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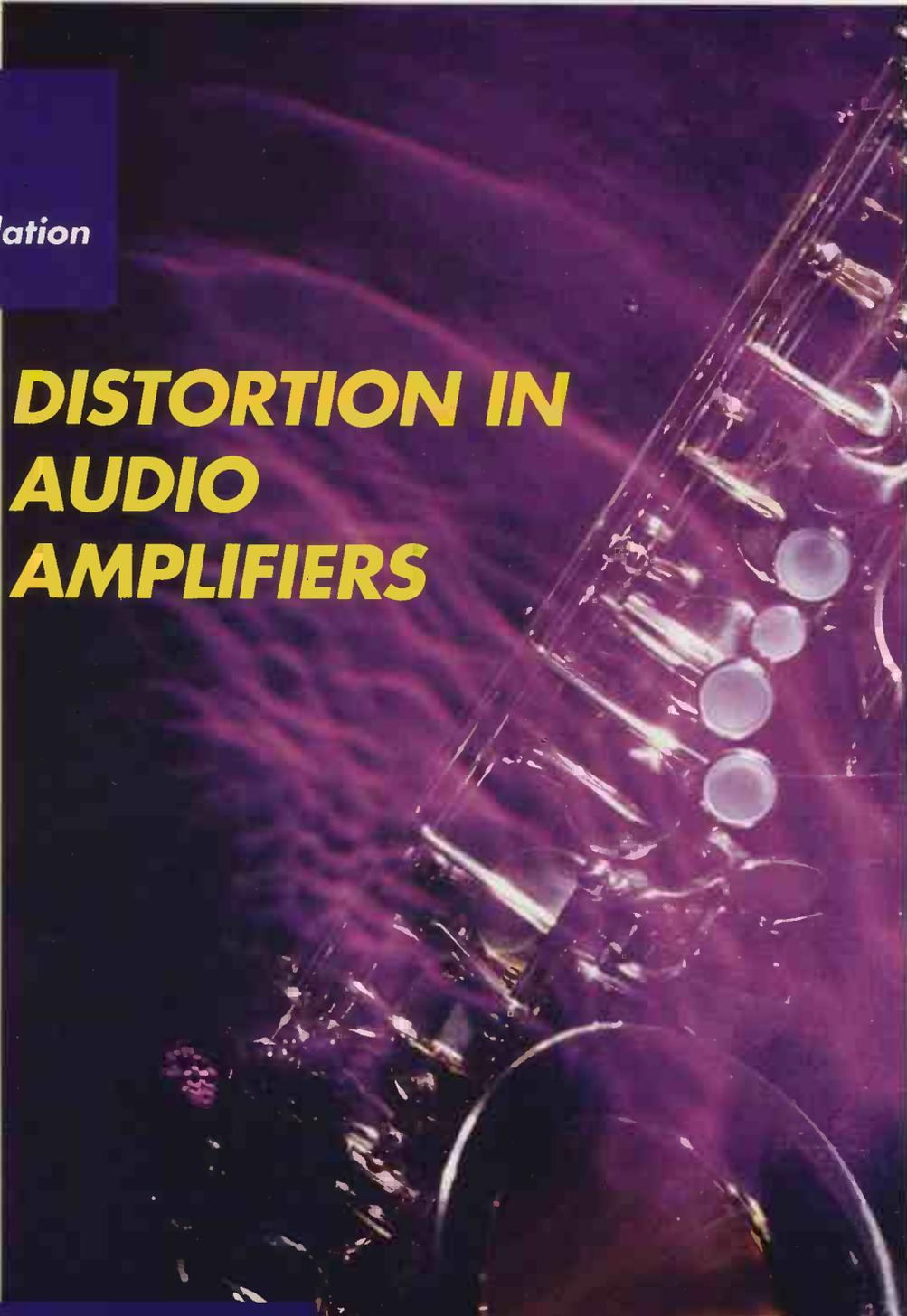
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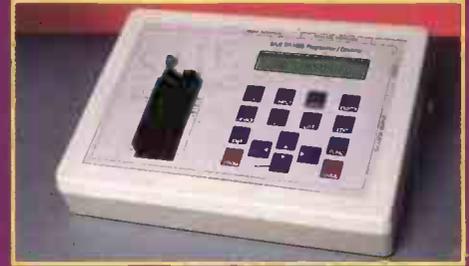
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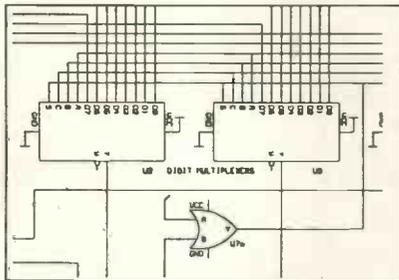
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In next month's issue: Spread Spectrum Radio. Until recently, this technology was the exclusive preserve of the military because of its resistance to jamming and interception. We present an in-depth design study on spread spectrum technology for civilian applications. The feature will include enough information to allow construction of an experimental system.

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Electronic Designs Right First Time?

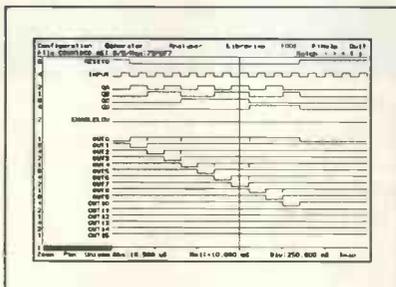
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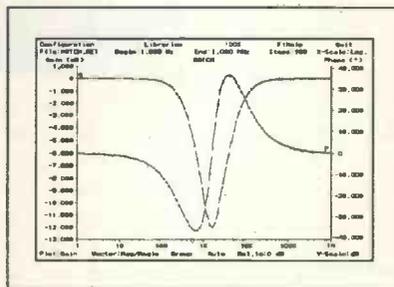
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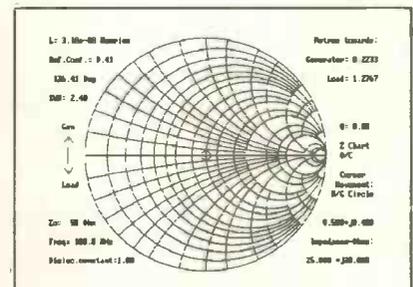
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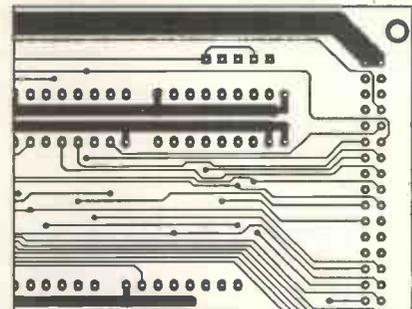
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Opportunity knocks digital TV

Any decisions currently being made on new broadcasting networks must take into account the possibilities offered by digital electronics.

In broadcast television for instance, it is now both feasible and practical to broadcast up to four digitally encoded programmes in the frequency space previously required for just a single analogue broadcast. Furthermore, the characteristics of the digital signal allow unprecedented frequency reuse. To bring home the point, currently available technology could provide up to eight new terrestrial TV channels to 97 per cent of the UK population using just two conventional UHF channels.

As Barry Fox points out in our news columns, the UK is likely to allocate UHF channels 35 and 37 to a new national network, Channel Five which will broadcast with conventional analogue technology. The Independent Television Commission will take the decision on the use to which these currently spare channels may be put, but the outcome will be largely influenced by the Treasury who will insist that the ITC raises short term cash with a franchise auction. This narrow objective can only be met with a standard TV service.

This would be a bad thing for European broadcast technology in general and this country in particular. We have it in our grasp to create

Europe's first national digital television service without disrupting existing broadcast systems or coercing viewers into chucking out their old televisions. Indeed, the new transmissions could be received on standard sets by simply adding a set-top converter.

The technologists promise that the frequency space which Channel Five will require to reach just 70 per cent of the UK population could bring eight standard channels or four standard channels plus a digital HDTV service to 97 per cent of the population, and with not a satellite in sight. The commercial value to European broadcast technology would be immense; we could receive major manufacturing benefits from having the World's first truly viable digital terrestrial TV service, an advantage of the sort which the US had – and lost – when it invented NTSC.

It is a different question as to what a doubling of network capacity might mean to programme quality. However, our electronics industry desperately needs the impetus which only a mass consumer advance can bring.

One thing is sure. If we allow the Treasury to take our technology decisions for us, we will be handing the potential world market for digital TV to the Far East on a solid gold plate.

Frank Ogden.

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UPDATE

Short term gain threatens digital TV

A twist of legal fate has put the Independent Television Commission in pole position to decide the future of TV in the UK. The ITC is legally obliged by the Broadcasting Act to try to sell two spare frequencies for a new analogue TV service, Channel 5.

This will raise money for the Treasury. But it will also obstruct a technical recommendation for the future of TV in Britain, which the ITC made at the end of June in a discussion document.

The document says viewers can have a far wider of choice of programmes if British broadcasters switch from analogue to digital transmission technology as quickly as possible. The report shows how analogue TV makes inefficient use of frequency.

If Ch 5 is an analogue service it will use

two frequencies to give 70% of the UK one new TV channel. Research by the BBC shows that if the two frequencies are used for digital TV, they will provide eight new services for 97% of the country.

When the document was published, the ITC was already working on a second report, due for publication in July. This was set to give the ITC's final verdict on Ch 5. The ITC will have to build a strong commercial case if the Home Office is to lift its obligation on the ITC to raise cash quickly by licensing Ch 5 to the highest bidder.

Today, 44 analogue Pal TV channels provide virtually the whole of the UK with four channels. Each occupies 8MHz of bandwidth. The same frequencies are reused many times over by transmitters that are far

enough away not to interfere with each other.

Two channels in the middle of the UHF band, 35 and 37, are used to connect VCRs to TV sets. The government's plan is to sell these for Ch 5. This will cause interference to millions of VCRs, which will have to be retuned at the broadcaster's expense. This is one reason why ITV turned down a bid for Ch 5 by a consortium led by Thames TV.

Because digital broadcasting is at lower power than analogue, there would be far less interference from digital TV on channels 35 and 37. Digital TV also makes more efficient use of broadcast frequencies.

Recent advances in technology will let each of today's 8MHz TV channels carry a digital data stream running at between 20 and 40Mbit/s.



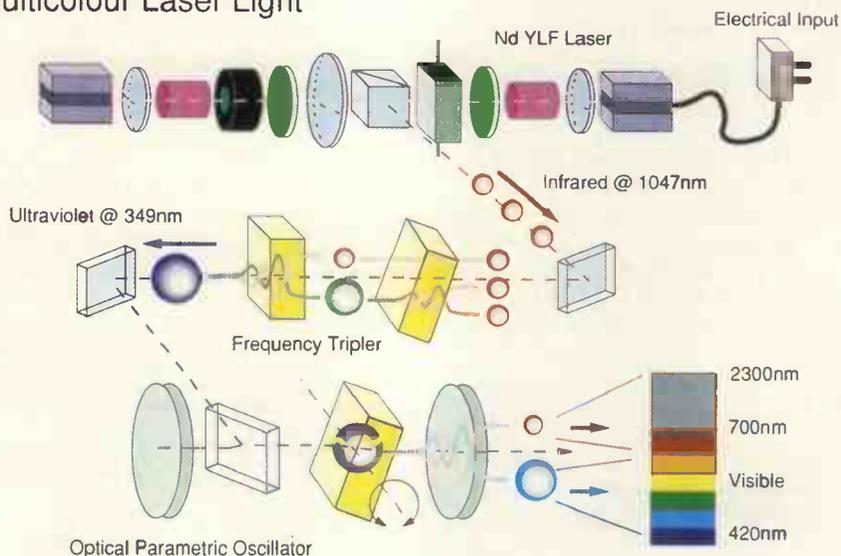
Mountain of soap: the 200 millionth Philips TV tube. The company recently hit the milestone of producing its 200 millionth colour picture tube – enough to go twice round the world at its Durham factory shown in the picture.

Crystal optical generator

Researchers at the University of St Andrews are using the non-linear effects in a crystal to produce a continuous range of frequencies from ultra-violet to infra-red from a single laser source. An infra-red light source is up-converted to higher frequency ultra-violet light, of high energy photons, using non-linear optics. These high energy

photons are split in the crystal oscillator and the resultant photon energies or frequencies are determined by the orientation of the crystal. Rotating the crystal produces light in wavelengths from 420 to 2300nm. The team has built a benchtop demonstrator and is seeking applications particularly in medical diagnostics.

Multicolour Laser Light



If analogue TV programmes are converted into digital code before transmission, then they each need between 5 and 10Mbit/s if there is to be no visible loss of picture quality.

One of today's analogue Pal TV programme channels can be replaced by at least one digital HDTV programme, four digital programmes of Pal picture quality, and eight of VHS quality. The ITC predicts that if analogue Pal broadcasting were shut down to release the 44 channels used to carry the four existing TV programmes, there would be room for a hundred digital programmes. But of course Pal TV cannot be shut down overnight; it would rob everyone of reception.

The ITC lists three routes. One is simulcasting. Existing broadcasters would be given access to digital frequencies, to transmit their programmes in analogue and digital. This would parallel the introduction of colour TV in the UK, when the new 625 line service ran alongside the old black and white 405 line service.

This worked well because the difference between black and white, and colour, was like chalk and cheese. Viewers will only invest in new digital equipment to receive simulcasts if the programmes look better. One idea is for broadcasters to transmit their digital programmes in widescreen format.

The second option is to make the digital frequencies available only for completely new programming.

The third option, which the ITC favours, is a mix of the other two. Broadcasters will simulcast the same programmes on analogue and digital frequencies and where there are extra digital frequencies available they, or

National puts squeeze on Euro video signals

Europe's first commercial operation to use broadcast quality compressed digital video signals is due to start in the autumn using technology developed by British firm National Transcommunications.

FilmNet, the Belgium-based provider of films for cable and satellite TV operators, is to use National's digital compression technology to squeeze three film channels into one satellite transponder channel, for distribution to cable operators in Belgium and Holland.

National is supplying the compression end of its MPEG based System 2000 front-to-back compressed video package, launched earlier this

year. The decompression hardware is coming from US firm Scientific Atlanta. The whole deal is believed to be worth around £2m.

Filmnet will digitise and compress signals in Brussels and then send them to a satellite uplink owned by the Belgian PTT. From there the signals will be sent to a transponder on the Intelsat 27.5° West satellite, which will broadcast them direct to cable operators.

National's compression technology makes it possible to squeeze eight TV channels on to a single satellite transponder, compared to just two standard Pal analogue channels.

licensed competitors, will broadcast different programmes.

BBC engineers said that research shows if the two Ch 5 frequencies were freed for digital TV, a network of 128 transmitters using UHF channels 35 and 37 could provide a new eight channel digital TV service for 97% of the UK population, without interfering with existing analogue services or VCRs.

This is possible because of a digital transmission system called orthogonal frequency division multiplexing. OFDM was developed for digital radio broadcasting. Instead of transmitting the digital code on a single radio carrier, it is spread over many narrow carriers. Several

hundred carriers per channel are used for radio, and several thousand for TV. OFDM widens the gaps between bits in each stream and thus stops transmitters interfering with each other even if they are using the same frequencies.

So if the ITC's second report can persuade the Home Office to drop its plans for an analogue Pal Ch 5, the UK can look forward to a smooth transition to digital TV in the second half of the decade. If the ITC's second report is wishy washy and the Home Office insists on Ch 5 being licensed now, the UK's transition to digital TV will be slow and painful.

Barry Fox

FIR design takes first prize in EW + WW/Xilinx contest

Bob Lawlor from County Dublin has won a programmable logic development system in the joint *EW + WW/Xilinx* competition. The challenge, which was in the December 1992 issue of *EW + WW* was to suggest the best use that can be made of the reprogrammability features of the Xilinx sram based FPGAs.

Lawlor's design used the products for implementing a digital finite impulse response (FIR) filter. These filters are used in digital signal processing.

For example, in decoding composite video to component RGB, the combination of a 13.5MHz (digital video sampling frequency) FIR bandpass filter centred on a subcarrier frequency (4.43MHz for Pal) and a 50Hz FIR comb filter facilitate very high quality separation of the luminance and chrominance data giving a much clearer picture than is possible with conventional TV sets.

Each of these FIR filters could be implemented using a Xilinx FPGA.

Frank Ogden, *EW + WW*'s editor, described Lawlor's application as

"ingenious". He said: "Enhanced Pal is very topical... An FIR implementation based on FPGA demonstrated the technology's capabilities to the full."

Panasonic and Motorola are helping the Science Museum update its telecommunications collection. The picture shows Roger Bridgman (left), the curator of communications at the Science Museum, being presented with a Panasonic I series handset by Peter Richardson, assistant marketing manager at Panasonic Business Systems. Motorola has given three mobile telephones to the museum. Bridgman said: "Mobile telephones are at the forefront of the revolution in personal communication and we are delighted that we have now been able to include them in our collection for the first time."



Consumer giants join in video CD challenge

Major consumer electronics companies from Europe and Japan are cooperating to set common standards for video CD to avoid costly and counter-productive format wars. Philips, Sony, Matsushita, and JVC have agreed on a world standard for putting VHS-quality video material on 12cm CDs. From the end of this year consumers will be able to buy a single player that reproduces either conventional audio CDs through a hi-fi system, or video CDs through a TV set. Unlike VHS video tapes, which follow different TV standards in different countries, and like audio CDs which are the same the world over, video CDs bought in any country will play on players anywhere in the world.

The new standard is called the White Book, and it builds on the Red Book standard set for audio CD by Philips and Sony when the CD system was launched ten years ago. The Red Book defines how to record 74min of digital stereo sound on a single-sided 12cm CD.

Over the last five years engineers have perfected ways of recording digitally coded video as well as stereo sound on a CD. The technique is known as full motion video, and a standard for the compression coding method was recently agreed by the ISO's moving picture experts group.

Philips originally planned to use MPEG FMV only to mix live video with games and educational programmes on its interactive disc system, CD-I. In January, Nimbus Records suggested that if the digital output

of a conventional CD audio player was connected to a simple MPEG FMV CD decoder, it could be used to play back movies and music videos through a TV. Nimbus designed a simple decoder and coined the name video CD.

In Japan, JVC is using MPEG FMV to put karaoke on CDs. JVC and Philips agreed a standard for karaoke CD in March this year.

The White Book pulls these strands together, so that any MPEG linear FMV disc will play on any player which has an MPEG decoder, either built in or added on. The White Book also cuts through all the national standards barriers that have so far bedeviled the video and TV industry.

Philips has already said it will release a selection of music videos and feature films on video CD this autumn. Nimbus is offering a service for companies which want to encode film and video material using MPEG technology.

Nimbus was initially angry at the White Book announcement. It was Nimbus' demonstration of video CD at the Midem music conference in Cannes, in January, which captured the imagination of the music and movie industries, and set the White Book ball rolling. There were two reasons for this.

