, at Barcelonia* increased the normal 80kHz bandwidth of a 4N25 optocoupler up to 250kHz[2].

When first introduced in 1969, some writers expected the OTA could even become the dominant technology[3]. Consequently one might expect that a variety of transconductance amplifier integrated circuits would be commercially available by now, but from memory I found I could name only a few part numbers.

Searching the Taiwan Global Datasheets site[4] for 'transconductance' revealed only a short listing of devices. I repeated this search at AltaVista, Google and Northern Light as well as those semiconductor makers sites, who I knew manufactured transconductance amplifiers, with similar results.

One possible reason for this lack of components is that many integrated circuits that incorporate transconductance amplifiers, are not classed as such. In some instances this is because the output from the transconductance stage is not externally accessible, in others because the final amplifier output is voltage not current.

Here I review three such examples, each used to design a 10 MHz bandwidth amplifier.

Video difference amplifier

In the October issue I briefly mentioned the AD830 from Analog Devices[5]. This circuit is described as a high speed video difference amplifier and is specified for both ±5V and ±15V supplies. It provides extremely good signal processing from DC to 10MHz and above, producing a 60dB common mode rejection at 4MHz.

This design uses two identical transconductance input amplifiers, having a fixed g_m. Their output currents are summed in their common load, which is a capacitor shunted with a resistor. One amplifier acts as the signal input channel, the other the feedback signal channel. Being identical, their functions can be exchanged compared to the figure, with no change in performance. Simply apply the signal to pins 3 and 4 and feedback to pins 1 and 2.

The final unity gain output amplifier, simply follows the summed voltage which develops across this capacitor. When powered from a ±5V supply or a single 12V supply as in the figure, at least ±3.2V drive into 150Ω, is guaranteed. As can be seen, this AC coupled circuit provides a superb 10MHz performance, **Fig. 1**.

High-frequency amplifier

The LT1228 integrated circuit from Linear Technology[6] is described as a 100MHz current feedback amplifier with gain control.

Signal amplification again passes through two distinct circuit stages. The first is a transconductance stage having its gain or g_m controlled by an external current, in similar fashion to the CA3080 illustrated in my October article. This transconductance amplifier produces a 75MHz peak bandwidth.

The second stage is a wide band current feedback amplifier designed to drive low impedance, coaxial cable loads. With a 100MHz bandwidth and slew rate of 1000V/µs, this output stage is well matched to the transconductance stage.

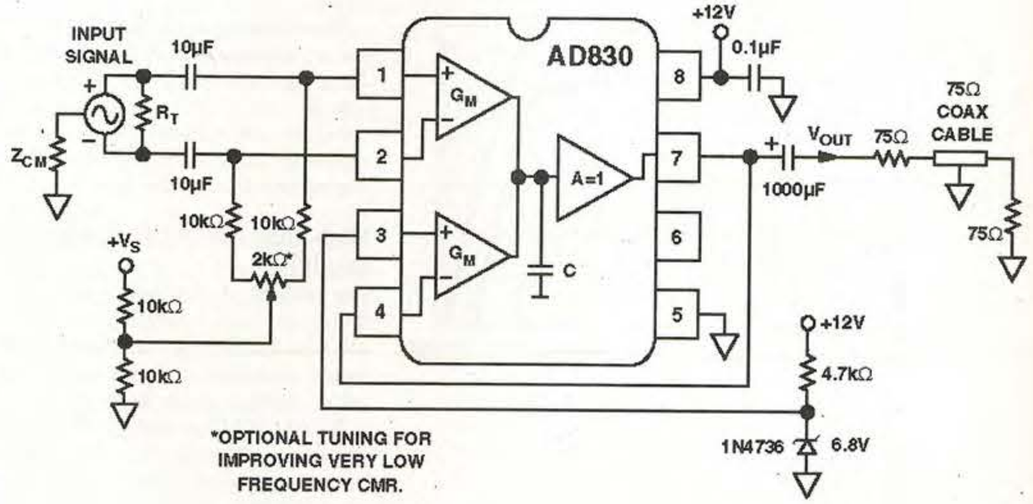

Fig.1 Schematic circuit of an AC coupled, differential line receiver, using Analog Devices AD830 integrated circuit. As can be seen this 'Video Amplifier' uses two OTA internally, with a unity gain voltage buffer as final output.

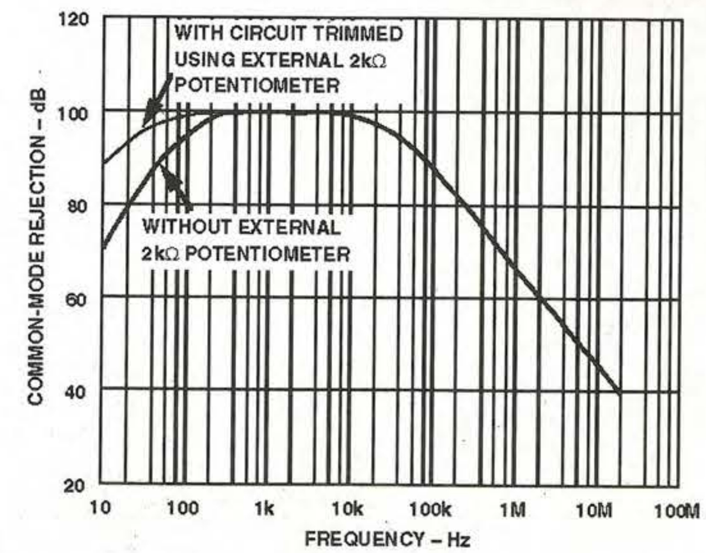

Fig. 1a). Using conventional components, this level of common-mode rejection would be almost impossible.

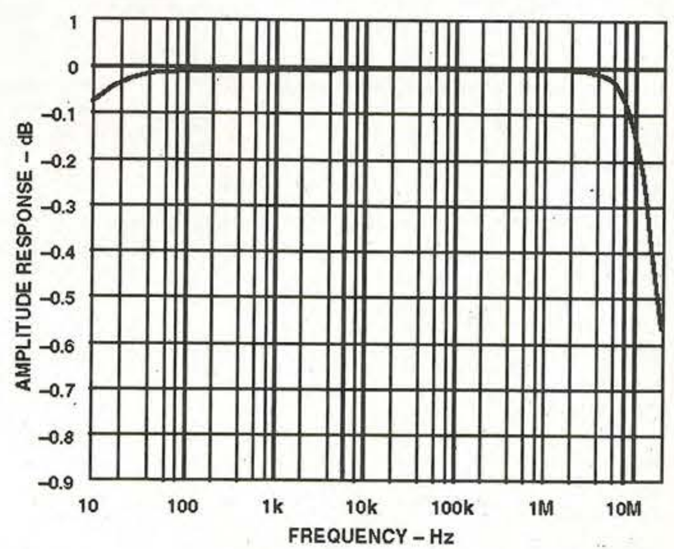

Fig. 1b). Showing the excellent 0.1dB flatness response from 10Hz to 10MHz for this AC coupled line receiver.

Where to look...

1. IE 5 Bug Parade Continues	http://www.bugnet.com/alerts/bugalert_102199.html
2. Circuit widens optocoupler's response to 1MHz	http://www.ednmag.com/reg/1995/122195/26dil.htm
3. 'OTA Obsoletes Op Amp'	C.F.Wheatley, H.A.Wittlinger, NEC Proc, Dec. 1969.
4. Global Semiconductor Datasheet Library	http://www.semi.com.tw
5. Analog Devices Inc	http://www.analog.com
6. Linear Technology Corporation	http://www.linear-tech.com
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Fig. 2. Gain of this 10MHz bandwidth differential input amplifier, is directly proportional to the current through R_5.