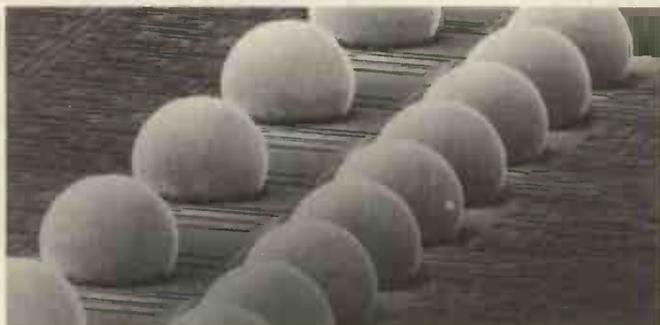
Nimbus was showing a double density CD, which could store a whole movie on a single disc. Double density CDs need tighter focusing of the player optics for guaranteed playback and Philips objected to this erosion of the Red Book standard. So Nimbus put

the double density idea on the back burner, and has ever since been using the name video CD to mean a single density disc. The White Book now defines a video CD format which will need two discs (of around 74min maximum) to hold a full length movie.

The other feature of the Midem demonstration was the system's simplicity and low cost. Video CD, according to Nimbus, will be a linear, passive format, like VHS tape, not an interactive format, like CD-I. This is what will let video CDs play on existing CD players connected to an add-on decoder, provided that (as most modern CD audio players do) the player has a digital output on the rear. Software sales will not have to wait for expensive CD-I players to become more popular. But here there is a hidden snag, which Nimbus failed to put across clearly.

As a pressing factory, Nimbus has access to a wide range of audio CD players. When tested, around 65% played a trick which is designed to protect owners of hi-fi systems but stymies their use for video CD. The trick is muting.

All CD-Rom data discs have a digital code mark or flag which identify the type of disc. On CD-Rom discs this flag is in the control sub-codes and runs continuously as the disc plays. On XA discs the flag is in the table of contents at the start of the disc. In each case the player recognises the flag and switches off the audio outputs. This stops the hi-fi trying to reproduce the code as music. Not only does the high pitched machine gun



Not reject tennis balls from Wimbledon, but tiny balls of solder used in flip chip mounting technology. These solder beads are applied to the bond pads of a circuit at the end of the usual manufacturing process. The circuit is turned over and placed on the board or substrate. Heat is applied, the solder melts, and a strong connection is formed.

Björn Hedlund, head of marketing at ABB Hafo, which offers the technology, said: "Flip chip mounting uses no packages and therefore reduces the overall height of the system. There are many benefits with this mounting method. It gives low resistance and inductance at the contact and is clearly cost effective where there are many connections.

Chip thieves hit Silicon Valley

Silicon Valley has been hit by a new wave of chip robberies with local police saying that semiconductors have become more valuable than gold or cocaine.

Armed gangs have been targeting small electronics firms. They usually hold the staff hostage for a while, ransack the place, and sometimes attack people.

Local police say that the number of these incidents is rising. They say that the robbers have no problem selling the chips on the black market. There is a high demand for chips toward the end of a quarter when chip allocations have run out and small firms are searching for chips to complete products.

Silicon Valley police are trying to find middlemen that buy stolen chips.

In the most recent incident, ten armed men attacked circuit board assembly firm Bestronics in San Jose.

They tied up more than a dozen workers with plastic handcuffs and stole a large amount of memory chips. One of the staff members was beaten badly.

MICROWAVE CONTROL PANEL. Mains operated, with touch switches. Complete with 4 digit display, digital clock, and 2 relay outputs one for power and one for pulsed power (programmable). Ideal for all sorts of precision timer applications etc. Now only £4.00 ref 4P151. Good experimenters board.

FIBRE OPTIC CABLE. Stranded optical fibres sheathed in black PVC. Five metre length £7.00 ref 7P29R or £2 a metre.

12V SOLAR CELL. 200mA output ideal for trickle charging etc. 300 mm square. Our price £15.00 ref 15P42R. Gives up to 15v.

PASSIVE INFRA-RED MOTION SENSOR. Complete with daylight sensor, adjustable lights on timer (8 secs -15 mins), 50' range with a 90 deg coverage. Manual override facility. Complete with wall brackets, bulb holders etc. Brand new and guaranteed. Now only £19.00 ref 19P29

Pack of two PAR38 bulbs for above unit £12.00 ref 12P43R

VIDEO SENDER UNIT. Transmit both audio and video signals from either a video camera, video recorder or computer to any standard TV set within a 100' range (tune TV to a spare channel). 12v DC op. £15.00 ref 15P39R Suitable mains adaptor £5.00 ref 5P191R. Turn your camcorder into a cordless cameral

FM TRANSMITTER Housed in a standard working 13A adaptor (bug is mains driven). £26.00 ref 26P2R. Good range.

MINATURE RADIO TRANSCEIVERS. A pair of walkie talkies with a range of up to 2 kilometres. Units measure 22x5x2x155mm. Complete with cases and earpieces. £30.00 ref 30P12R

FM CORDLESS MICROPHONE. Small hand held unit with a 500' range! 2 transmit power levels. Reqs PP3 battery. Tuneable to any FM receiver. Our price £15 ref 15P42AR.

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TOWERS INTERNATIONAL TRANSISTOR SELECTOR GUIDE. The ultimate equivalent book. New ed. £20.00 ref 20P32R.

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FM BUG Built and tested superior 9v operation £14.00 ref 14P3R

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SINCLAIR C5 MOTORS. 12v 29A (full load) 3300 rpm 6"x4" 1/4" O/P shaft. New. £22.00 ref 20P22R. Limited stocks.

As above but with fitted 4 to 1 inline reduction box (800rpm) and toothed nylon belt drive cog £45.00 ref 40P8R. 800 rpm.

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12V 19A TRANSFORMER Ex equipment £20 but OK

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LED PACK. .50 red, 50 green, 50 yellow all 5mm £8.00 ref 8P52

MINIMUM GOODS ORDER £5.00. TRADE ORDERS FROM GOVERNMENT, SCHOOLS, UNIVERSITIES, & LOCAL AUTHORITIES WELCOME. ALL GOODS SUPPLIED SUBJECT TO OUR CONDITIONS OF SALE AND UNLESS OTHERWISE STATED GUARANTEED FOR 30 DAYS.

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THIS MONTHS SPECIAL OFFERS

AMSTRAD MP3 #19.00

VHF/UHF TV Receiver, converts RGB or composite monitor into colour TV. Brand new and cased.

#19.00 each. REF: EW19P1

EMERGENCY LIGHTING #5.00

Fully cased units with two swivel lamps mounted on the top. Just add a 6V sealed lead acid battery and you have a complete automatic system.

Gives about 3 hours of light as soon as the mains fail.

Our Price #5.00. REF: EW5P1

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You can have a 2 Mtr length of mains cable with a fitted 13A plug (with sleeved pins) for just 30p!

Minimum quantity: PKT 10 #3.00 REF: EW3P1 or you can have PKT 100 #27.00 REF: EW27P2

CORDLESS PHONES (non working)

#30.00 for 3

These are BT Curlews. They are not new and have been stored in sunlight so the plastic cases are discoloured. They also have handset aerial missing. This is a standard telescopic aerial - piece of wire will do! These are NON RETURNABLE.

Our Price PKT 3 for #30.00. REF: EW30P1

ANSWER PHONES from only

#18.00

Yes its true BT approved push button dialling.

These are customer returned units and have two faults. We will supply you with a chip to cure one of the faults then you have to sort out the other problem! NON RETURNABLE.

Price #18.00 each REF: EW18P2

or #60.00 PKT 4 REF: EW60P2

ANSWER PHONES - HIGH SPEC LCD DISPLAY

Once again, BT approved push button dialling answerphones. LCD display, digital speech chip, fast rewind, memories plus all the features found on a BT Response 400 machine. Again, we will supply you with a chip to cure one of the faults then you have to sort out the other problem!

NON-RETURNABLE.

Price #28.00 each REF: EW28P2

or #100.00 PKT 4 REF: EW100P1

POWER SUPPLIES for ANSWER PHONES above

If you need a DC plug in power supply for the above answer phones they are #4.00 each.

Price #4.00 each REF: EW4P5

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MICE 2 button serial mouse with 3.5" s/ware. £8.40 ref BPCM6, 3button serial mouse with 3.5" s/ware £9.60 ref BPCM7.

KEYBOARDS 102 AT UK standard keyboard £18.60 ref BPCM4, Deluxe keyboard 102 AT UK £26.40 ref BPCM5.

SOFTWARE MS DOS V5 OEM version. £39.60 ref BPCM8, MS WINDOWS V3.1 OEM version. £42 ref BPCM9.

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HARD DRIVES 42MB IDE 17ms £99 ref BPCDD01, 89MB IDE 16ms ref BPCDD02, 130MB IDE 15ms £215 ref BPCDD03, 213MB IDE 14ms £298 ref BPCDD04.

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EXAMPLES OF COMPLETE SYSTEMS

386SX-33 SYSTEM
386SX-33 board at £82.80, case £51.60, 2MB ram £52.80, 42MB drive £99, 512K SVGA card £31.20, 3.5" FDD £32.34, multi I/O card £11 SVGA colour monitor £174, 102 kboard, £25 build fee if required. Total £579.34

486DX-33 SYSTEM
486DX-33 board £378, case £51.60, 2MB ram £52.80, 89MB drive £166, 512 SVGA card £31.20, 3.5" FDD £32.34, multi I/O card £11, SVGA monitor £174, 102 kboard £18.60, £25 build fee if required. Total £939.84.

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code sound nasty, it can quickly destroy loudspeakers.

But of course the player cannot know that its output is connected to a video CD decoder which needs the code to make pictures. So 65% of all CD players will not work with video CD decoders unless the factory making the disc plays a very simple trick of its own.

The factory simply leaves the muting flag off any data disc that is a video CD. It's as easy as that. Nimbus sent a report on muting to Philips and is understandably angry that Philips is talking about modifying future CD players to ignore the muting flags, instead of writing the no-flag requirement into the White Book standard for video CDs.

The Nimbus option makes video CD backwards-compatible with many existing CD players, but the Philips option makes video CD compatible only with expensive

CD-I players and future generations of modified audio CD players.

The wrong decision on this small point will mean the difference between commercial success and failure for video CD.

In a parallel move, the same four giants, have joined with Thomson Consumer Electronics, Hitachi, Mitsubishi, Sanyo, Sharp, and Toshiba, to agree a single standard for a new generation of digital video tape recorder for home use.

The aim is to find a digital standard which will be suitable for recording existing TV programmes, and future high definition TV formats. Matsushita, Sony, Philips, and Thomson have all been working on this separately.

The ten companies are already agreed on a basic specification. The digital VCR will convert the analogue picture signals into

digital code with a system similar to the MPEG system used for video CD. But whereas video CD makes do with a digital data rate of 1.5MB/s, the new DVCR will handle 25MB/s to give much higher picture quality.

The DVTR will record onto a tape which is 6.35mm wide, half the width of VHS tape, and the digital tape will be coated with an evaporated film of pure metal, rather than the magnetic oxide powder used for VHS. There will be two types of cassette, one large for domestic use with 4.5 hours playing time, the other small for portable use with 1 hour capacity.

The HDTV version will run the tape twice as fast, to record a data rate of 50MB/s, but half the playing time from the same cassettes.

BF

Blue Lightning strikes at Pentium

With massive shortages of the Pentium microprocessor lasting for most of this year and well into 1994, IBM is betting that its superfast 486 microprocessors will find a widespread market.

IBM's triple-clocked 486 microprocessors, dubbed *Blue Lightning*, is available in a 25/75MHz version; a 33/100MHz version is due in the fourth quarter.

Blue Lightning has twice as much cache as Intel's 33/66MHz DX2 and uses half the power. The 25/75MHz version runs at 26Mips compared with 25Mips

for Intel's 33/66MHz 486. The 33/100MHz Blue Lightning runs at 34Mips.

IBM's licensing agreement with Intel prevents it from selling the chips directly. Customers will have to buy the chips in sub-system products.

IBM believes that the Blue Lightning will win over some customers waiting for Intel's expensive Pentium. IBM says that it also has manufacturing rights to the Pentium.

The first IBM Pentium is not due until mid-1994, just when the Pentium market is expected to start taking off.

Korea aims for world dram market

Korea plans to be dominant in the memory chip market of the 1990s, just as Japan dominated it in the 1980s and America in the 1970s.

Samsung, the world's number one dram supplier, has built enough manufacturing capacity to take 20% of the world market for the upcoming generation of 16Mbit drams and has targeted the end of 1994 for achieving that share.

Hyundai, Korea's number two chip maker, is also aiming for a 20% share of the world 16Mbit market, but does not expect to catch up with Samsung in terms of unit output until the third quarter of 1994. Hyundai has stabilised its 16Mbit process and intends to start commercial production in October and to hit the one million a month mark in early 1994. Full capacity is 2.5 to 3 million units a month.

Both are making drams on 8in wafers. It is not known if any of the Japanese chip companies are doing this in production quantities. However, according to the Koreans, it will not be economic to make them on 6in wafers.

Samsung expects to be producing between 2 to 3 million 16Mbit drams a month by the start of the fourth quarter.

The price is beginning to erode fast, around \$65 with reports of DM90 (\$55) pricing in Germany. And the Koreans are projecting the crossover on price per bit with the 4Mbit will occur in the first half of 1994. That would imply a 16Mbit price of \$40 to \$45.

David Manners, *Electronics Weekly*

GSM delays hit communicator launch

Delays in agreeing the GSM standards for cellular telephones in Europe have forced EO to launch its personal communicator without the cellular phone option. Sales could be hit as a result.

The product was launched in the USA in July with a cell phone built in, but the EO PC440, due for launch in Europe in September or October, will have no phone.

Victor Hrovat, general manager of AT&T Microelectronics, which makes the products's Hobbit microprocessor, said: "In Europe the digital GSM standard does not support data over voice. The technical standard is fixed but approval will probably not happen until the first quarter of next year".

Buyers of the model about to be launched will only be able to upgrade to using cell phones via the machine's PCMCIA slot. The PC440 is a pen-based communicator that can send and receive faxes and E-mail as well as acting as a personal organiser. The link with cellular phones to make it truly

portable has been one of the main hype points in the publicity build up and it seems certain that initial sales will be hit because of this missing feature.

As well as transmitting data, users can add 10s of voice to any message sent. Messages can be longer if more than the standard 8Mbyte of memory is fitted.

Operating system is *PenPoint* from Go for which a number of applications are already available, and many more are in the pipeline. Algorithms are included to recognise handwriting.

The product can be linked to an external keyboard (a soft keyboard is included), printer, or desk-top computer.

Initial target market is the business user, though it is envisaged that the will eventually become a consumer item too.

Hrovat's prediction is that within two years they will be in every shop".

In the US the product is already being sold through AT&T's phone shops.

Steve Rogerson

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- HP Vector Voltmeter type 8405A - £400 to £600 - old or new colour.
- HP Synthesiser/signal generator type 8672A - 2 to 18GHz £4000.
- HP Oscillographic recorder type 7404A - 4 track - £350.
- HP Plotter type 9872B - 4 pen - £300.
- HP Sweep Oscillators type 8690 A & B + plug-ins from 10Mc/s to 18GHz also 18-40GHz. P.O.R.
- HP Network Analyser type 8407A + 8412A + 8601A - 100Kc/s - 110Mc/s - £500-£1000.
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- HP Signal Generator type 8640B - Opt001 + 003 - .5-512Mc/s AM/FM - £1000.
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- Racal/Dana 9301A-9302 RF Millivoltmeter - 1.5-2GHz - £250-£400.
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- Alltech Stoddart receiver type 1727A - .01-32Mc/s - £2500.
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- Alltech Stoddart receiver type NM65T - 1 to 10GHz - £1500.
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- Infra-red Binoculars in fibre-glass carrying case - tested - £100. Infra-red AFV sights £100.
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- Tektronix 491 spectrum analyser - 1.5GHz-40GHz - as new - £1000 or 10Mc/s 40GHz.
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- Knott Polyskaner WM1001 + WM5001 + WM3002 + WM4001 - £500.
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- SE Lab Eight Four - FM 4 Channel recorder - £200.
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- H.P. 3438A digital multimeter.
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- Fluke 80 high voltage divider.
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- Avo VCM163 valve tester+book £300.
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- PPM 8000 programmable scanner.
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- HP 3763A error detector. £250.
- Racal/Dana signal generator 9082 - 1.5-520Mc/s - £800.
- Racal/Dana signal generator 9082H - 1.5-520Mc/s - £900.
- Claude Lyons Compuline - line condition monitor - in case - LMP1+LCM1 £500.
- Efratom Atomic FX standard FRT - FRK - 1-1-5-10Mc/s. £3K tested.
- Racal 4D recorder - £350-£450 In carrying bag as new.
- HP8350A sweep oscillator mainframe + HP11869A RF PI adaptor - £1500.
- Alltech - precision automatic noise figure indicator type 75 - £250.
- Adret FX synthesizer 2230A - 1Mc/s. £250.
- Tektronix - 7S12-7S14-7T11-7S11-S1-S52-S53.
- Rotek 610 AC/DC calibrator. £2K + book.
- Marconi TF2512 RF power meter - 10 or 30 watts - 50 ohms - £80.
- Marconi multiplex tester type 2830.
- Marconi digital simulator type 2828A.
- Marconi channel access switch type 2831.
- Marconi automatic distortion meter type TF2337A - £150.
- Marconi mod meters type TF2304 - £250.
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- HP 3763A error detector.
- HP 8016A word generator.
- HP 489A micro-wave amp - 1-2GHz.
- HP 8565A spectrum analyser - .01-22GHz - £4k.
- HP 5065A rubidium vapour FX standard - £5k.
- Fluke 893A differential meters - £100 ea.
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- Takeda Rilken TR4120 tracking scope + TR1604P digital memory.
- EG&G Parc model 4001 indicator + 4203 signal averager PI.
- Systron Donner 6120 counter/timer A+B+C inputs - 18GHz - £1k.
- Racal/Dana 9083 signal source - two tone - £250.
- Systron Donner signal generator 1702 - synthesized to 1GHz - AM/FM.
- Systron Donner microwave counter 6057 - 18GHz - Nixey tube - £600.
- Racal/Dana synthesized signal generator 9081 - 520Mc/s - AM-FM. £600.
- Farnell SSG520 synthesized signal generator - 520Mc/s - £500.
- Farnell TTS520 test set - £500 - both £900.
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- Tektronix TM515 mainframe + TM5006 mainframe.
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- Farnell electronic load type RB1030-35.
- Fairchild interference analyser model EMC-25 - 14kc/s-1GHz.
- Fluke 1720A instrument controller+keyboard.
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- HP3730B down converter - £200.
- Bradley oscilloscope calibrator type 192 - £600.
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- Marconi TF2370 spectrum ANZ + TK2373 FX extender 1250Mc/s+trk gen - £2.5k-£3k.
- Racal receivers - RA17L-RA1217-RA1218-RA1772-RA1792 - P.O.R.
- Systron Donner microwave counter 6057 - 18GHz - nixey tube - £600.
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RESEARCH NOTES

New laser targets commercial uses

Sandia National Laboratories scientists have reported (*Electronics Letters*, Vol 29, No 10) the first electrically injected visible-light, vertical-cavity, surface-emitting laser. Hidden behind this mouthful of a description is a device that could push edge-emitting solid-state lasers into the background.

Vertical-cavity surface-emitting lasers (VCSELs) emit light directly from the top layer of the semiconductor, giving them several fundamental advantages over edge emitters. As emission is perpendicular to the surface, it is easier to fabricate closely packed arrays of devices. VCSELs are also well known for their tight circular beam which greatly facilitates interfacing with optical fibres.

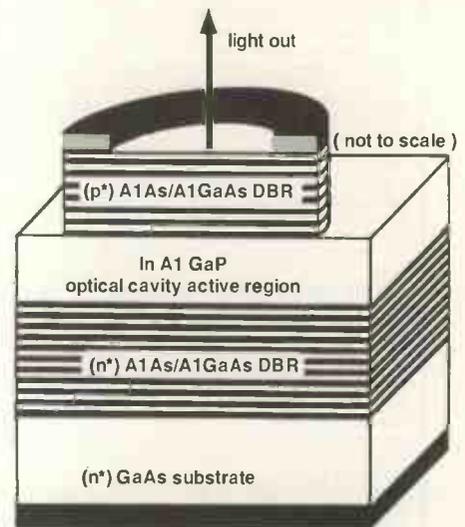
Surface emitting lasers are not in themselves new; optically pumped devices have been demonstrated in the laboratory for about two years. But such lasers have to be pumped by other lasers and so are scarcely practical. Electrical injection is essential if a

laser is to have any real commercial application.

Until this latest report, the shortest wavelength reported for a VCSEL diode was 699nm, just at the edge of the visible light spectrum. The new Sandia device emits bright red light at between 639nm and 661nm, making it potentially suitable for a whole range of commercial uses, including bar-code scanners, displays, holographic memories and fibre-optic networks. It could render the cumbersome helium-neon laser obsolete.

The new surface emitting laser is made using strained layer superlattice (SLS) technology, which allows stacking multiple alternating layers of very thin compound semiconductor materials. The layers are so thin that strain between them can accommodate slight mismatches between their lattice spacing, thus allowing a much wider variety of compounds with precisely tailored properties.

Operation at room temperature is in pulsed



Highlight: Visible laser light comes out at the top of the chip rather than the side.

mode. Peak power exceeds 3.3mW at a wavelength of 650nm. At that wavelength threshold current is 30mA and beam diameter is 20µm. The researchers say that substantially improved performance is expected from devices processed with more advanced techniques such as ion implantation.

Computers model stairway to heaven

How would an intelligent computer face its own approaching death? Would it burst into song, as did the HAL 9000 in the film *2001*. Or sink into terminal electronic depression?

A recent report (*Scientific American*, May 1993) suggests that some of the more fanciful ideas may not be too wide of the mark. The report describes how Stephen Thaler, a physicist for McDonnell Douglas, spent his evenings devising a program that would in effect kill off a neural network. Neural networks (when healthy) attempt to mimic the learning processes of the human brain by establishing links, or synapses, between neurons. Memories exist in the strengths, or weightings, of these various synapses. In an electronic (and possibly human) network, the weightings vary with use, thus allowing the system to learn from experience. Electronic neural networks are particularly good at tasks where self-learning plays a big part, such as pattern

recognition, machine translation etc, the very areas where humans have traditionally out-performed computers. Thaler's main work with neural networks has been to optimise process control for growing diamond crystals.

Thaler's lateral thinking led him to wonder what would happen if a neural network were subject to the same sort of destruction that happens when a human brain suffers from a degenerative disease or terminal asphyxia. In his spare time he wrote a "death" program designed to mimic the depolarisation of the synapses that occurs when biological brains die.

When tested on an eight-unit network that had been trained to model an exclusive-OR function, the results were intriguing. After between 10% and 60% of the network had effectively been destroyed, it produced gibberish. But when the level of synaptic destruction reached 90% – corresponding to "near death" – the

output began to settle on distinct values. It produced what Thaler describes as "whimsical" states, values that would not be created by a healthy network and which were not trained into it. Yet training does appear to be a prerequisite of these "whimsical" states; an untrained network produces only gibberish.

Comparing all this with human near death experiences is clearly irresistible, even if it does rather stretch a point to extrapolate to the human situation from an eight-neuron network. Could the whimsical states of the neural network correspond in any way to the experiences reported by dying people, the mystical experiences, the flashbacks, the slowing down of time?

Obviously it's difficult to do more than speculate. But it is intriguing to think that mathematics might soon be able to model our very demise. Let's hope it reveals some light at the end of the tunnel.

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Record recovery for silicon photodetector

Researchers from the University of Minnesota and the University of Rochester (US) have developed what they claim is the fastest silicon photodetector yet reported – a device capable of operating at up to 75GHz (*Applied Physics Letters*, 17.5.93).

The key to the new MSM (metal-semiconductor-metal) detector is said to be the speed at which it recovers after registering a signal. Less than 20ps after detecting one optical pulse, the detector is ready to detect another.

Structurally, the device consists of alternating strips of metal and silicon, each less than a third of a micron wide. Light hitting the surface creates electrons which move from the semiconductor to the metal electrodes. Because the spacings are so small, electrons have only a short distance to

travel to create a signal – hence the speed.

The team, led by Professor Thomas Hsiang at Rochester's laboratory for Laser Energetics, has studied the device's response to light of varying wavelengths, knowing that red (long wavelength) light penetrates silicon more deeply than violet (short wavelength). They showed that the detector's performance is limited mainly by the fact that when photons penetrate deeply, the resultant electrons must travel further to the electrodes to generate a signal. So while 75GHz has been achieved with violet light, the performance drops off to around 38GHz with pulses of red light.

To improve this result at longer wavelengths, Hsiang suggests burying an insulating material, such as silicon oxide, just below the surface of the photodetector. The extra layer, he thinks, would prevent

unwanted electrical activity and should improve the detector's response to the longer wavelengths commonly employed in fibre optic communications.

Hsiang's group is one of the very few in the world able to measure such ultrafast optical devices. As an optical signal source, the group uses a tuneable titanium-doped sapphire laser capable of producing powerful pulses of light lasting less than a picosecond.

Although the latest achievement is by no means the fastest photodetector on record (Hsiang's team last year fabricated a gallium arsenide device capable of 510GHz), the fact that 75GHz has been achieved with silicon is noteworthy. Silicon technology is much commoner, much cheaper and more easily integrated into a whole range of low-cost devices.

Taking an ocean's temperature

A multi-million dollar programme to use sound to map temperatures in the world's oceans has been launched in the US. You might imagine that measuring the temperature of the sea would be a simple matter of dangling a thermometer (or possibly a thermocouple) on a length of string over the side of a ship. That indeed is how oceanographers and meteorologists have been doing it for more than a century. But to build any meaningful picture over whole oceans has proved impossible.

Surface temperatures from all over the world can of course be measured by infra-red satellites, but such measurements can

not indicate in any detail what is going on at depth. To come up with meaningful figures for the temperature of bulk sea water would require a vast number of thermometers to be dangled in an equally huge number of representative locations. In practice thousands of ships or buoys would need to be stationed everywhere from the equator to the poles.

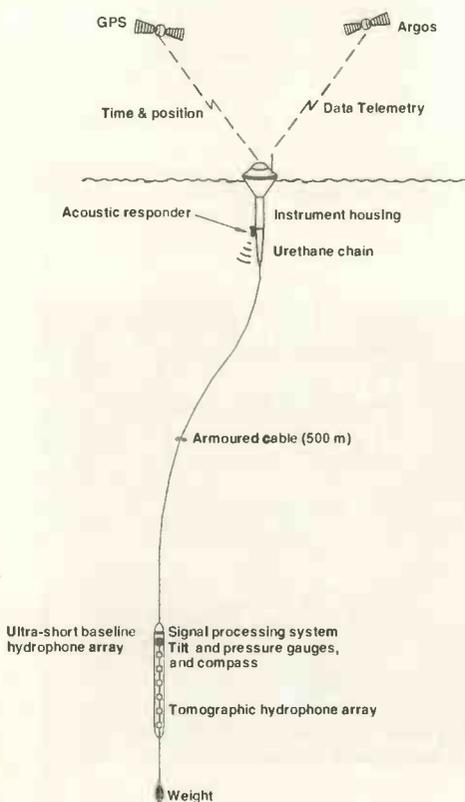
A novel technique that will measure changes in average sea temperatures, but at a fraction of that cost, was dreamt up in the early 80s by Dr John Spiesberger, associate professor of meteorology at Pennsylvania State University. Spiesberger realised that it might be possible to exploit the fact that the velocity of sound through water varies with its temperature. So by measuring how long it takes for a pulse of sound to traverse an ocean, it should be possible to get a very accurate measure of its average temperature.

While working at Wood's Hole Oceanographic Institution, Spiesberger began transmitting acoustic pulses from Hawaii to receivers distributed from the Aleutian Islands in the north, to California in the south. The journey time in each case is about 40 minutes.

Now Spiesberger is leading a multi-institution team that has been awarded \$10.4 million grant by the US Defense Advanced Projects Agency. Its aim is to take the temperature of all the world's oceans to within a fiftieth part of a degree Celsius. Mapping involves specially developed instruments which can either generate or detect sounds at up to 1000m below the surface. The time it takes for sound to travel between these instruments will be transmitted via satellite to laboratories.

The team hope to be able to watch, virtually in real time, how the interior of the oceans change temperature. They aim to produce, with the help of powerful computers, temperature maps that change from hour to hour, just like weather maps.

Given the sort of accuracy possible with acoustic sounding, the team hopes to make useful predictions about long term climate change. In particular, they hope to be able to predict the occurrence of El Nino, the periodic change in ocean currents off the west coast of the Americas that has had such a damaging effect of fisheries and agriculture.



Stay out of the rain if you want to avoid a tan?

The University of Arizona Health Sciences Center has been providing some helpful hints on how to identify lightning strike victims and what to do if you find yourself in a "lightning situation".

"Twenty million volts of electricity generally pass through a person directly struck by lightning...", a victim who is often found "unconscious, disorientated and/or suffering from paralysis in the extremities". So take care not to mistake such people for an assault casualty, especially (says Arizona) if you find them with "their cloth-

ing in disarray or blown off".

Lightning-strike victims, as well as being divested of their apparel, may also have flash burns that masquerade as a deep sunburn. So be especially vigilant next time you come across someone deeply tanned and scantily-clad. Rather than sunbathing, they may have been struck by a bolt from the blue.

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

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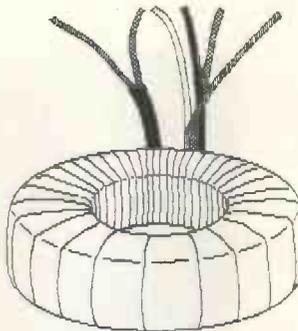
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CIRCLE NO. 110 ON REPLY CARD

In the first article of a new series, Douglas Self examines the origins of distortion in audio amplifiers based on research work to isolate the individual mechanisms. His results are surprising and will overturn much conventional wisdom in the world of hi-fi as well as having implications for general analogue design.



Distortion in power amplifiers

1: the sources of distortion

It seems surprising that in a world which can build the Space Shuttle and detect the echoes of the birth of the universe, we still have to tolerate distortion in power amplifiers. Leafing through recent reviews and specifications shows claims for full-power total harmonic distortion ranging more than three orders of magnitude between individual designs, a wider range than any other parameter.

Admittedly the higher end of this range is represented by subjectivist equipment that displays dire linearity, presumably with the intention of implying that other nameless audio properties have been given priority over the mundane business of getting the signal from input to output without bending it.

Given the juggernaut rate of progress in most branches of electronics this seems to me anomalous, and especially notable in view of the many advanced analogue techniques used in op-amp design; after all power amps are only op-amps with boots on. One conclusion seems inescapable: a lot of power amplifiers generate much more distortion than they need to.

This series attempts to show exactly why amplifiers distort, and how to stop them doing it, culminating in a practical design for an ultra-linear amplifier. It should perhaps be said at the outset that none of this depends on excessively high levels of negative feedback. Many of the techniques described here are also entirely applicable to discrete op-amps,

headphone drivers, and similar circuit blocks. Since we are almost in the twenty-first century I have ignored valve amplifiers.

Since mis-statements and confusions are endemic to audio, I have based these articles almost entirely on my own experimental work backed up with spice circuit simulation; much of the material relates specifically to bipolar transistor output stages, though a good deal is also relevant to mosfet amplifiers. Some of the statements made may seem controversial, but I believe they are all correct. If you think not, please tell me, but only if you have some real evidence to offer.

The fundamental reason why amplifier distortion persists is, of course, because it is a difficult technical problem to solve. An Art proverbially becomes a Science when there are more than seven variables, and since it will emerge that there are seven major distortion mechanisms to the average amplifier, we would seem to be nicely balanced on the boundary of the two cultures. Given so many significant sources of unwanted harmonics, overlaid and sometimes partially cancelling, sorting them out is a non-trivial task.

Make your amplifier as linear as possible before applying NFB has long been a cliché, (one that conveniently ignores the difficulty of running a high gain amp without any feedback) but virtually no dependable advice on how to perform this desirable linearisation has been published. The two factors are the basic linearity of the forward path, and the amount of negative feedback applied to further straighten it out. The latter cannot be increased beyond certain limits or high-frequency stability is put in peril, whereas there seems no reason why open-loop linearity could not, in principle, be improved without limit, leading us to the Holy Grail of the distortionless amplifier. This series therefore takes as its prime aim the understanding and improvement of open-loop linearity. As it proceeds we will accrete circuit blocks to culminate in two practical amplifier designs that exploit the techniques presented here.

How an amplifier (really) works

Figure 1 shows the usual right, trusty and well-beloved power amp circuit drawn as standard is possible. Much has been written about this configuration, though its subtlety and quiet effectiveness are usually overlooked, and the explanation below therefore touches on several aspects that seem to be almost unknown. It has the merit of being docile

enough to be made into a workable amplifier by someone who has only the sketchiest of notions as to how it works.

The input differential pair implements one of the few forms of distortion cancellation that can be relied upon to keep working in all weathers. This is because the transconductance of the input pair is determined by the physics of transistor action rather than matching of variable parameters such as beta; the logarithmic relation between I_c and V_{be} is proverbially accurate over some eight or nine decades of current variation.

The voltage signal at the voltage amplifier stage (hereafter VAS) transistor base is typically a couple of millivolts, looking rather like a distorted triangle wave. Fortunately the voltage here is of little more than academic interest, as the circuit topology essentially consists of a transconductance amp (voltage-difference input to current output) driving into a transresistance (current-to-voltage converter) stage. In the first case the exponential V_{be}/I_c law is straightened out by the differential-pair action, and in the second the global (overall) feedback factor at LF is sufficient to linearise the VAS, while at HF shunt negative feedback (hereafter NFB) through C_{dom} conveniently takes over VAS-linearisation while the overall feedback factor is falling. The behaviour of Miller dominant-pole compensation in this stage is exceedingly elegant, and not at all just a case of finding the most vulnerable transistor and slugging it. As frequency rises and C_{dom} begins to take effect, negative feedback is no longer applied globally around the whole amplifier, which would include the higher poles, but instead is seamlessly transferred to a purely local role in linearising the VAS. Since this stage effectively contains a single gain transistor, any amount of NFB can be applied to it without stability problems.

The amplifier operates in two regions; the LF, where open-loop gain is substantially constant, and HF, above the dominant-pole breakpoint, where the gain is decreasing steadily at 6dB/octave. Assuming the output stage is unity-gain, three simple relationships define the gain in these two regions:

$$LF \text{ gain} = g_m \times \beta \times R_c \quad (1)$$

At least one of the factors that set this (beta) is not well-controlled and so the LF gain of the amplifier is to a certain extent a matter of pot-luck; fortunately this doesn't matter as long as it is high enough to give a suitable level of NFB to eliminate LF distortion. The use of the word 'eliminate' is deliberate, as will be seen later. Usually the LF gain, or HF local feedback-factor, is made high by increasing the effective value of the VAS collector impedance R_c , either by the use of a current-source collector-load, or by some form of bootstrapping.

The other important relations are:

$$HF \text{ gain} = g_m / (\omega \times C_{dom}) \quad (2)$$

Dominant pole freq $P1 =$

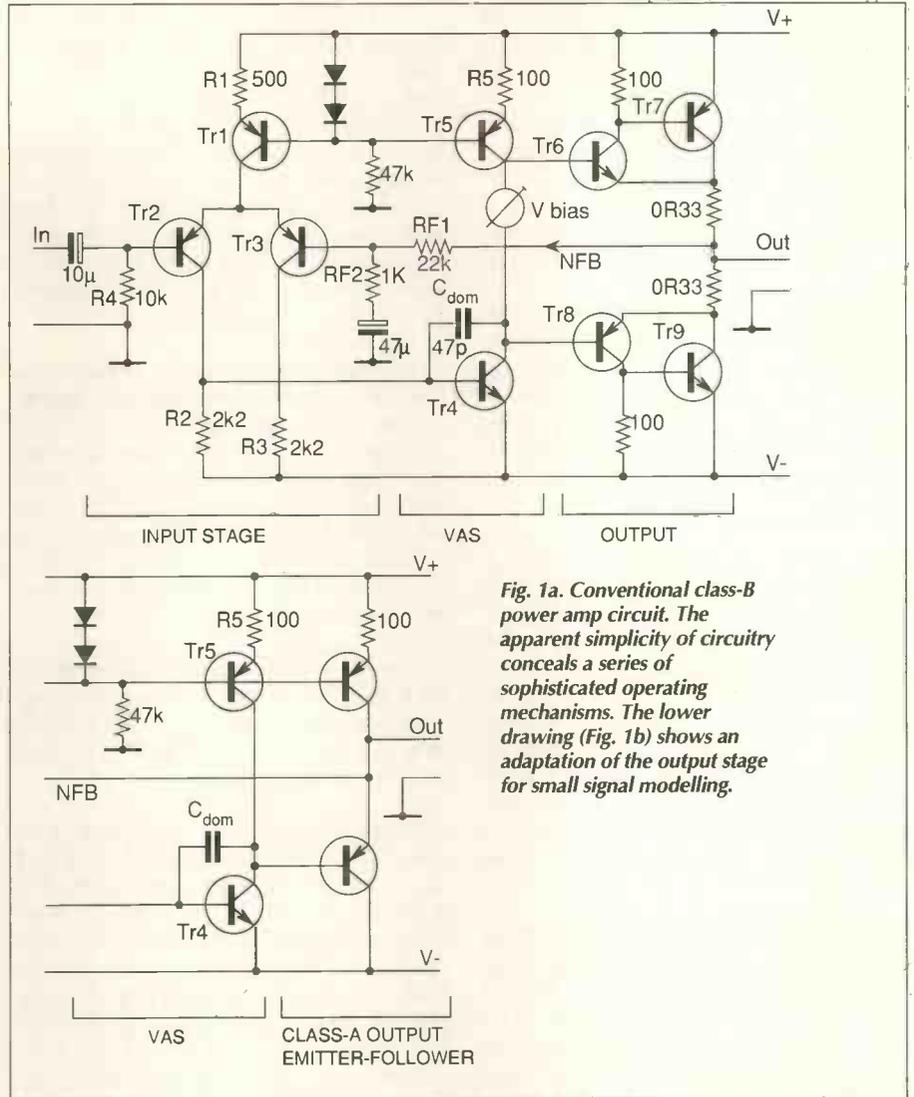


Fig. 1a. Conventional class-B power amp circuit. The apparent simplicity of circuitry conceals a series of sophisticated operating mechanisms. The lower drawing (Fig. 1b) shows an adaptation of the output stage for small signal modelling.

$$1 / (\omega \times C_{dom} \times \beta \times R_c) \quad (3)$$

where $\omega = 2\pi f$

In the HF region, things are distinctly more difficult as regards distortion, for while the VAS is locally linearised, the global feedback-factor available to linearise the input and output stages is falling steadily at 6dB/octave. For the time being we will assume that it is possible to define an HF gain (say N dB at 20kHz) which will assure stability with practical loads and component variations. Note that the HF gain, and therefore both HF distortion and stability margin, are set by the simple combination of the input stage transconductance and one capacitor, and most components have no effect on it at all.