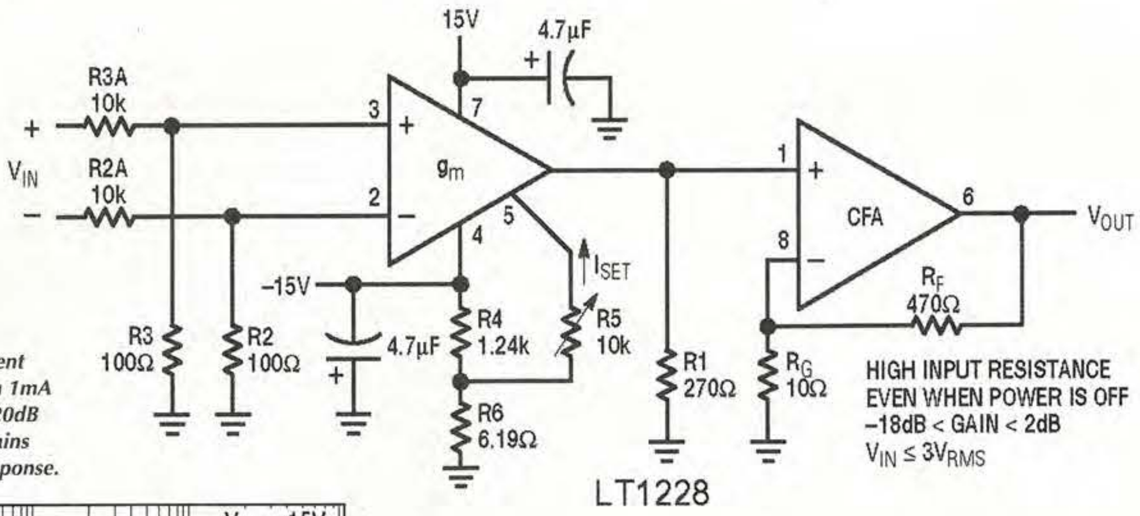
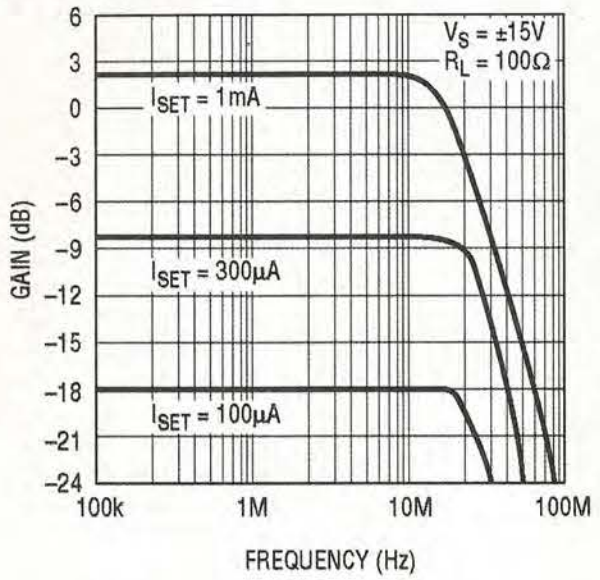


Fig. 2a). Varying the current through R_5 by 20dB, from 1mA to 100µA, produces this 20dB reduction in gain and retains the 10MHz frequency response.



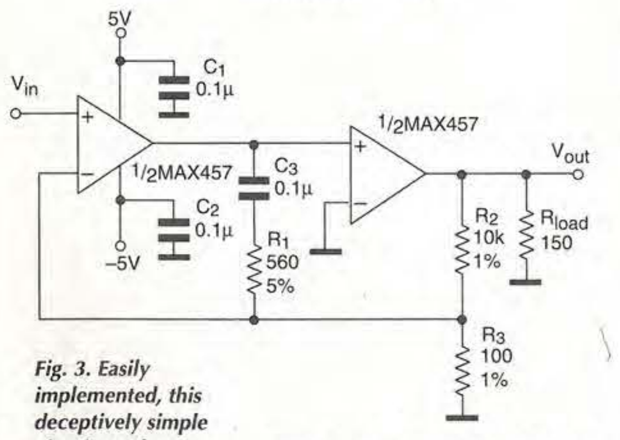


Fig. 3. Easily implemented, this deceptively simple circuit produces a 1GHz gain bandwidth product. With values shown, gain is 40dB with a -3dB response of 10MHz. The circuit can drive up to ±3.3 volts into a back terminated 75Ω coaxial cable.

The data sheet provides a range of applications, including RF levelling, voltage-controlled state-variable filters and video DC restoring circuits. The circuit I chose is a variable-gain differential-input amplifier having a constant 10MHz bandwidth over a 20dB change of gain, **Fig. 2**.

Variation of the I_{set} resistor R_5 varies the g_m, or gain, of the transconductance stage. The resistor used to convert the first stage output current into a voltage input for the second stage, namely R_1, determines the amplifier's bandwidth. Values larger than the 270Ω shown act with the capacitance at pin one, the input to the current feedback amplifier, to form a pole, further reducing bandwidth.

This circuit's gain can be controlled both by current or an external voltage. These can be provided digitally using a voltage output d-to-a converter and voltage to current converter, or more simply by using a digitally controlled potentiometer to replace R_5.

High-gain 10MHz bandwidth amplifier

One difficult design area is attaining high gain over moderate bandwidths and with a high input impedance. You would need such performance for an active oscilloscope probe for example.

The MAX457 is a dual CMOS video amplifier offering a very high impedance low capacitance input stage and 70MHz bandwidth. It was designed for use as a low gain 75Ω cable driver for video distribution amplifiers and is unity gain stable.

It is pin compatible with normal eight pin dual op-amps, so with suitable component changes could be used to upgrade many circuits. It provides a 72dB isolation between amplifiers at 5MHz and differential phase and gain of 0.2° and 0.5% respectively.

While Maxim[7] does not promote it as a transconductance amplifier, output current is proportional to the difference voltage at its inputs, with a G_m of 0.6A/V. Unlike the MAX435 and 436 transconductance amplifiers, discussed in the December issue, the gain for the MAX457 uses the normal non-inverting op-amp two resistor closed loop feedback method. Using a 1kΩ resistor to ground from the inverting input pin, the data sheet lists values of feedback resistor by desired gain and load impedance.

Its gain bandwidth product, while specified for a 75Ω load, increases with increased load impedance. With unity gain and a 75Ω load, its output is -3 dB at 70MHz. At a gain of two and a 150Ω load, bandwidth becomes 50MHz. Increase this load impedance to 750Ω and amend the feedback network to give a gain of ten; output is now -3 dB at 25MHz, representing an increase in gain bandwidth product from 70 to 250MHz.

As described in A0317.PDF this feature has been used to design a composite amplifier having a gain bandwidth product of 1GHz. This amplifier provides a 10MHz bandwidth at 40dB gain and can drive ±3.3V into a 150Ω load, i.e. a back-terminated 75Ω coaxial cable, **Fig. 3**.

Operating with no DC load, the first stage produces its maximum possible voltage gain, around 660V/V. The second amplifier, which drives the 150Ω load, has an open-loop voltage gain of 65V/V. This results in a total open-loop gain for the composite amplifier of 92dB.

Feedback components shown, together with the phase compensation components R_1 and C_3, ensure the required 40dB gain and bandwidth. However without these phase compensation components, the circuit can oscillate. Download A0317.PDF from Maxim's site for full details.

Perhaps by now you might think that transconductance amplifiers are always targeted to high frequency use. Not so, my next application by comparison is almost DC.

Current-sense amplifier

A current sensing amplifier from Maxim, the MAX4172, takes advantage of the transconductance amplifier benefits already mentioned.[7] It uses a differential measurement of the voltage dropped across an external sense resistor to measure the unknown current. Its output is a current proportional to the voltage drop being measured.

The device has a common-mode input range of 32V, regardless of the chip's supply voltage which can be as little as 3V. It has an 800kHz bandwidth when measuring a voltage drop of 100mV, reducing to 200kHz at 6.25mV.

A ground referred output is provided, which represents the floating current passing through the sense resistor. By suitable choice of load resistance, the subsequent output voltage can be scaled to suit your needs, **Fig. 4**.

In December, I highlighted Burr-Brown's OPA660 amplifier[8]. Sometimes called the 'perfect transistor', this transconductance amplifier provides the highest gain bandwidth of the transconductance amplifiers found in my searches. So perhaps it is fitting to end this exploration of transconductance amplifiers by introducing their dual version.

Two 'perfect transistors'

By omitting the buffer voltage amplifier part of the OPA660, Burr-Brown has been able to provide a dual transconductance amplifier in a six-teen-pin package. The OPA2662 is developed from two of these 'perfect transistors'. In the process, output current has been increased to ±75mA for each amplifier, but bandwidth has reduced to 370MHz.

In addition, each OTA can be turned off in 30ns and on in 200ns using TTL level logic provided in the sixteen pin package. These changes result in a device that is particularly suited to applications like analogue/digital video tape head driving, LED and laser diode driving and high-density disk drives reading and writing **Fig. 5**.

To facilitate your appraisal of this device, the company details a suitable two channel demonstration board, together with nine other application circuits, in the data sheet. ■

Are your files being read remotely?

As Windows 2000 – previously known as NT5 – is announced for release to computer makers in December and the shrink wrapped version planned for February, yet another Internet Explorer 5 security bug has been uncovered. In the December issue, I reported that Georgi Guninski of Bulgaria had discovered an 'ActiveX' security hole that allowed hostile code buried in a Web page or in an e-mail, to run on a computer without the user's knowledge. Disabling the option 'Active Scripting' prevents this. Alternately a patch can now be downloaded from Microsoft.