It is often said that the use of a high VAS collector impedance provides a current drive to the output devices, often with the implication that this somehow allows the stage to skip quickly and lightly over the dreaded crossover region. This is a misconception – the collector impedance falls to a few kΩ at HF, due to increasing local feedback through C_{dom} . In any case it is very doubtful if true current drive would be a good thing since calculation shows that a low-impedance voltage drive minimises distortion due to beta-unmatched

output halves¹, and it certainly eliminates distortion mechanism four described later

The seven distortions

In the typical amplifier THD is often thought to be simply due to the Class-B nature of the output stage, which is linearised less effectively as the feedback factor falls with increasing frequency. However the true situation is much more complex as the small-signal stages can generate significant distortion in their own right in at least two different ways. This can easily exceed the output stage distortion at high frequencies. It seems inept to allow this to occur given the freedom of design possible in the small-signal section.

Include all the ills that a class-B stage is prone to and then there are seven major distortion mechanisms.

Distortion in power amplifiers arises from:

1) *Non-linearity in the input stage.* If this is a carefully-balanced differential pair then distortion is typically only measurable at HF, rises at 18dB/octave, and is almost pure 3rd harmonic.

If the input pair is unbalanced (which from published circuitry it usually is) then the HF

distortion emerges from the noise floor earlier. As frequency increases, it rises at 12dB/octave as it is mostly 2nd harmonic.

2) *Non-linearity in the voltage amplifier stage* surprisingly does not always figure in the total distortion. If it does, it remains constant until the dominant-pole freq P1 is reached, and then rises at 6dB/octave. With the configurations discussed here, it is always 2nd harmonic.

Usually the level is very low due to linearising negative feedback through the dominant-pole capacitor. Hence if you crank up the local VAS open-loop gain, for example by cascoding or putting more current-gain into the local VAS/C_{dom} loop, and attend to mechanism four below, you can usually ignore VAS distortion.

3) *Non-linearity in the output stage*, which is naturally the obvious source. This, in a Class-B amplifier, will be a complex mix of large-signal distortion and crossover effects, the latter generating a spray of high-order harmonics, and in general rising at 6dB/octave as the amount of negative feedback decreases. Large-signal THD worsens with 4Ω loads and worsens again at 2Ω. The picture is complicated by dilatory switch-off in the relatively slow output devices, ominously signalled by supply current increasing in the top audio octaves.

4) *Loading of the VAS by the non-linear input impedance of the output stage*. When all other distortion sources have been attended to, this is the limiting distortion factor at LF (say below 2kHz). It is simply cured by buffering the VAS from the output stage. Magnitude is essentially constant with frequency, though overall effect in a complete

amplifier becomes less as frequency rises and feedback through C_{dom} starts to linearise the VAS.

5) *Non-linearity caused by large rail-decoupling capacitors feeding the distorted signals on the supply lines into the signal ground*. This seems to be the reason many amplifiers have rising THD at low frequencies. Examining one commercial amplifier kit, I found that rerouting the decoupler ground-return reduced THD at 20Hz by a factor of three.

6) *Non-linearity caused by induction of Class-B supply currents into the output, ground, or negative-feedback lines*. This was highlighted by Cherry³ but seems to remain largely unknown; it is an insidious distortion that is hard to remove, though when you know what to look for on the THD residual, it is fairly easy to identify. I suspect that a large number of commercial amplifiers suffer from this to some extent.

7) *Non-linearity resulting from taking the NFB feed from slightly the wrong place near where the power-transistor Class-B currents sum to form the output*. This may well be another common defect.

Having set down what Mao might have called The Seven Great Distortions – Fig. 2 shows the location of these mechanisms diagrammatically – we may pause to put to flight a few Paper Tigers. The first is common-mode distortion in the input stage, a spectre that tends to haunt the correspondence columns. Since it is fairly easy to make an amplifier with less than <0.00065% THD (1kHz) without paying any special attention to this, it cannot be too serious a problem. A more severe test is to apply the full output voltage as a

common-mode signal, by running the amplifier as a unity-gain voltage-follower. If this is done using a model (see below for explanation) small-signal version of Fig. 1, with suitable attention to compensation, then it yields less than 0.001% at 8V rms across the audio band. It therefore appears that the only real precaution required against common-mode distortion is to use a tail current-source for the input pair.

The second distortion conspicuous by its absence in the list is the injection of distorted supply-rail signals directly into the amplifier circuitry. Although this putative mechanism has received a lot of attention⁴, dealing with Distortion five above by proper grounding seems to be all that is required... Once again, if triple-zero THD can be attained using simple unregulated supplies and without specifically addressing power supply rejection ratio, (which it reliably can be) then much of the work done on regulated supplies may be of doubtful utility. However, PSRR does need some attention if the hum/noise performance is to be of the first order.

A third mechanism of doubtful validity is thermal distortion, allegedly induced by parameter changes in semiconductor devices whose instantaneous power dissipation varies over a cycle. This would presumably manifest itself as a distortion increase at very low frequencies, but it simply does not seem to happen. The major effects would be expected in Class-B output stages where dissipation can vary wildly over a cycle. However drivers and output devices have relatively large junctions with high thermal inertia. Low frequencies are of course also where the NFB factor is at its maximum.

To return to our list of the unmagnificent seven, note that only Distortion three is *directly* due to O/P stage non-linearity, though numbers 4-7 all result from the Class-B nature of the typical output stage.

The performance

The THD curve for the standard amplifier is shown in Fig. 3. As usual the distortion increases with frequency and, as we shall see later, would give grounds for suspicion if it did not. The flat part of the curve below 500Hz represents non-frequency-sensitive distortion rather than the noise floor, which for this case is at about the 0.0005% level. Above 500Hz the distortion rises at an increasing rate, rather than a constant number of dB/octave, due to the combination of Distortions 1, 2, 3 and 4. (In this case Distortions 5,6 and 7 have been carefully eliminated to keep things simple. This is why the distortion performance looks good already, and the significance of this should not be overlooked). It is often written that having distortion constant across the audio band is a Good Thing. This is a most unhappy conclusion as the only practical way to achieve this with a Class-B amplifier is to *increase* the distortion at LF, for example by allowing the VAS to distort significantly.

It should now be clear why it can be hard to wring linearity out of a snake-pit of contend-

The advantages of being conventional

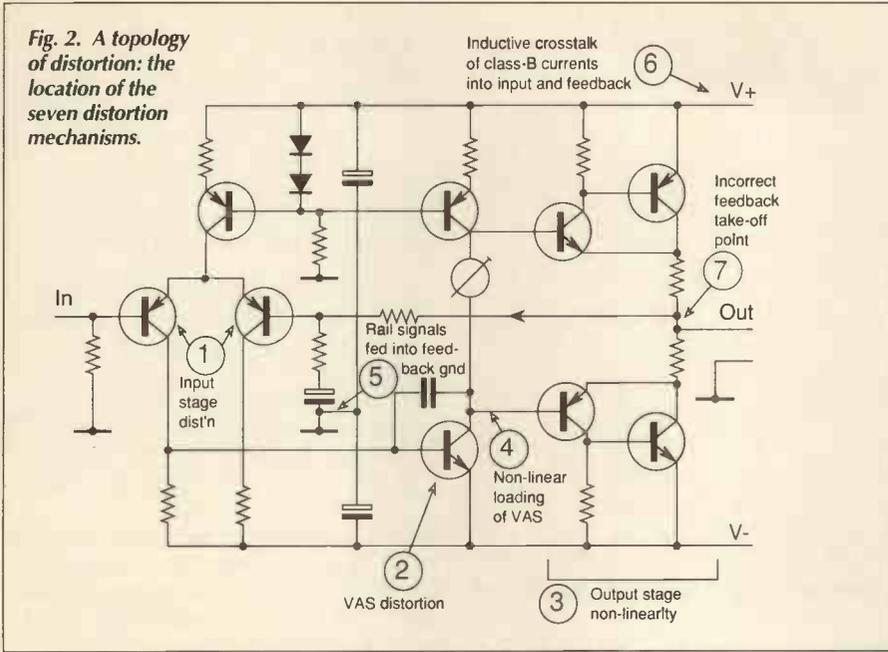
The input pair not only provides the simplest way of making a DC-coupled amp with a dependably small output offset voltage, but can also (given half a chance) completely cancel the 2nd-harmonic distortion which would be generated by a single-transistor input stage. One vital condition must be met: the pair must be accurately balanced by choosing the associated components so that the two collector currents are equal. (The 'typical' component values shown in Fig 1 *do not* bring about this most desirable state of affairs)

The input devices work at a constant and near-equal V_{ce}, giving good thermal balance.

The input pair has virtually no voltage gain so no low-frequency pole can be generated by Miller effect in the Tr₂ collector-base capacitance. All the voltage gain is provided by the VAS

stage, which makes for easy compensation. Feedback through C_{dom} lowers VAS input and output impedances, minimising the effect of input-stage and output stage capacitance. This is often known as pole-splitting²; the pole of the VAS is moved downwards in frequency to become the dominant pole, while the input-stage pole is pushed up in frequency.

The VAS Miller compensation capacitance smoothly transfers NFB from a global loop which may be unstable, to the VAS local loop that cannot be. It is quite wrong to state that *all* the benefits of feedback are lost as the frequency increases above the dominant pole, as the VAS is still being linearised. This position of C_{dom} also swamps the rather variable C_{cb} of the VAS transistor.



ing distortions. A circuit-value change is likely to alter at least two of the distortion mechanisms, and probably change the open-loop gain as well. In the coming articles I shall demonstrate how each of these mechanisms can be measured and manipulated separately.

Determining open-loop linearity

Improving something demands its measurement, and so it is essential to examine the open-loop linearity of typical power-amp circuits. This cannot in general be done directly, so it is necessary to measure the NFB factor and calculate open-loop distortion from the usual closed-loop data. It is assumed that the closed-loop gain is fixed by operational requirements.

Finding the feedback-factor is at first sight difficult, as it means determining the open-loop gain. The standard methods for measuring op-amp open-loop gain involve breaking feedback-loops and manipulating closed-loop gains, procedures that are likely to send the average power-amplifier into fits. However, the need to measure this parameter is inescapable, as a typical circuit modification – eg changing the value of R_2 – will change the open-loop gain as well as the linearity, and to prevent total confusion it is necessary to keep a very clear idea of whether the observed change is due to an improvement in open-loop linearity or merely because the open-loop gain has risen. It is wise to keep a running check on the feedback-factor as work proceeds, and so the direct method of open-loop gain measurement shown in Fig. 4 was evolved.

AUDIO PRECISION POWRAMP THD+N(%) vs FREQ(Hz)

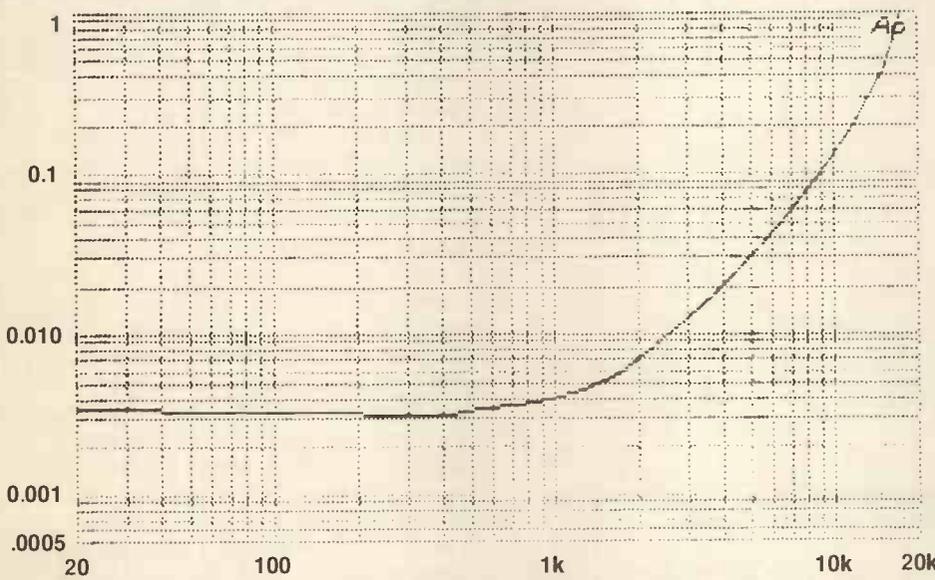
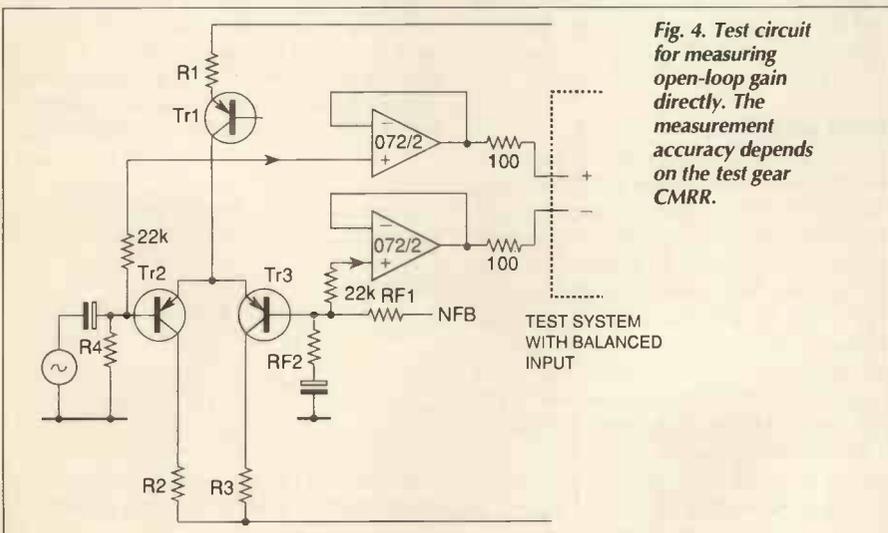


Fig. 3. The distortion performance of the class-B amplifier shown in Fig. 1a.



Direct open-loop gain measurement

Since the amplifier shown in Fig. 1 is a differential amplifier, its open-loop gain is simply the output divided by the voltage difference between the inputs. If the output voltage is kept effectively constant by providing a swept-frequency constant voltage at the +ve input, then a plot of open-loop gain versus frequency is obtained by measuring the error-voltage between the inputs, and referring this to the output level. This yields an upside-down plot that rises at HF rather than falling, as the differential amplifier requires more input for the same output as frequency increases, but the method is so quick and convenient that this can be lived with. Gain is plotted in dB with respect to the chosen output level (+16dBu in this case) and the actual gain at any frequency can be read off simply by dropping the minus sign. Fig. 5 shows the plot for the amplifier in Fig 1.

The HF-region gain slope is always 6dB/octave unless you are using something special in the way of compensation and, by the Nyquist rules, must continue at this slope until it intersects the horizontal line representing the feedback factor provided that the amplifier is stable. In other words, the slope is not being accelerated by other poles until the loop gain has fallen to unity, and this provides a simple way of putting a lower bound on the next pole P2; the important P2 frequency (which is usually somewhat mysterious) must be above the

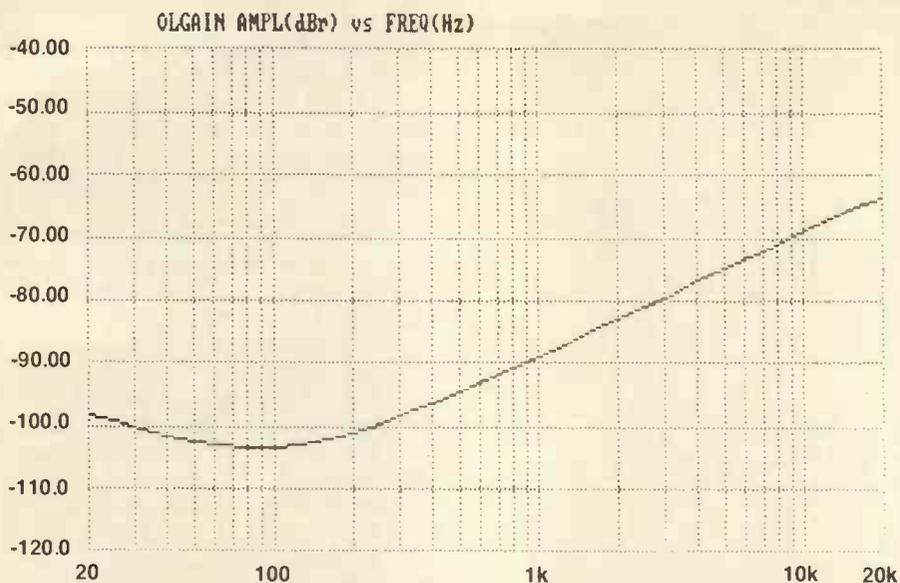


Fig. 5. Open-loop gain versus frequency plot for Fig. 1. Note that the curve rises as gain falls, because the amplifier error is the actual quantity measured.