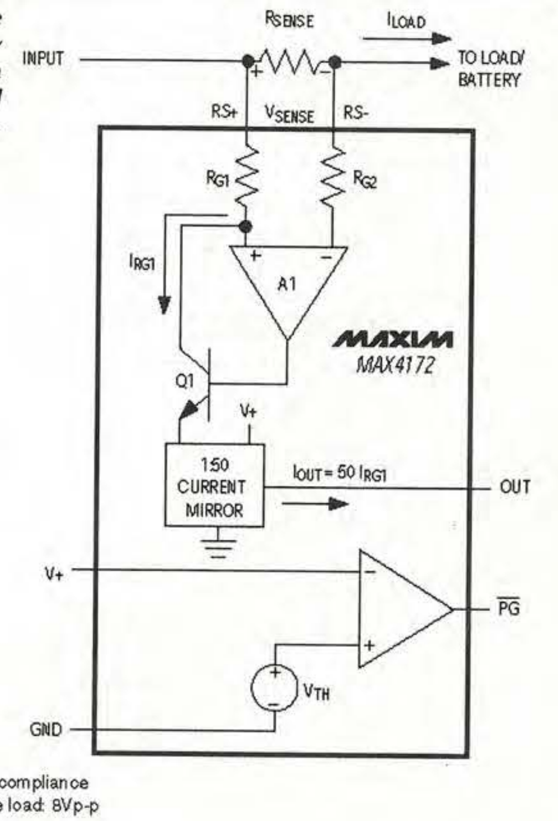
Now Guninski has reported yet another IE5 bug, making a total of three IE5 security holes he has reported within the space of a month. This latest uses the 'JavaScript Redirect' function with a little domain redirection, to trick IE5 into exposing the files on your computer, **Fig. A.**

Normally, should an Internet server request access to your data files, IE5 will prevent access. However following suitable domain redirections, ownership of your files can be made unclear and IE5 does not then prevent access.

This access will be 'read only' so your files will remain intact, but their contents will have been exposed. Microsoft plans to post another software patch, but in the meantime it recommends users should disable Active Scripting in their 'Internet Options'.[1]

***Fig. A.** If you use Explorer 5 to access Internet, review your IE 5 Internet ActiveX Scripting settings.*

***Fig. 4.** Able to sustain a 32V common mode voltage at its input terminals with a supply voltage as low as 3V, this circuit outputs a current proportional to the voltage dropped across R_{sense}.*



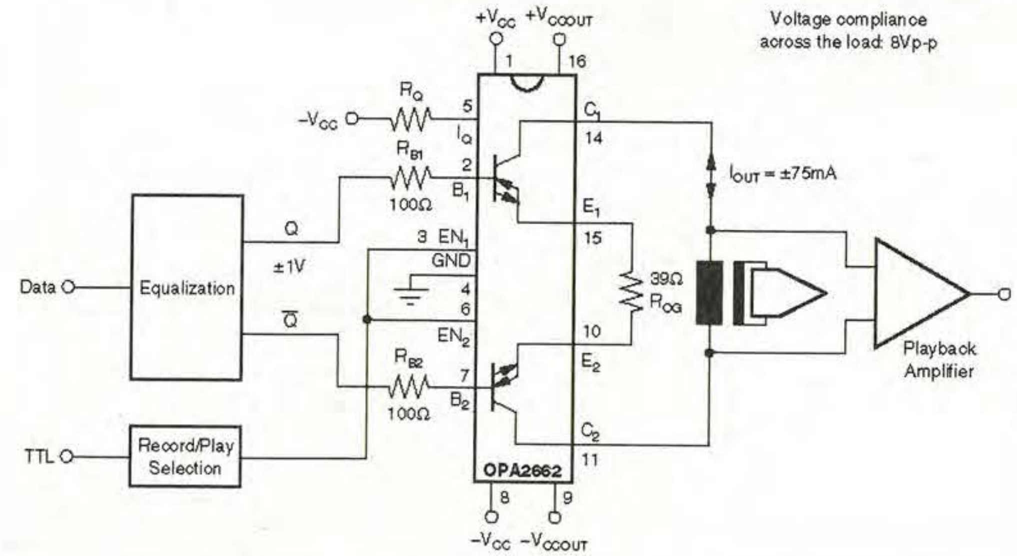


***Fig. 5.** Particularly suited to inductive loading, the OPA2662 provides a 250MHz large-signal bandwidth. With its 'EN' inputs grounded, both outputs become high impedance so record/playback head mechanical switches are not needed.*

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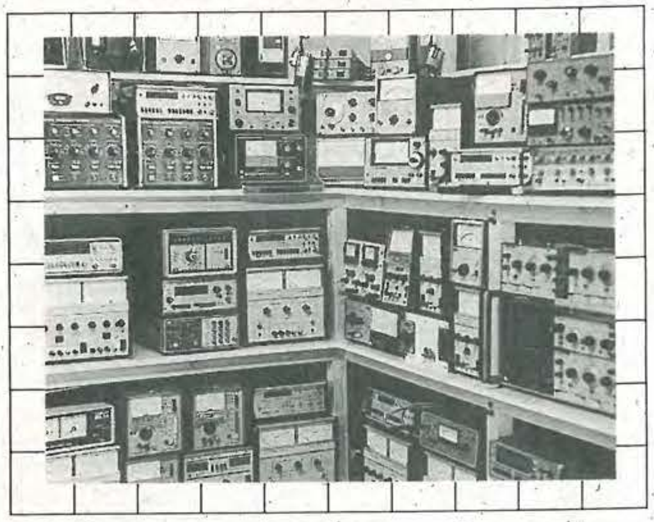

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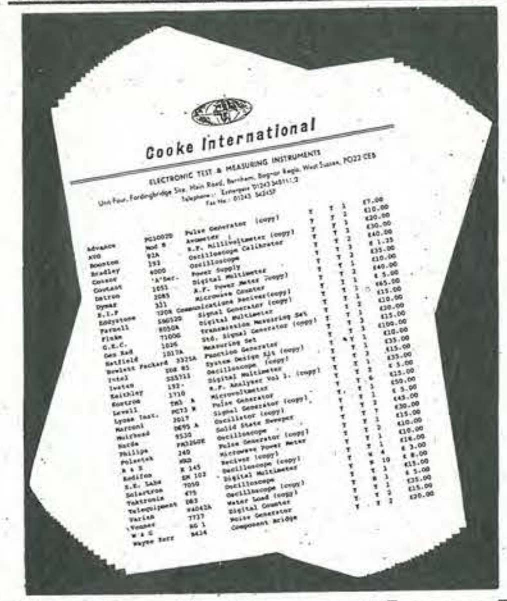

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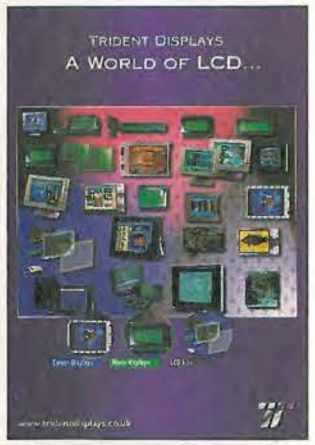

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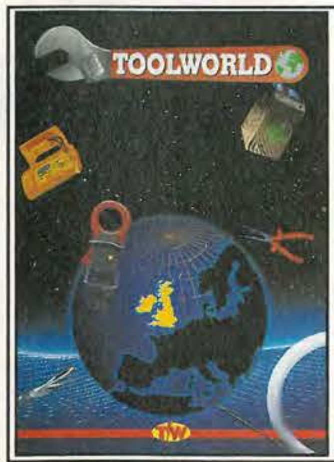

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WiNRADiO

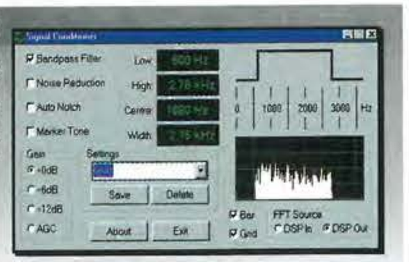
TAKiNG THE EUROPEAN RADiO MARKET BY STORM

TAKE A LOOK AT WiNRADiO's DiGiTAL SUiTE SOFTWARE (AWARDED 5 STARS BY WRTH)

1. WEFAX / HF Fax
2. Packet Radio for HF and VHF
3. Aircraft Addressing and Reporting System (ACARS)
4. Audio Oscilloscope, real time Spectrum Analyzer with calibration cursors
5. Squelch-controlled AF Recorder
6. DTMF, CTSS decode and analyse

The DSP applet provided with the WR3100i spectrum monitor ISA card (£995+VAT) allows continuous control of audio bandwidth and other signal conditioning functions.

ONLY £81.07 inc vat
(requires SoundBlaster 16 compatible sound card)



WiNADiO™ PC RECEIVERS

Available as either an internal ISA card that slips inside your PC, or as an external (portable) unit. WiNRADiO combines the power of your PC with the very latest, and greatest, synthesised receivers.