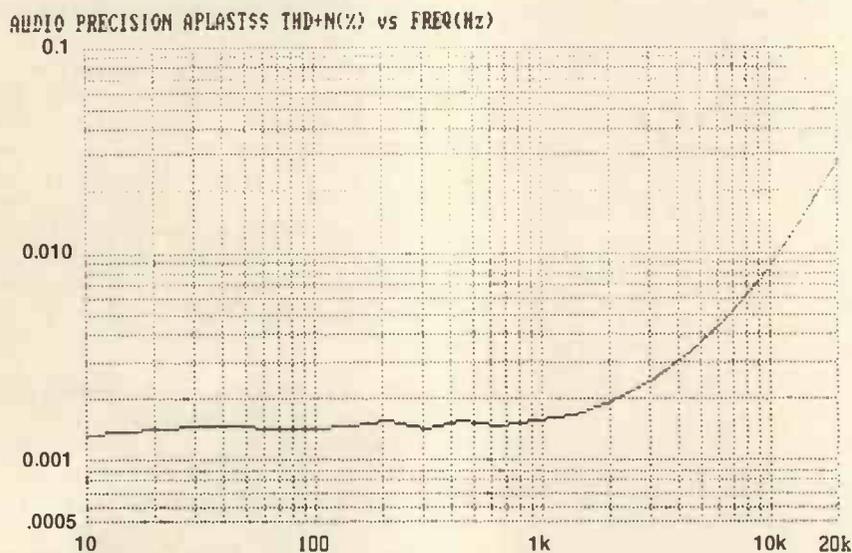


Fig. 6. The distortion from a model amplifier, produced by the input pair and the voltage amplifier stage. Note increasing slope as input pair distortion begins to add to VAS distortion.

intersection frequency if the amplifier is seen to be stable.

Given test gear with a sufficiently high common-mode-rejection-ratio balanced input, the method of Fig 4 is simple; just buffer the differential inputs from the cable capacitance with *TL072* buffers, placing negligible loading on the circuit if normal component values are used. Be particularly wary of adding stray capacitance to ground to the -ve input, as this directly imperils amplifier stability by adding an extra feedback pole. Short wires from power amplifier to buffer IC can usually be unscreened as they are driven from low impedances.

The test gear input CMRR defines the maximum open-loop gain measurable; I used an Audio Precision System-1 without any special alignment of CMRR. A calibration plot can be produced by feeding the two buffer inputs from the same signal; this will probably be found to rise at 6dB/octave, being set by the inevitable input asymmetry. This must be low enough for amplifier error signals to be above

it by at least 10dB for reasonable accuracy. The calibration plot will flatten out at low frequencies, and may even show an LF rise due to imbalance of the test gear input-blocking capacitors; this can make determination of the lowest pole P1 difficult, but this is not usually a vital parameter in itself.

Model amplifiers

The first two distortions on the list can dominate amplifier performance and need to be studied without the complications introduced by a Class-B output stage. This can be done by reducing the circuit to a model amplifier that consists of the small-signal stages alone, with a very linear Class A emitter-follower attached to the output to allow driving the feedback network. Here 'small-signal' refers to current rather than voltage, as the model amplifier should be capable of giving a full power-amp voltage swing, given sufficiently high rail voltages. From Fig. 2 it is clear that this will allow study of Distortions 1 and 2 in

Glossary

Several abbreviations will be used throughout this series to keep its length under control.

LF...Relating to amplifier action below the dominant pole, where the open-loop gain is assumed to be essentially flat with frequency.

HF...Amplifier behaviour above the dominant pole frequency, where the open-loop gain is usually falling at 6dB/octave.

I/P...Input.

P1...The first open-loop response pole, and its frequency in Hz.

NFB...Negative feedback.

isolation, and using this approach it will prove relatively easy to design a small-signal amplifier with negligible distortion across the audio band. This is the only sure foundation on which to build a good power amplifier.

A typical plot combining Distortions 1 and 2 from a model amp is shown in Fig. 6, where it can be seen that the distortion rises with an accelerating slope, as the initial rise at 6dB/octave from the VAS is contributed to and then dominated by the 12dB/octave rise in distortion from an unbalanced input stage.

The model can be powered from a regulated current-limited PSU to cut down the number of variables, and a standard output level chosen for comparison of different amplifier configurations. The rails and output level used for the results in these articles was $\pm 15V$ and +16dBu. The rail voltages can be made comfortably lower than the average amplifier HT rail, so that radical bits of circuitry can be tried out without the creation of a silicon cemetery around your feet. It must be remembered that some phenomena such as input-pair distortion depend on absolute output level, rather than the proportion of the rail voltage used in the output swing, and will be worse by a mathematically predictable amount when the real voltage swings are used. The use of such model amplifiers requires some caution, and cannot be applied to bipolar output stages whose behaviour is heavily influenced by the sloth and low current gain of the power devices. As another general rule, if it is not possible to lash on a real output stage quickly and get a stable and workable power amplifier; the model may be dangerously unrealistic.

Continued next month

Douglas dedicates this series to Thalia, his girlfriend, who died a year ago.

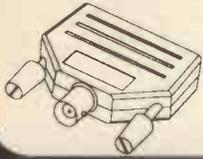
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2. Feucht *Handbook of Analog Circuit Design* Academic Press 1990. p256 (Pole-splitting)
3. Cherry *A New Distortion Mechanism in Class-B Amplifiers* JAES May 1981, p327.
4. Duncan *PSU Regulation Boosts Audio Performance*. *EW + WW* Oct 1992, p818.

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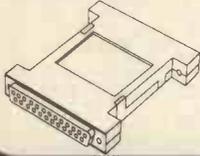
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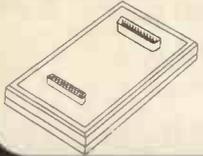
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Source JUNE 1991 Practical Electronics

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United approach to

CAD

How effectively do Supercad and TopSpice operate together? Probably better than they do apart
says
John Anderson.

Combining separate, but complementary, schematic capture and *Spice* simulation packages has a lot to recommend it – reducing errors through transcribing the model definition file is one example.

TopSpice *Spice* simulator and *Supercad* schematic capture editor are two such products from different companies. Combined, they offer an integrated route from schematic capture through to simulation. But to judge how well they cooperate, it is worth looking first at each in isolation, and then seeing how they work together.

Looking first at *Supercad*, the *Windows*-based schematic capture program, installation is fast and straightforward, invoked from either *Windows* or *dos*. Information provided in the manual appears to be just about adequate, bearing in

mind that there is some limited help available as part of the normal *Windows* help system. Inclusion in the documentation of application note “Using *Supercad* with *IsSpice*” gives a clue that *Supercad* is targeted at a variety of add on third party products.

The basic *Supercad* display follows the standard *Windows* format of FILE, EDIT etc together with a short tool-bar. At times the program really shows its insular US market base, offering only US paper sizes “A” through to “E, and US spellings in the menus.

Editing is based around objects. Objects are placed, selected, and once selected can be moved, mirrored, rotated and erased. A wide range of library symbols is included with the package and full details are given about how the symbol libraries are built, so that custom ones can be added.

Unfortunately the libraries make up nearly 1000 tiny files hungry for disc space and it would have been so much better if the components had been compiled into a single library file.

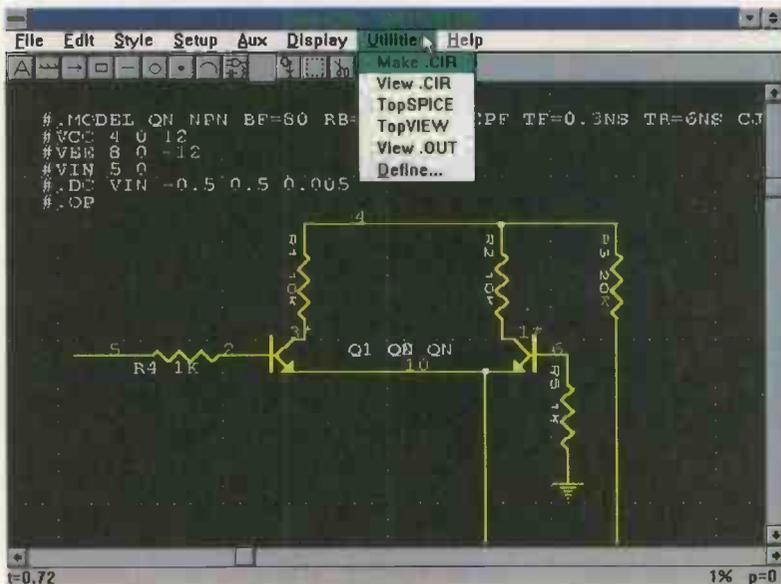
Link to *TopSpice*

One point of distinction (the only one?) for this very average schematic capture package is that it can be used to generate data for *TopSpice*. This is accomplished by a combination of the schematic and text strings entered directly onto the schematic itself. It seems rather strange to have *Spice* commands as part of the circuit documentation: but on second thoughts – why not?

Entry to *TopSpice* is through the utilities menu, a list of commands which can be user defined. In this case, there is a MAKE_CIR command to produce a netlist from the schematic and embedded text, a VIEWER, which is simply a command to view the resulting netlist through the *Windows* notepad and *TopSpice* and *TopView* entry points.

TopSpice itself is marketed as a mixed-mode circuit simulator: the simulator will work from the systems level

Embedded *Spice* commands



through to analogue component level, and has an event driven logic simulator. It is presented as the sensible alternative to *Spice* simulation of logic circuits, reducing simulation durations by twenty times.

In reality no one would simulate a complex logic circuit using *Spice* analogue methods unless looking for a pseudo-analogue effect such as dynamic hazards. But the system is capable of simulating a combined digital and analogue system described in a single *Spice* file (using *Spice2G6* standard).

Supercad actually executes *TopSpice* as a dos command with the *Spice* net filename appended to the command. The result is that *TopSpice* runs the simulation immediately.

The screen presented during the simulation is not really useful, comprising sets of numbers representing the various nodes of the circuit changing at high speed as the simulation progresses. But it is worth saying that the simulation does happen at high speed and even quite complex analogue simulations are completed in only a few seconds.

When the simulation is complete, *TopSpice* automatically switches to its companion program *TopView*, to display the results. Graphs are automatically scaled and detailed with the node names, though the editing scheme that allows alteration of the how the graph is presented is somewhat cranky. A neat touch in the graphical display program is the facility to run the cursor along a graph trace (while being glued to it) with the values of the cursor position being reported.

All the data from the simulation is saved in an ascii (large) output file. This forms a useful part of design records, as the graphical performance data does not present all the variables and settings used in the simulation.

A multi-window dos-based editor can be invoked from within *TopSpice* to modify the circuit description files, and the editor itself is fast and neat - very similar to Borland's *Turbo* user interface.

Extended language

Several variants of the *Spice* language are around, and this version includes a variety of enhancements over the original 1975 Berkley program. Additions include extended syntax, library files, parameter passing and many new device models.

TopSpice offers a further degree of design description abstraction through a top down systems simulation approach.

The net effect is that the language has become quite complex, with the manual running to several hundred pages, describing the commands, their syntax and semantics. So take it as a warning that although some of the grind associated with producing a *Spice* net from a circuit diagram has been eased, this does not necessarily address all the aspects of control of the simulation. Clearly, getting the best from *TopSpice*, as with learning any language, needs significant learning effort.

On the plus side, the original *Spice* is contained in the kernel of the program, so those with only limited knowledge of *Spice* can get the simulator running quickly.

Performance libraries

TopSpice has its own set of libraries describing the performance of components. The range and number of devices included is impressive and includes, by manufacture number, many transistors and fets; TTL in standard, LS and HC for-

Window with generated *Spicenet*

Spice simulation screen

SYSTEM REQUIREMENTS

Supercad

Windows v3.0 or later
Windows standard mode
80286 processor
1Mbyte ram
Windows 386 Enhanced mode provides access to Windows virtual memory and multi tasking.
Windows-supported pen plotter or printer
Gerber format photoplotter.

TopSpice

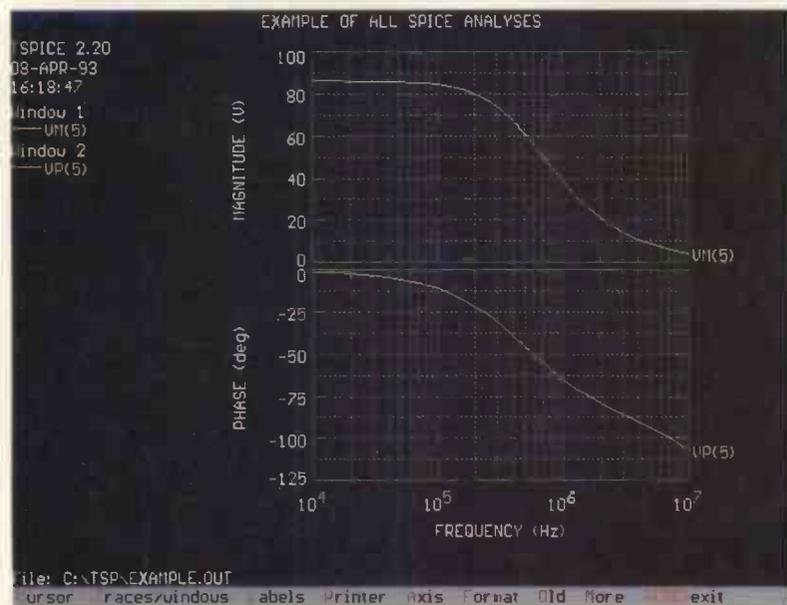
Dos 3.3 or later
520K memory
3Mbytes of hard disc
Maths processor (optional)
EMS (optional)
Dot matrix, HPGL plotter or LaserJet printer

TopSpice Plus is also available to run on 386/486 machines with at least 2Mbyte memory free.

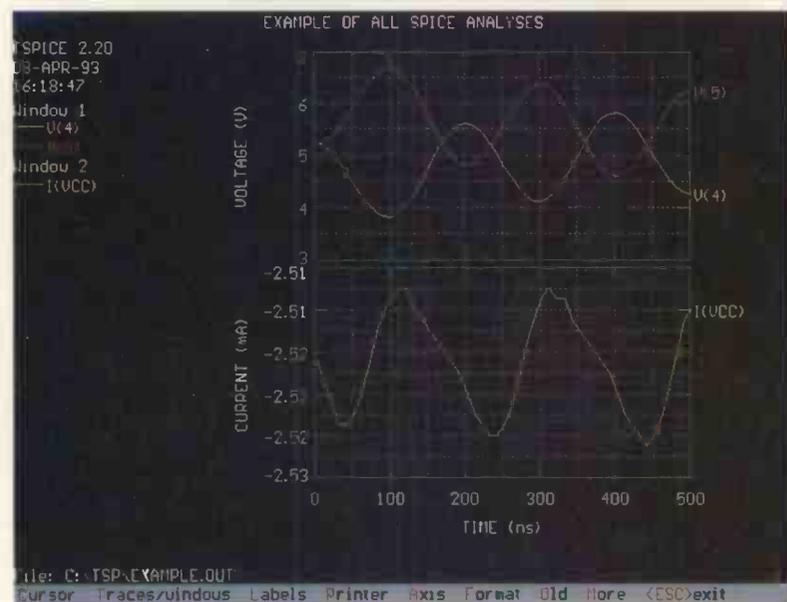
Reviewed on a 33MHz 486 machine with 8Mbytes ram



TopSpice main menu



Frequency response



Transient analysis

mats; SCRs and a set of libraries for components from seven US semiconductor manufacturers, including Texas, Harris and Analogue Devices.

No models are supplied for European or Japanese manufacturers, although these may be available from the manufacturer directly or may be derived from data sheet parameters.

Combined solution

Despite differences in modes of operation of the two programs, their combination does offer an integrated route from schematic capture through to simulation.

Supercad on its own is a rather limited US product produced for the home market. It uses Windows features reasonably well, but the symbols in the libraries look rather amateurish and the whole editor has a not quite complete product look about it.

TopSpice is a different matter. Simulation performance is fast and there is potential for real improvements in engineer productivity using this tool – once the language is mastered.

On the down-side, its mixed mode simulation capability, ranging from components to systems, may look very attractive on paper but its final performance is less impressive, and there is a poor user interface which looks as if it was designed a decade ago.

Using both products together does work. But I'm sure that this can only be seen as at best a stop gap until a Windows version of TopSpice becomes available. ■

SCHEMATIC CAPTURE/SPICE

Spice, (Simulation Program with IC Emphasis) modelling dates back to work done at Berkley in the mid 1970s, producing a simulation system which was essentially text based. In the practical application of electronic simulation it is essential that there is a direct correspondence between schematic and model, otherwise there is a distinct possibility of errors in transcribing the model definition file – hence the need to marry schematic capture and Spice simulation.

SUPPLIER DETAILS

Supercad for Windows by Mental Automation Inc and TopSpice by Penzar Development (USA). TopSpice £295 TopSpice Plus £695 Supercad £200 Available from CRaG Systems, 8 Shakespeare Road, Thatcham, Newbury, Berks RG13 4DG. Tel/Fax: 0635 873670

FREE SOFTWARE OFFER

To obtain your FREE working version of the TopSpice mixed analogue and digital simulator, just send a formatted 1.44MB HD 3.5in disk and a stamped addressed envelope to: **Electronics World Offer, CRaG Systems, 8 Shakespeare Road, Thatcham, Newbury, Berks RG13 4DG. Phone 0635-873670.** Copied disks will be dispatched by return of post.

If you are not already familiar with SPICE based simulation, then for only £34.95+£2p&p you can buy from CRaG Systems the excellent book by John Keown **PSpice and Circuit Analysis** containing lots of circuit examples to run on the free software. It explains in detail how to use the various Spice analysis and modelling commands.

CIRCUIT IDEAS

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

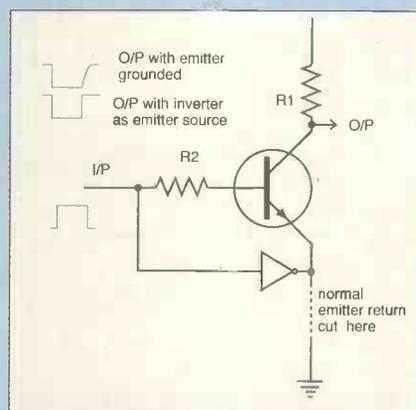
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Fast turnoff, open collector

Directly interfacing logic gates to open collector outputs generally corrupts output square wave trailing edges. When sluggish turn-off of a saturated open collector transistor is a problem, introduce an inverter into the emitter circuit which nips off emitter current to give a fast trailing edge.