YOU CAN USE WiNRADiO™ SCANNiNG PC COMMUNiCATiON RECEiVERS FOR: Broadcast, media monitoring, professional & amateur radio communications, scanning, spot frequency, whole spectrum monitoring, instrumentation surveillance and recording.

If you're after the ultimate receiver-in-a-PC with full DSP then smile and say, "Hello" to the new **WR3100i-DSP** with its hardware for real-time recording, signal conditioning and decoding applications. It's all you need.

NEW EXTERNAL MODELS

EXTERNAL WiNRADiO™
We are now able to offer you a complete range of stand-alone WiNRADiO comms systems:

- **WR1000e - £359 INC VAT**
- **WR1500e - £429 INC VAT**
- **WR3100e - £1169 INC VAT**

Each stand-alone unit connects to your PC through either the basic RS232, or through an optional PCMCIA adapter (for high speed control).
The units are powered through either your existing 12v supply, or through an (entirely optional) NiMH rechargeable 12v battery pack.

"It's software is excellent.. more versatile and less idiosyncratic than that of the Icom IC-PCR1000"
WRTH 1999 Review

"Five stars for its mechanical design"
WRTH 1999 Review

"Most Innovative Receiver"
WRTH 1998 Awards

Model Name/Number	WR-1000	WR-1500	WR-3100
Construction of internals	WR-1000i/WR-1500i-3100iDSP- Internal full length ISA cards		
Construction of externals	WR-1000e/WR-1500e - 3100e - external RS232/PCMCIA (optional)		
Frequency range	0.5-1300 MHz	0.15-1500 MHz	0.15-1500 MHz
Modes	AM,SSB/CW,FM-N,FM-W	AM,LSB,USB,CW,FM-N,FM-W	AM,LSB,USB,CW,FM-N,FM-W
Tuning step size	100 Hz (5 Hz BFO)	100 Hz (1 Hz for SSB and CW)	100 Hz (1 Hz for SSB and CW)
IF bandwidths	6 kHz (AM/SSB), 17 kHz (FM-N), 230 kHz (W)	2.5 kHz(SSB/CW), 9 kHz (AM) 17 kHz (FM-N), 230 kHz (W)	2.5 kHz(SSB/CW), 9 kHz (AM) 17 kHz (FM-N), 230 kHz (W)
Receiver type	PLL-based triple-conv. superhet		
Scanning speed	10 ch/sec (AM), 50 ch/sec (FM)		
Audio output on card	200mW	200mW	200mW
Max on one motherboard	8 cards	8 cards	3-8 cards (pse ask)
Dynamic range	65 dB	65 dB	85dB
IF shift (passband tuning)	no	±2 kHz	±2 kHz
DSP in hardware	no - use optional DS software		YES (ISA card ONLY)
IRQ required	no	no	yes (for ISA card)
Spectrum Scope	yes	yes	yes
Visitune	yes	yes	yes
Published software API	yes	yes	yes (also DSP)
Internal ISA cards	£299 inc vat	£369 inc vat	£1169.13 inc
External units	£359 inc vat	£429 inc vat	£1169.13 inc (hardware DSP only internal)

PCMCIA Adapter (external): £69.00 inc when bought with 'e' series unit (otherwise: £99 inc)
PPS NiMH 12v Battery Pack and Charger: £99 inc when purchased with 'e' series unit (otherwise: £139 inc)
The WiNRADiO Digital Suite: £74.99 inc when purchased with a WiNRADiO receiver (otherwise: £81.05 inc)

To receive your completely free (no obligation) info pack and WiNRADiO software emulation demo disk all you have to do is get on the internet and go to our website at http://www.broadercasting.com. If you don't yet have easy access to the internet then by all means feel free to telephone us or send a fax.