Bob Philp
 L-5215
 Luxembourg



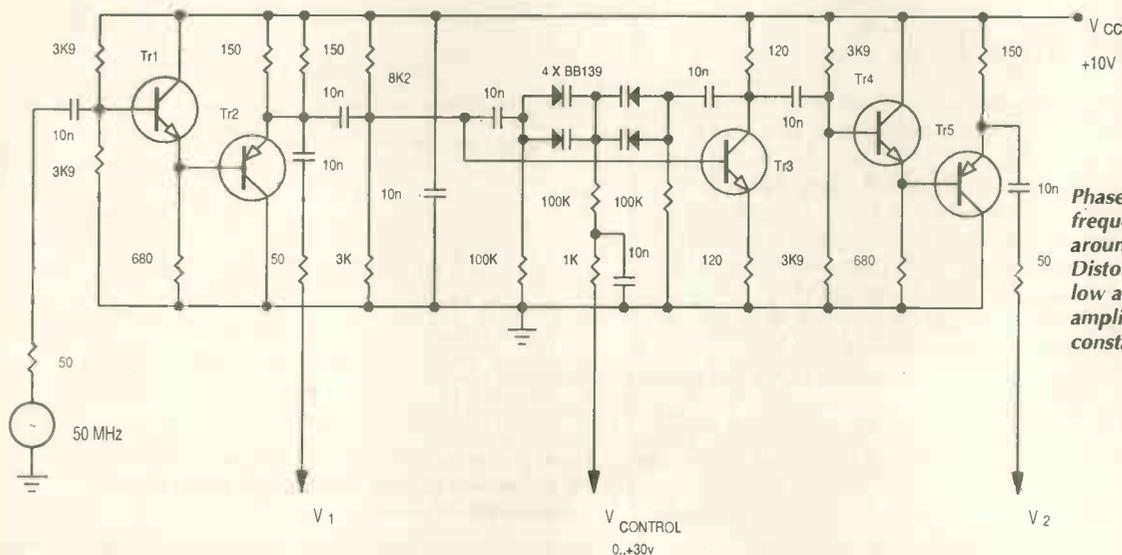
50MHz phase shifter

Phase difference between the two outputs of the 50MHz phase shifter is controlled by the DC input at $V_{control}$, V_1 leading V_2 . Its transfer characteristic varies somewhat for slight variations in

frequency. Driving 50Ω loads, output varies less than 3dB over the 30V control-voltage excursion and distortion remains reasonably low even when the control voltage is near zero volts. Layout needs

watching: avoid stray coupling and parasitic capacitances.

Alexandru Ciubotaru
 University of Texas at Arlington
 Texas



Phase shifter for frequencies around 50MHz. Distortion is low and amplitude fairly constant.

Low-noise M-C head amplifier

Based on an amplifier design published in the September 1992 CI, this new design possesses two further features to reduce noise: parallel input-transistor connection and inclusion of the cartridge internal resistance in the feedback loop. The measurement setup is shown to allow verification of the amplifier's performance. Components are Ukrainian types, but it is only necessary to use a low-noise input device and medium-power HF output transistors.

Equivalent unweighted noise voltage is -140.5dB relative to 0.775V RMS at 73nV over the range $200\text{Hz}-20\text{kHz}$; input transistors with a lower R_B should improve this figure, the theoretical minimum being 33nV and -147.4dB , which is equivalent to the thermal noise of a 3.3Ω resistor.

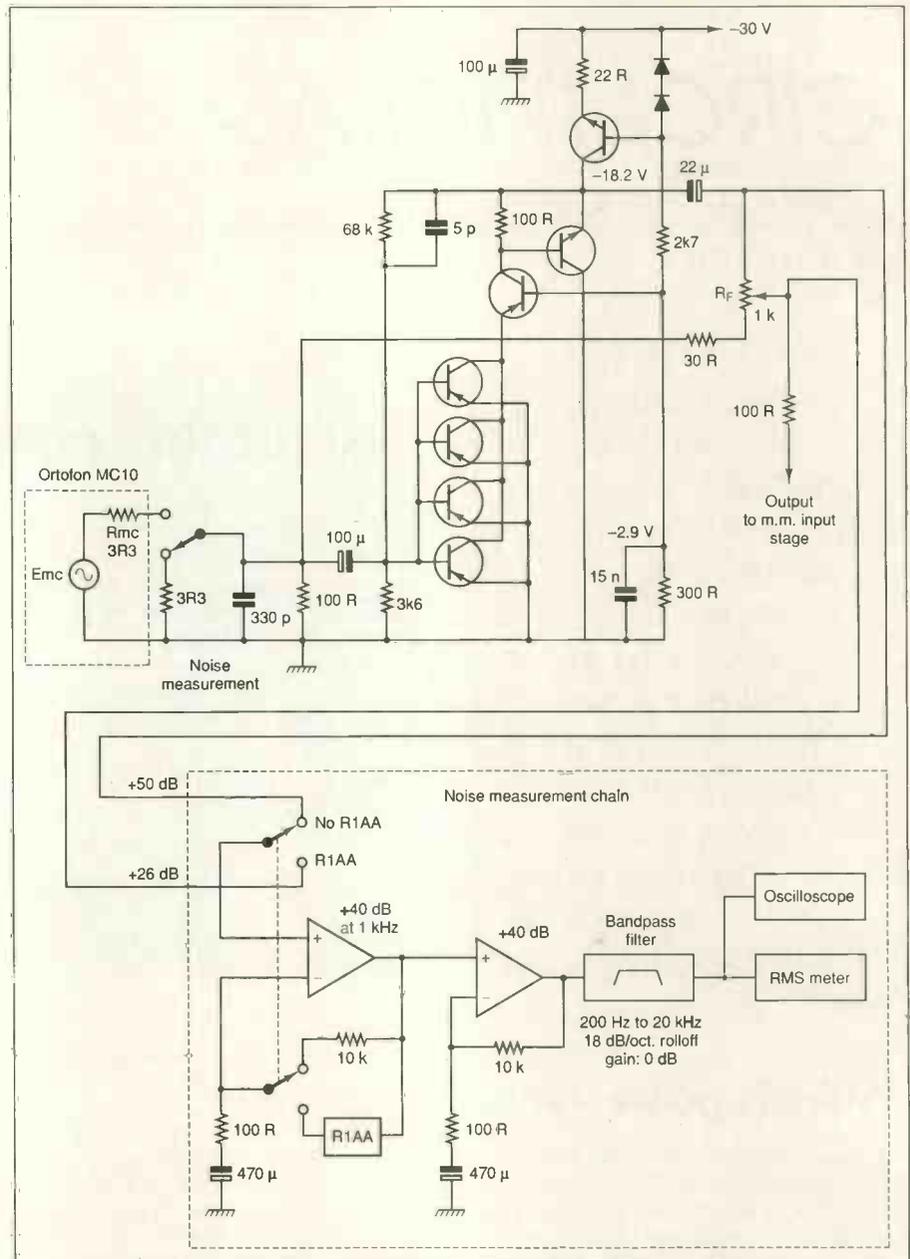
Gain, set by R_F , is 26dB for the Ortofon MC10; at this gain setting and with RIAA equalisation in the measurement chain, amplifier output noise is $0.78\mu\text{V RMS}$, which is -76dB relative to that of a 5mV moving-magnet input stage.

The $+0/-1\text{dB}$ frequency response is $20\text{Hz}-200\text{kHz}$, slew rate $150\text{V}/\mu\text{s}$, open-loop gain 5000 up to 100kHz and total harmonic distortion $<0.05\%$ for 2V RMS output in the audio range (still only 0.08% at 200kHz).

Vladimir Katkov

Priluki
Ukraine

Moving-coil head amplifier based on an earlier amplifier design, giving 0.05% THD and -140.5dB noise level referred to 0.775V RMS , shown together with set up for test measurement purposes.

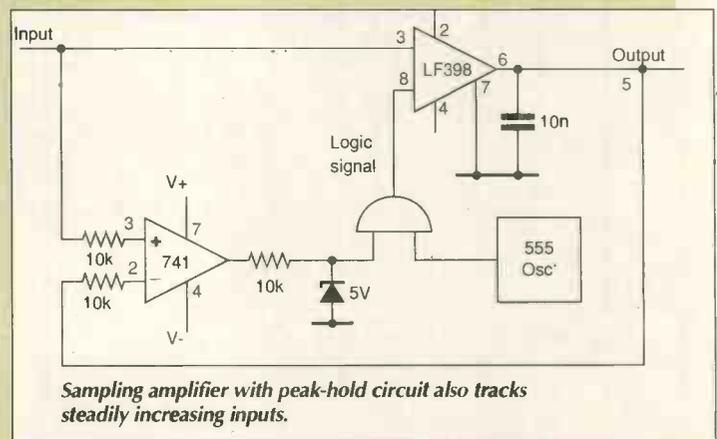


Peak hold amplifier

Three ICs and a spare AND gate with some passive components make a simple circuit to detect and hold the highest voltage encountered.

Sample-and-hold amplifier 398 samples the input, but only when the control logic on pin 8 generated by the 741 comparator is high. In the case of a continuously increasing input, the output should change at regular intervals; the 555 oscillator ANDED with the comparator output allows both operating regimes to be met. Possible loss of rapidly changing data when the oscillator output is low is avoided as much as possible by ensuring a 50% oscillator duty cycle. Use a tantalum or polyester capacitor to avoid "droop".

K V Madanagopal
University of Roorkee
Uttar Pradesh
India



Sampling amplifier with peak-hold circuit also tracks steadily increasing inputs.

Current-controlled inductance

The XR13600 contains two transconductance amplifiers, complete with controlled-impedance buffers, and will simulate a single-ended inductor; two would simulate a double-ended L , but the circuit might be considered a little too much.

In the circuit shown,

$$i_{out2} = gV_{out1} \quad (1)$$

$$V_{out1} = \frac{1}{sC} I_{out1} = \frac{g}{sC} V_{in} \quad (2)$$

Since $V_{in} = V_{out2} R_A / (R_A + R_B)$, from (2),

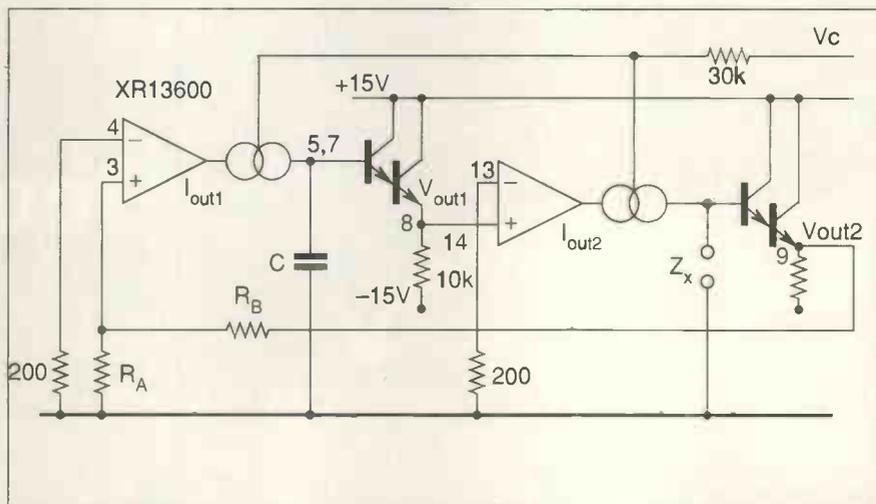
$$V_{out1} = \frac{g}{sC} \frac{R_A}{R_A + R_B} V_{out2} \quad (3)$$

Substitute (3) into (1),

$$I_{out2} = \frac{g^2}{sC} \frac{R_A}{R_A + R_B} V_{out2}$$

Hence,

$$Z_x = \frac{V_{out2}}{I_{out2}} = s \frac{C}{g^2} \frac{R_A + R_B}{R_A}$$



$$L = \frac{C}{g^2} \left(1 + \frac{R_B}{R_A} \right)$$

Since $g = 19.2 I_C$, L is inversely proportional to I_C^2 , which is used to adjust the value of

inductance.
Kamil Kraus
 Rokycany
 Czech Republic

Simple frequency indicator

Using the circuit shown, frequencies up to about 200kHz can be measured by means of a multimeter, with no need for a power supply. The circuit fits into a probe case.

A diode-pump with a resistive load is a simple and reasonably linear method of frequency-to-current conversion, the relevant components here being $C_1 D_1 D_2$, with integrating capacitor C_2 . Average current seen by the $20\mu A$ meter is $f C_1 V$. To obtain $1\mu A$ at 10kHz with $V = 1V$, C_1 is 100pF; in practice V is about 1.3V and C_1 75pF. Linearity is good from about 50Hz to 10kHz, diverging by about 5% high below and 5% low above these frequencies, and can be considered a straight line from the origin to 100kHz/10 μA .

Fet T_{r1} - a symmetrical current source due to the effect of $R_{a,b}$ - and diodes $D_{a,b}$ form a clipper to reduce the effect of input amplitude changes. The graph shows this effect for an input change 350mV-10V RMS and can be used to correct the meter reading; simply divide the meter reading by G .

CJD Catto
 Cambridge

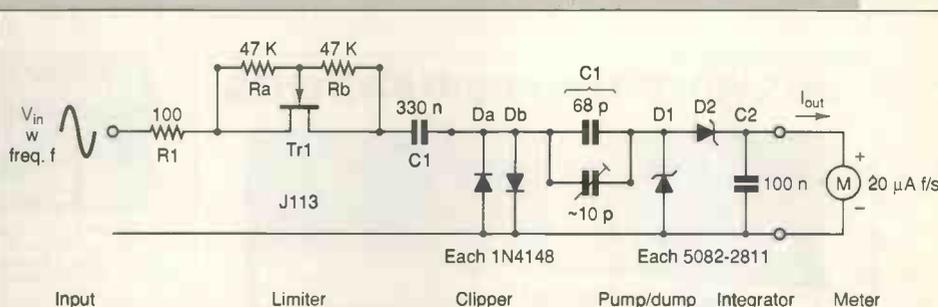


Fig.1. Very simple frequency indicator working up to a few hundred kilohertz with reasonable linearity.

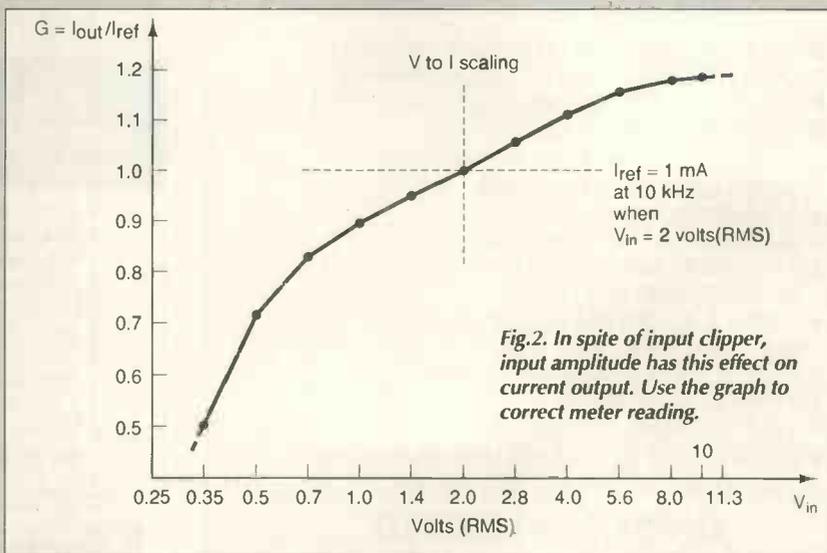


Fig.2. In spite of input clipper, input amplitude has this effect on current output. Use the graph to correct meter reading.

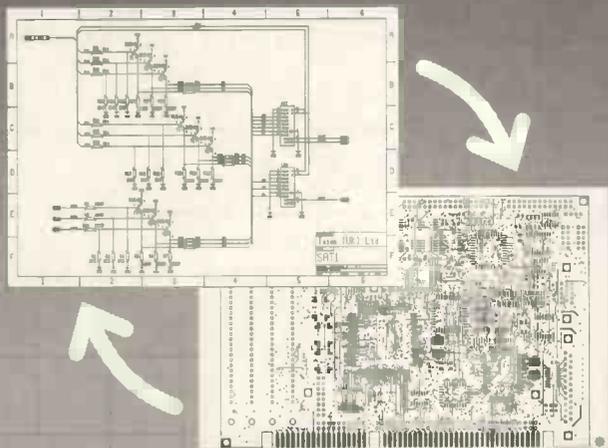
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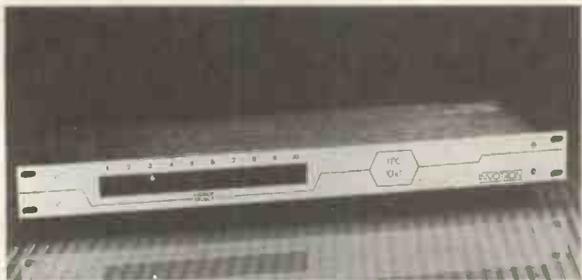
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CIRCLE NO. 119 ON REPLY CARD

Combing technology for a higher performing pal

The pal TV broadcast system has been in use in the UK since the nineteen-sixties, and seems set to continue for a long time yet. Today's receivers will benefit from decoding techniques which until recently could only be implemented in expensive hardware. In this article John Watkinson looks at advanced pal decoding theory and its implementation in new chips.

Pal video was selected for use in UK broadcasting in preference to NTSC or Secam on the introduction of the colour broadcasting. Pal itself is an enhancement of the NTSC system which dates from the 1950s. But broadcasting standards have to be around for a long time owing to the vast quantities of receiving equipment involved. There are currently a large number of developments occurring in TV formats and out of these a future TV standard will eventually emerge. This may not be for some time, but there is little doubt that it will be entirely digital and will rely heavily on data reduction. But even if a

new TV system were to be standardised tomorrow, pal would carry on for a long time after that.

In this context there is justification for improving the picture quality of pal receivers. TV displays cannot accept a composite signal, but need component, or RGB, signals. Accurate pal decoding is a complex process and, until recently, only high-cost broadcast decoders approach the performance theoretically available.

These techniques are now available in little more than a single chip.

The difficulty in properly decoding pal is the

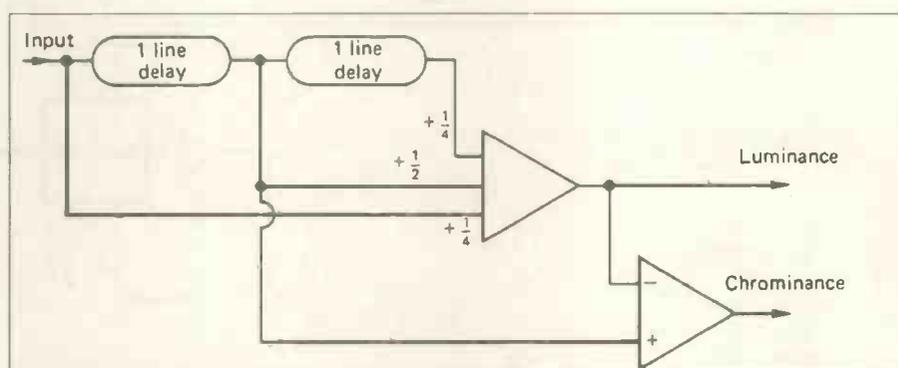


Fig. 1. Simple digital line comb filter. Although the spectral response is ideal with minimal cross-colour and cross-luminance, there are shortcomings.