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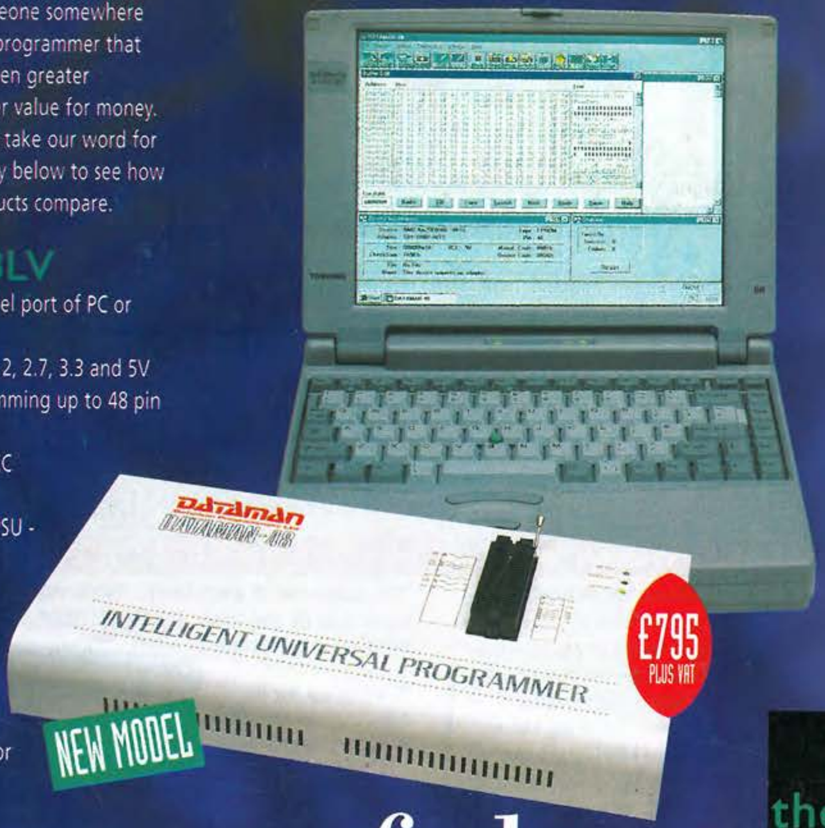
CIRCLE NO. 102 ON REPLY CARD

Surely not. Surely someone somewhere has developed a portable programmer that has even more features, even greater flexibility and is even better value for money.

Actually, no. But don't take our word for it. Use the feature summary below to see how other manufacturers' products compare.



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If you do not agree that these truly are the most powerful portable programmers you can buy, simply return your Dataman product within 30 days for a full refund

Dataman-48LV

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- Programs and verifies at 2, 2.7, 3.3 and 5V
- True no-adaptor programming up to 48 pin DIL devices
- Free universal 44 pin PLCC adaptor
- Built-in world standard PSU - for go-anywhere programming
- Package adaptors available for TSOP, PSOP, QFP, SOIC and PLCC
- Optional EPROM emulator

30 FREE DAY TRIAL

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free software upgrades + technical support for life

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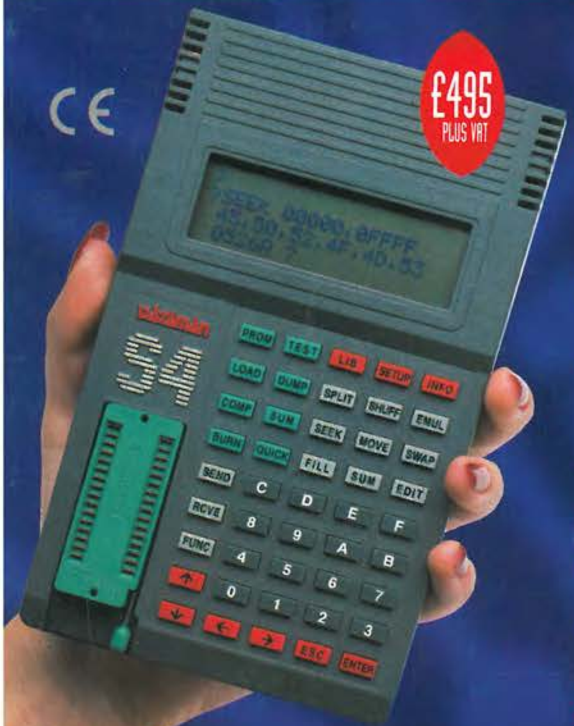

Dataman S4

- Programs 8 and 16 bit EPROMs, EEPROMs, PEROMs, 5 and 12V FLASH, Boot-Block FLASH, PICs, 8751 microcontrollers and more
- EPROM emulation as standard
- Rechargeable battery power for total portability
- All-in-one price includes emulation leads, AC charger, PC software, spare library ROM, user-friendly manual
- Supplied fully charged and ready to use

S4 GAL module

- Programs wide range of 20 and 24 pin logic devices from the major GAL vendors
- Supports JEDEC files from all popular compilers

Still as unbeatable as ever!

Beware of cheap imitations. Beware of false promises. Beware of hidden extras.

If you want the best, there's still only one choice - Dataman.

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Alternatively, request more detailed information on these and other market-leading programming solutions.

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