Details from a TV picture before and after adaptive comb filtering. Conventional colour signal extraction removes much high frequency luma information (left) as shown by the smeared verticals. The right hand picture uses digital filtering (ITT).

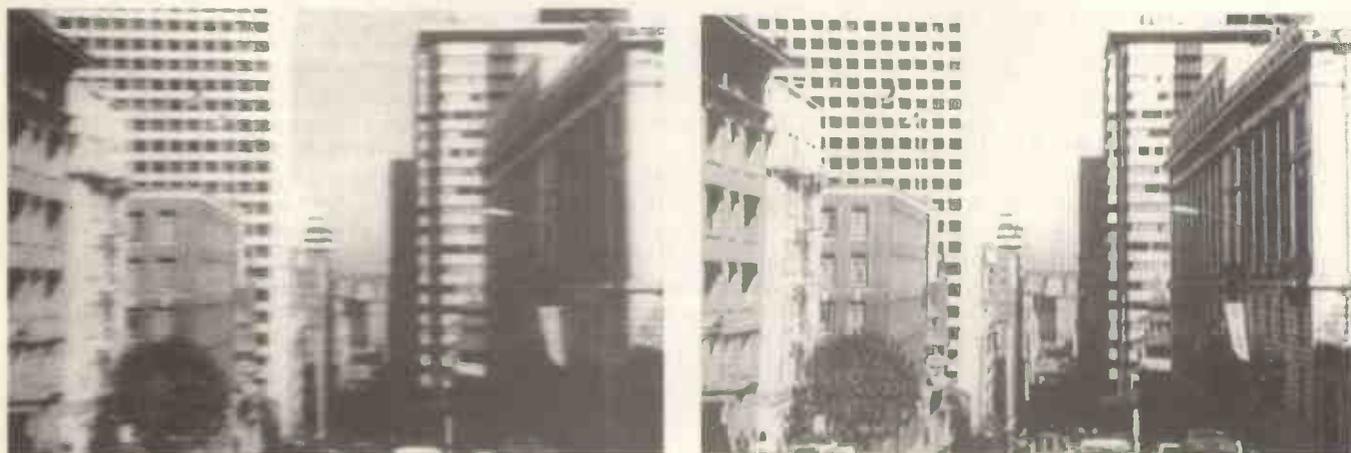
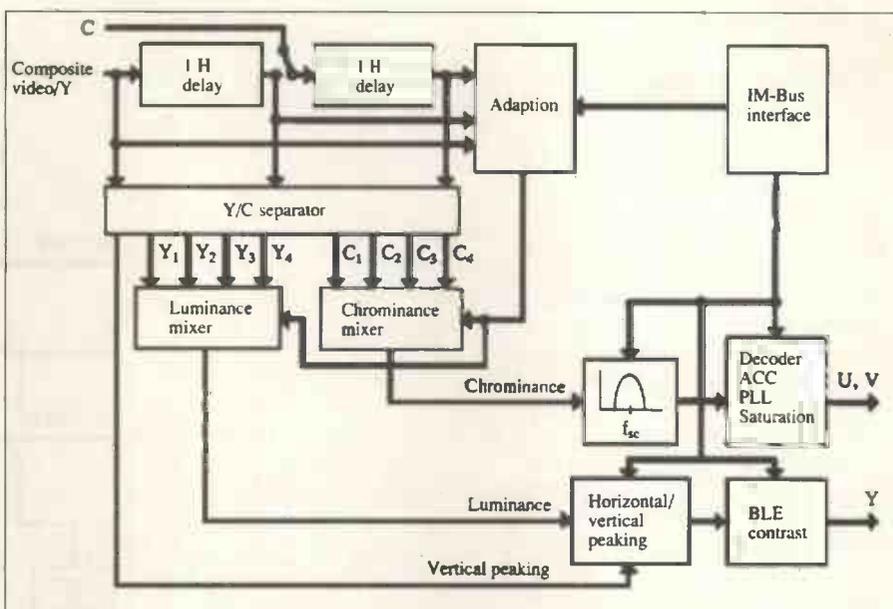


Fig. 3 Block diagram of the ITT ACVP2205 comb filter chip. Like the Motorola device, it has a composite analogue input with an ADC running at $4F_{sc}$ and two one-line delays.

olution loss of such a filter can be largely restored by running the comb filter only in a passband centred around subcarrier. Within the passband, combing is used to remove luminance from the chroma. This chroma is then subtracted from the composite input signal to leave luminance. Below the passband, the entire input spectrum is passed as luminance and the vertical resolution loss is restored¹.

Figure 2 shows a block diagram of the Motorola MC141625 comb filter chip. It will be seen that each tap of the line comb filter is in series with a bandpass filter. Thus the combed response is restricted to the chroma passband. Chroma is subtracted from the com-



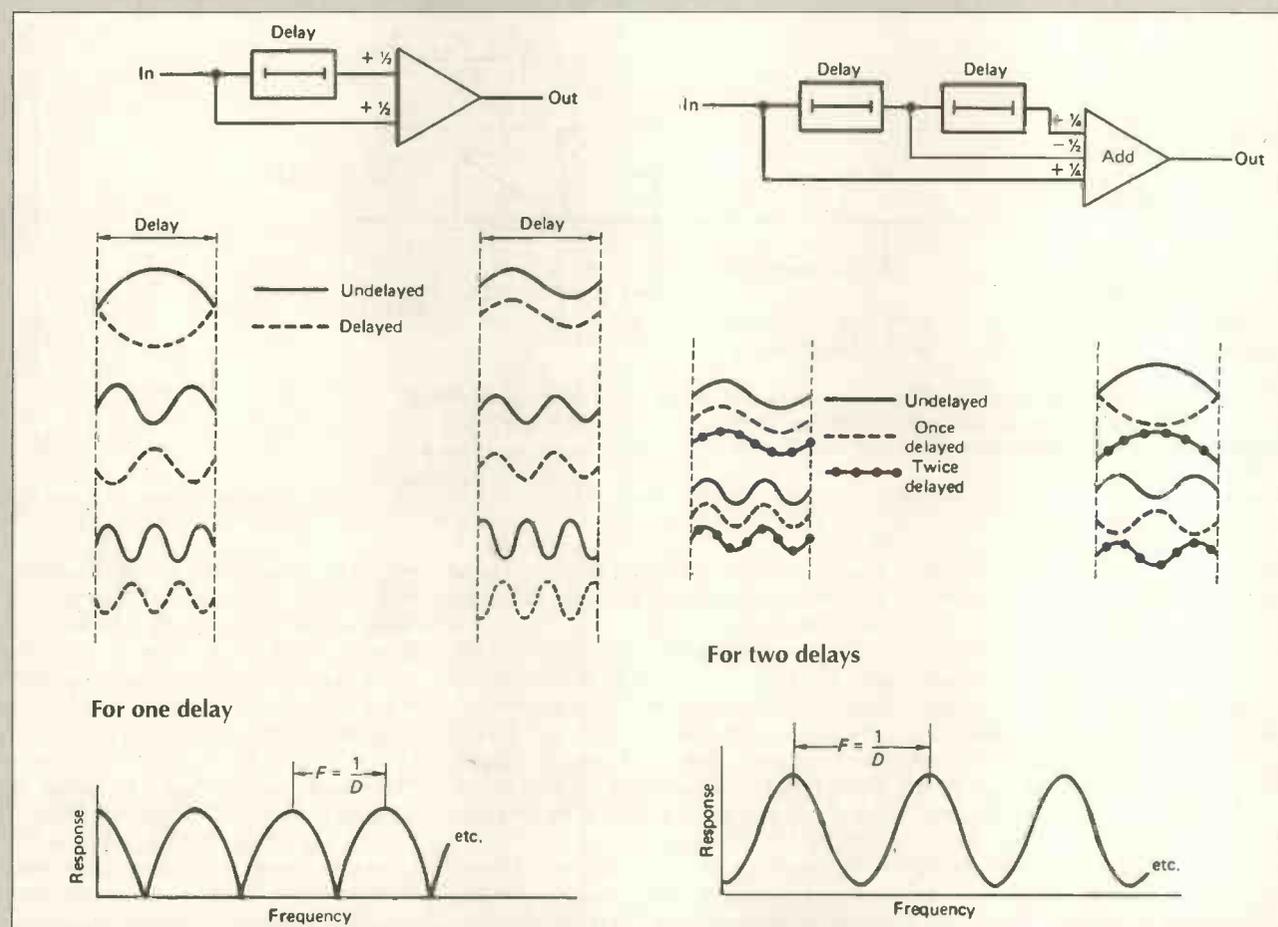
Comb filtering

The drawing shows a simple comb filter consisting of a ram delay and an adder. The phase difference between the delay output and input is a function of the signal frequency. When the delay exceeds the signal period, there will be a number of evenly spaced frequencies which suffer the same phase shift and which therefore have the same gain. The frequency response is repetitive, resembling the teeth of a comb. A

sharper response can be obtained with two delays and a three input adder as shown in Fig. 1. The frequency response is a cosinusoid with the peaks spaced at the reciprocal of the delay. The spectrum of composite video is interleaved at multiples of line rate and so a comb filter will need to use delays of one horizontal line period.

Chroma and luminance have a fundamental spacing of 25Hz due to the

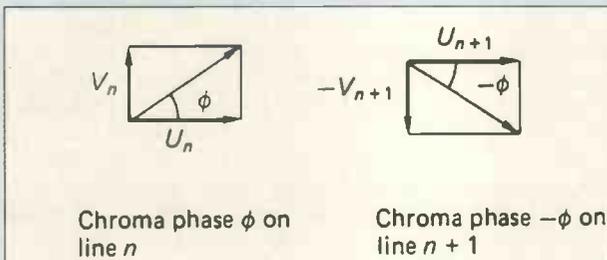
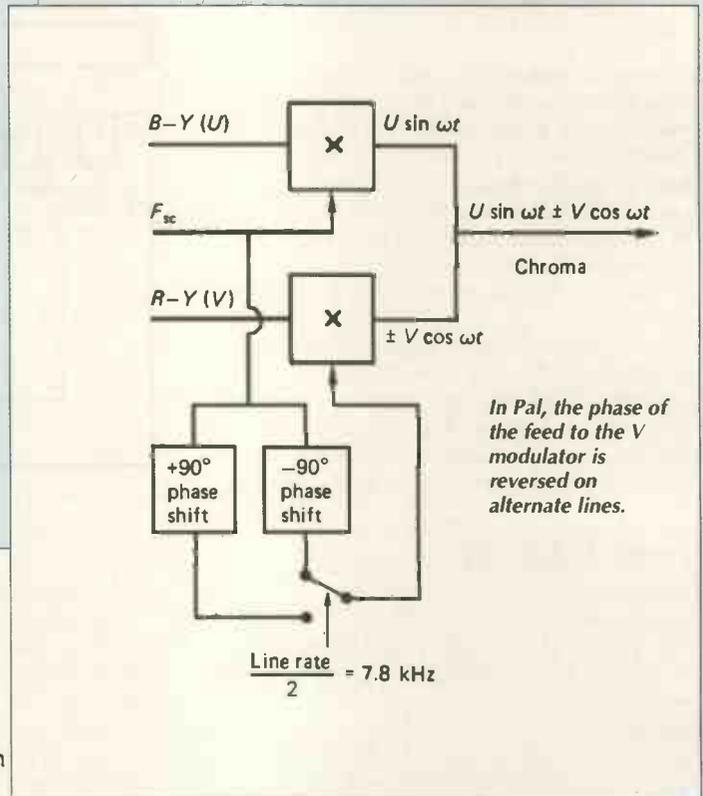
frame rate. A comb filter can be made in which the teeth are 25 Hz apart using two delays of one field each. This gives extremely precise Y/C separation, but this mode of operation is restricted to stationary or slow moving pictures, as subject movement will result in differences between successive fields which causes loss of resolution.



The PAL system

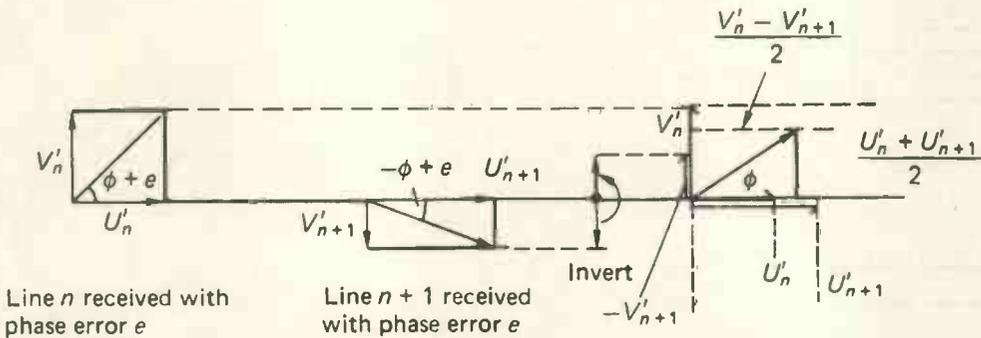
For colour television broadcast in a single channel, the pal and NTSC systems interleave into the spectrum of a monochrome signal a subcarrier which carries two colour difference signals of restricted bandwidth using quadrature modulation. The subcarrier is intended to be invisible on the screen of a monochrome television set. A subcarrier based colour waveform is generally referred to as composite video, and the modulated subcarrier is called chroma.

When the pal (phase alternating line) system was being developed, it was decided to achieve immunity to the received phase errors to which NTSC is susceptible. The figure right shows how this was achieved. The two colour difference signals ($[i]U[r]$ and $[j]V[r]$) are used to modulate a



Chroma phase ϕ on line n

Chroma phase $-\phi$ on line $n + 1$



Line n received with phase error e

Line $n + 1$ received with phase error e

Top, chroma as transmitted. Bottom, chroma received with phase error. V is reversed on second line and the two lines are averaged, eliminating phase error and giving a small saturation error.

Average of line n and line $n + 1$ removes error e restoring phase ϕ

posite signal which has been taken from the centre tap of the line comb to achieve time alignment.

Comb mesh failure can be overcome by switching back to the traditional low-pass luminance/high pass chroma separation technique until the comb meshes once more. The switching condition can be detected in two ways. The phase of chroma at the three filter taps can be compared to predict mesh failure. Alternatively the emergence of subcarrier frequency from the luminance output can be detected. The Motorola chip uses the former technique in the ACF (advanced comb filter)

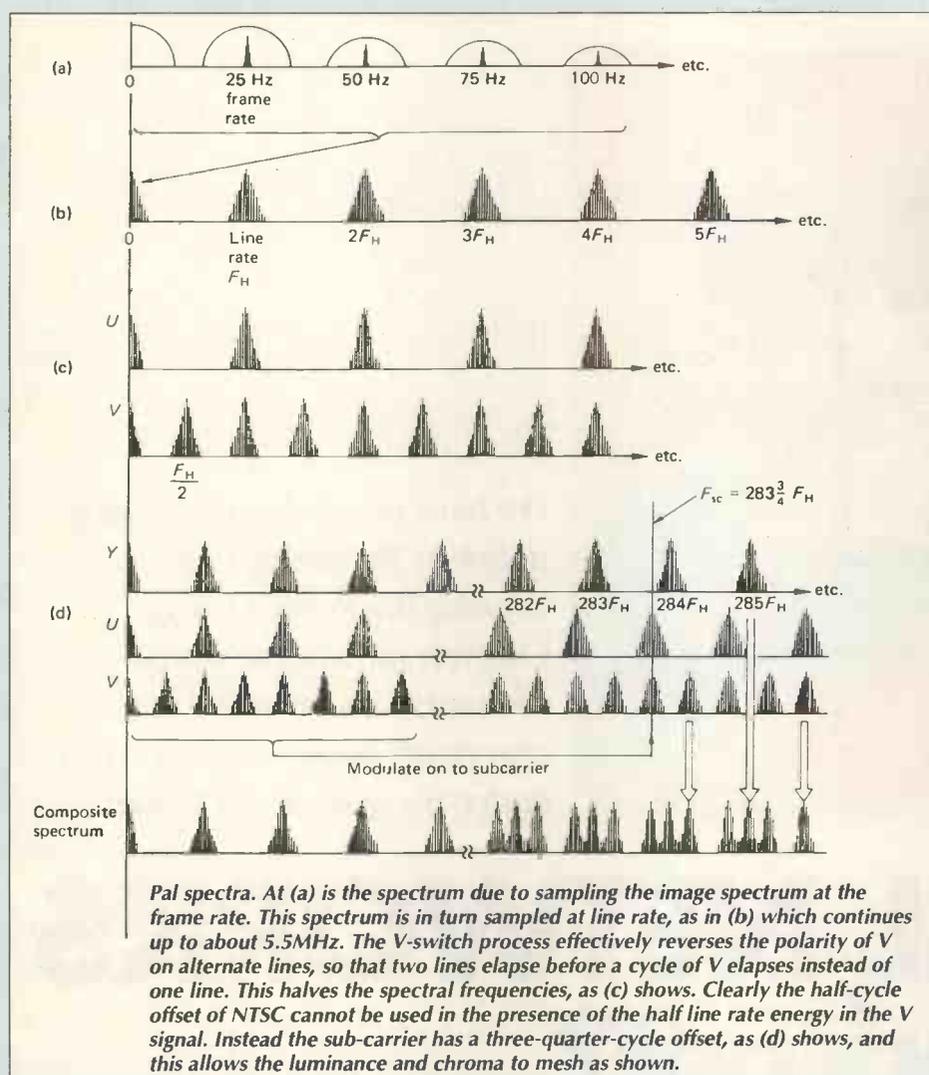
processing block. In addition an external input (pin 46) may be used to force the filter into bandpass mode.

This chip contains its own composite ADC and Y/C DACs and so can be used in an analogue application with little more external circuitry than a four-times-subcarrier ($4F_{sc}$) sampling rate clock. However, two parallel digital ports, C and D, are available allowing a variety of digital interfacing tasks to be combined with the Y/C filtering.

In digital output mode, the comb filtered composite analogue input is made available as digital luminance on port C and digital chroma

on port D. The chip also acts as two independent sections, a composite ADC with digital output on port C, and a comb filter/DAC with digital input on port D. There is also a mode which applies additional noise reduction filtering to the chroma output.

Figure 3 shows the block diagram of the ITT ACVP2205 comb filter chip. Like the Motorola device, this has a composite analogue input with an ADC running at $4F_{sc}$, and two one-line delays. The Y/C separator block contains a variety of low pass and bandpass filters and below this means to mix the filtered signals in various ways under the control of



subcarrier in quadrature in a similar way as for NTSC, except that the phase of the $[i]V[r]$ signal is reversed on alternate lines. The receiver must then re-invert the $[i]V[r]$ signal in sympathy. If a phase error occurs in transmission, it will cause the phase of $[i]V[r]$ to alternately lead and lag, as shown in the figure left. If the colour difference signals are averaged over two lines, the phase error is eliminated and the replaced

with a small saturation error which is subjectively much less visible. This does, however, have a fundamental effect on the spectrum.

Graph (a) above shows the spectrum due to sampling at a frame rate of 25/Hz. The line rate sampling of 15625/Hz then gives the spectrum of (b). The colour difference signals will have the same spectral structure, but less bandwidth. The $[i]B[r]-[i]Y[r]$ signal

becomes $[i]U[r]$, but the $[i]R[r]-[i]Y[r]$ signal will be inverted on alternate lines to become $[f]\pm[r][i]V[r]$. The inversion signal is a square wave at half line rate, and this effectively modulates the $[i]V[r]$ signal at half line rate so that the $[i]V[r]$ spectrum is divided by two, and so contains spectral entries at half line rate spacing. When $[i]U[r]$ and $[i]V[r]$ modulate the subcarrier, the spectrum of the chroma also contains energy at multiples of half line rate, and so spectral interleaving with a half cycle offset of subcarrier frequency will not work, as graph (c) shows. The solution is to adopt a subcarrier frequency with a quarter cycle per line offset. Multiplying the line rate by $283\frac{3}{4}$ allows the luminance and chrominance spectra to mesh as in graph (d).

The receiver needs to know whether or not to invert $[i]V[r]$ on a particular line, and this information is conveyed by swinging the phase of the burst plus and minus 135° on alternate lines. A damped PLL in the receiver will run at the average phase, i.e. subcarrier phase, but it will develop a half line rate phase error whose polarity determines the sense of the V-switch.

The quarter cycle offset means that there are now sets of four lines, and four frames or eight fields have to elapse before the same relationship of subcarrier to frame timing repeats. It also means that the line pair cancellation due to the half cycle offset of NTSC is absent, and another means has to be found to achieve visibility reduction. This is done by adding half frame rate to subcarrier frequency, such that an inversion in subcarrier is caused from one field to the next. Since in an interlaced system, lines one field apart are adjacent on the screen, cancellation is achieved. The penalty of this approach is that subcarrier phase creeps forward with respect to H-sync at one cycle per frame. The eight field sequence contains 2500 unique lines all having the subcarrier in a slightly different position. Observing burst on an H-triggered oscilloscope shows a stable envelope with a blurred interior.

the adaption circuit which predicts mesh failure.

In addition to Y/C separation, the ITT chip also has an on board chroma decoder which outputs the U and V colour difference signals directly. The S-VHS recording format can deliver separate Y and C signals and it is possible to bypass the comb filter so that these can be fed directly to the chroma decoder and the luminance processor. Luminance processing provides software controlled peaking filters which can be adjusted to compensate for aperture effect in the CRT and losses in converter filters. Additional a contrast enhance-

ment is available by making the luminance transfer function non-linear at low levels.

Figure 4 provides a comparison of adaptive comb filtering and conventional filtering. Note particularly the enhanced edge sharpness in the horizontal direction which results from the wider luminance bandwidth allowed by comb filtering. ■

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ACVP 2205 - ITT Semiconductors, Rosemount

House, Rosemount Avenue, West Byfleet, Surrey, KT14 6LB

I. Watkinson, J.R., *The Art of Digital Video*, Focal Press, Oxford, ISBN 0 240 51287 1, Chapter 5.

* John Watkinson is an independent consultant specialising in digital audio and video technology. He is a Fellow of the Audio Engineering Society and is the author of seven books



We have more chance of being struck by lightning than winning the Pools. George Overton shows how neural networks – so powerful for processing imprecise data – can nudge the odds in our favour.

NEURAL NETWORKS hit the jackpot?

Computer programs based on the structure of the brain can form the basis for artificial neural networks. Like the brain they are capable of learning to recognise patterns in data applied to their inputs, with their main advantage being an ability to solve problems where data is imprecise. Artificial neural networks are not restricted to pure logic and are able to simulate the fuzzy reasoning of the brain.

In the brain, neurons receive signals from other neurons through connections known as dendrites and output electrical signals through connections called axons. Axons themselves diverge into thousands of branches, each terminated by a synapse whose purpose is to inhibit, or excite activity in other neurons to which it is connected. Learning is achieved by adjusting these synapses, so changing the influence that one neuron has on others further down the network.

For a neural network, learning is an optimisation procedure where numerical values at each neuron are automatically adjusted in an iterative process to obtain the best performance. At the end of each iteration the output of the network is analysed and an index of performance (commonly called the network error) is evaluated. The weights associated with each node/neuron are then adjusted to reduce the network error until no further improvement can be achieved or a preset level is reached.

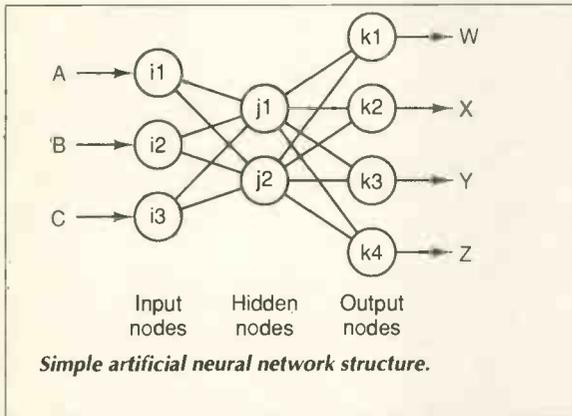
```
'Simple Example Of a Neural Network
GOSUB setupConstants
GOSUB setupNetwork
CLS
LOCATE 1, 40: PRINT "Start: "; TIMES
GOSUB trainingMode
LOCATE 1, 65: PRINT "Stop: "; TIMES
END

setupConstants:
  samples = 4
  idata = 2
  hidden = 4
  odata = 4
RETURN

setupNetwork:
  DIM sample(samples, idata) 'sample data for training
  DIM ipdata(idata) 'input data for network
  DIM iout(idata) 'output data from input units
  DIM wih(idata, hidden) 'weights between input and hidden units
  DIM hin(idata, hidden) 'input data to hidden units
  DIM hsum(hidden) 'sum of inputs to hidden units
  DIM hout(hidden) 'output data from hidden units
  DIM who(hidden, odata) 'weights between hidden and output units
  DIM oin(odata, hidden) 'input data to output units
  DIM osum(odata) 'sum of inputs to output units
  DIM oout(odata) 'output data from output units
  DIM dout(samples, odata) 'desired output
  DIM obeta(odata) 'error derivative for output units
  DIM hbeta(hidden) 'error derivative for hidden units
  DIM hthr(hidden) 'hidden threshold value
  DIM othr(odata) 'output threshold value
  DIM oerror(samples) 'network error
```

Typical structure

Simple artificial neural networks can be constructed using only three layers, the input, hidden and output layers. The network is free to construct its own internal representation of the input data within the hidden layer as it learns. Using two nodes for the hidden layer (as in the figure) is an



arbitrary choice and depends on the application. But a rule of thumb demands that the number of weights in the network must be much less than the number of patterns multiplied by the number of outputs. Other factors that must be considered are the complexity of the input data and the speed of response of the network. The relationship of "number of weights" to network convergence time is not linear.

Although not shown in the figure, each of the connections from a node passes through a modifiable weight (equivalent to the synapse in the brain) before the information is passed along to the next layer. During learning, the network automatically adjusts the weights to minimise the error present at the output nodes.

Training a Neural Network

An artificial neural network, needing to make predictions based on input data it has never seen, must first learn accurate recognition of patterns from a set of training data. This "supervised learning" aims to provide a set of examples from which the network can determine the rules.

A typical training sequence would be:

- Sample data continuously presented at the input to the network;
- generate an error value representing how close the actual output of the network comes to the desired output;
- update weights associated with each connection to reduce the error and bring the actual output closer to the desired output.

Networks presented with large amounts of data during learning will be able to recognise patterns never encountered before. Such systems represent artificial intelligence and will continue to improve as more data is added.

Adjusting the weights is the tricky part; in practice, the back-propagation algorithm is the most popular method and can be used¹ in such complex tasks as recognition of handwritten digits, predicting currency exchange rates and managing the yield of chemical processes. Artificial neural networks² are even being used in the fight against crime.

Error minimisation

During learning, a neural network, in effect, performs an optimisation in *n*-dimensional variable space, finally

```

RANDOMIZE TIMER
'Assign random values to weights between input and hidden layer
FOR i = 1 TO idata
  FOR j = 1 TO hidden
    wih(i, j) = RND * 2 - 1
  NEXT j
NEXT i
'Assign random values to threshold levels
FOR i = 1 TO hidden
  hthr(i) = RND * 2 - 1
NEXT i
FOR i = 1 TO odata
  othr(i) = RND * 2 - 1
NEXT i
delta = .5 'set step size for weight adjustment
tolerance = .1 'set error tolerance for network
'Assign random values to weights between hidden and output layers
FOR i = 1 TO hidden
  FOR j = 1 TO odata
    who(i, j) = RND * 2 - 1
  NEXT j
NEXT i
fl1$ = "example.dat" 'sample data file for training
fl2$ = "example.wht" 'calculated weights from training
noConvergence = 1
RETURN

trainingMode:
GOSUB readSampleData
GOSUB displayTitles
pass = 0
WHILE noConvergence
  FOR scan = 1 TO samples
    LOCATE 1, 1: PRINT "Pass = "; pass; TAB(19); "SAMPLE = "; scan;
    " "
    GOSUB calculateOutputs
    GOSUB calculateErrorForOutputLayer
    GOSUB calculateErrorForHiddenLayer
    GOSUB adjustWeights
    GOSUB calculateNetworkError
  NEXT scan
  pass = pass + 1
  GOSUB checkForConvergence
WEND
LOCATE 22, 45: PRINT "Network has converged"
GOSUB storeCalculatedWeights
RETURN

readSampleData:
OPEN fl1$ FOR INPUT AS #1
FOR i = 1 TO samples
  INPUT #1, sample$
  FOR j = 1 TO idata
    sample(i, j) = VAL(MID$(sample$, j, 1))
  NEXT j
  INPUT #1, sample$
  FOR j = 1 TO odata
    dout(i, j) = VAL(MID$(sample$, j, 1))
  NEXT j
NEXT i
CLOSE #1
RETURN

calculateOutputs:
'Assign input data
FOR i = 1 TO idata
  ipdata(i) = sample(scan, i)
NEXT i
'Calculate output of input units
FOR i = 1 TO idata
  iout(i) = ipdata(i)
  LOCATE 4 + i, 20: PRINT iout(i)
NEXT i
'Calculate input to hidden units
FOR i = 1 TO idata
  FOR j = 1 TO hidden
    hin(i, j) = iout(i) * wih(i, j)
  NEXT j
NEXT i
'Calculate output of hidden units
FOR j = 1 TO hidden
  hsum(j) = 0
  FOR k = 1 TO idata
    hsum(j) = hsum(j) + hin(k, j)
  NEXT k
  hsum(j) = hsum(j) + hthr(j)
  hout(j) = 1 / (1 + EXP(-hsum(j)))

```

converging to an optimum point at which the hyper-surface yields the lowest network error. But convergence to the global minimum is not guaranteed. In fact one of the criticisms of the back-propagation algorithm method (see box) is that it is possible for it to converge onto a local minimum which may not represent the most optimum value.

Not all input variables may be of equal importance. Network structure is designed to take account of this weighting by providing factors for each node (neuron). The numerical value at the input to a node is multiplied by the associated weight before being combined with other weighted inputs to produce the output function.

Two main methods may be employed to evaluate the error measured at the output of the network. The first involves combining errors from each output node by taking the sum of the moduli:

$$E = \sum_{i=1}^n |e_i|$$

The procedure simply involves taking the sum (Σ) of all the errors (e_i^2) at the output nodes (from $i = 1$ to n) and ignoring the sign of the error during the summation process. E represents the index of performance, or network error.

Drawback to the approach is that it does not distinguish between cases where there is one large error together with a number of smaller errors. In reality the most practical method for combining the errors at the output nodes is to take the sum of the squares of the errors:

$$E = \sum_{i=1}^n e_i^2$$

The error is obtained by taking the sum (Σ) of the squared errors (e_i^2) at the output nodes (from $i = 1$ to n). Generating the network error in this way has the advantage that larger errors are given more weight than smaller ones, concentrating minimisation on reducing the largest errors first.

Binary sample

The process required to produce an artificial neural network program is not as difficult as it appears from the

```

LOCATE 4 + j, 31: PRINT INT(hout(j) * 1000) / 1000; "
NEXT j
'Calculate input of output units
FOR j = 1 TO odata
  FOR k = 1 TO hidden
    oin(j, k) = hout(k) * who(k, j)
  NEXT k
NEXT j
'Calculate output of output units
FOR j = 1 TO odata
  osum(j) = 0
  FOR k = 1 TO hidden
    osum(j) = osum(j) + oin(j, k)
  NEXT k
  osum(j) = osum(j) + othr(j)
  oout(j) = 1 / (1 + EXP(-osum(j)))
  LOCATE 4 + j, 47
  PRINT INT(oout(j) * 1000) / 1000; "
NEXT j
RETURN

calculateErrorForOutputLayer:
FOR i = 1 TO odata
  obeta(i) = oout(i) - dout(scan, i)
NEXT i
RETURN

calculateErrorForHiddenLayer:
FOR j = 1 TO hidden
  hbeta(j) = 0
  FOR k = 1 TO odata
    hbeta(j) = hbeta(j) + who(j, k) * oout(k) * (1 - oout(k)) * obeta(k)
  NEXT k
NEXT j
RETURN

adjustWeights:
FOR i = 1 TO idata
  FOR j = 1 TO hidden
    FOR k = 1 TO odata
      wih(i, j) = wih(i, j) - (delta * iout(i) * hout(j) * (1 - hout(j)) * hbeta(j))
    NEXT k
  NEXT j
NEXT i
FOR i = 1 TO hidden
  hthr(i) = hthr(i) - (delta * hout(i) * (1 - hout(i)) * hbeta(i))
NEXT i
FOR i = 1 TO hidden
  FOR j = 1 TO odata
    who(i, j) = who(i, j) - (delta * hout(i) * oout(j) * (1 - oout(j)) * obeta(j))
  NEXT j
NEXT i
FOR i = 1 TO odata
  othr(i) = othr(i) - (delta * oout(i) * (1 - oout(i)) * obeta(i))
NEXT i
RETURN
    
```

Football goal

Millions of people gamble on the football pools, but how many realise the enormous odds stacked against them in favour of the pools companies? In fact, the odds against an eight-draw pool from ten selections filled in solely by chance are a staggering 42,595,495 to 1. These are very long odds indeed⁴ – the chance of being struck by lightning is a meagre 600,000 to 1.

My goal has been to create an artificial neural network to provide a more accurate prediction of the weekly football results. For anyone interested in lowering the odds, here are some observations:

- A draw in any form (whether it is of the jackpot, high-score or no-score variety) is

extremely difficult to predict. It only requires one goal to transform a dead cert draw into a result. Accuracy of predictions is increased by focusing on the home and away output nodes of the neural network.

- Data available for each match on the coupon is extremely fuzzy, and in many cases is just not available. My best recommendation is to use the form figures supplied by some of the Sunday papers, in conjunction with the teams' division, as a source of input data.

- Do not use full perms. Number of selections is governed by cost of the perm, and full perms are expensive. In addition, no matter how good a neural network is at predicting the results, it will never be 100% correct, and many outsiders still end up as draws. Casting a wider net by using a non-

consecutive permutation will increase chances of winning⁵.

- Although my winnings have not been spectacular, I have won a small amount using this method – a great improvement on my past performance of zero wins

- Basic structure required for construction of an artificial neural network, designed for football pools prediction, is contained in the example program. All that is required is a second option to allow the input of data by the user, and the ability to read in the recorded weights determined during training. Using the knowledge it has learned by examining past results, the network is then able to make predictions concerning new data (applied to the input nodes) which it has not previously encountered.

Good luck!

mathematics, and is best explored by a simple example.

A network is configured to learn how to convert binary coded numbers, applied to its inputs, into single outputs: as the input counts up in binary from 0 to 3, the network learns to represent each of the possible combinations by a high signal at one of the output nodes.

The program (EXAMPLE.BAS) is written in Qbasic on the PC, for simplicity, and has been designed to be easily changed for other configurations. The flexibility is to allow readers to explore different combinations of input to output relationships with relative ease, and to determine the effect of varying the number of hidden nodes.

Subroutine SETUPCONSTANTS allows the following values to be set, prior to running the program:

- samples = number of training examples to expect.
- idata = number of input nodes
- hidden = number of hidden nodes
- odata = number of output nodes

Two other files are required by the program:

- EXAMPLE.DAT - contains the training data.
- EXAMPLE.WHT - saved weight values for use in the calculation mode.

Training data simply consists of consecutive lines of input node levels and desired output levels. For the binary to single output conversion example, the training data contained in the file EXAMPLE.DAT is:

00, 0001, 01, 0010, 10, 0100, 11, 1000

Convergence occurs when the network error is reduced to 0.1 at which point the weights are saved into the file EXAMPLE.WHT.

Information in the brain is transmitted along the axons as short electrical impulses known as action potentials.

```

calculateNetworkError:
  oerror(scan) = 0
  FOR i = 1 TO odata
    oerror(scan) = oerror(scan) + (oout(i) - dout(scan, i)) ^ 2
  NEXT i
  oerror(scan) = oerror(scan) / 2
RETURN

checkForConvergence:
  neterror = 0
  FOR k = 1 TO samples
    neterror = neterror + oerror(k)
  NEXT k
  LOCATE 22, 1
  PRINT "Delta = "; delta
  PRINT "Network error = "; neterror; "
  IF neterror < tolerance THEN noConvergence = 0 ELSE noConvergence = 1
  IF RIGHT$(STR$(pass), 1) = "00" THEN GOSUB storeCalculatedWeights
RETURN

storeCalculatedWeights:
  OPEN fl2$ FOR OUTPUT AS #1
  PRINT #1, delta
  FOR i = 1 TO idata
    FOR j = 1 TO hidden
      PRINT #1, wih(i, j)
    NEXT j
  NEXT i
  FOR i = 1 TO hidden
    PRINT #1, hthr(i)
  NEXT i
  FOR j = 1 TO hidden
    FOR k = 1 TO odata
      PRINT #1, who(j, k)
    NEXT k
  NEXT j
  FOR i = 1 TO odata
    PRINT #1, othr(i)
  NEXT i
  CLOSE #1
  LOCATE 23, 46: PRINT "Saving PASS "; pass; " to disk"RETURN

displayTitles:
  LOCATE 3, 18: PRINT "inputs   hidden nodes   output nodes"
  LOCATE 4, 18: PRINT "____   _____"
RETURN
    
```

Back-propagation algorithm

Many algorithms have been applied to artificial neural networks with varying success. The back-propagation algorithm is one of the simplest to understand and, though it is unlikely to be the process employed by biologically based neural networks, it does provide good results despite its tendency to converge on a locally optimum solution instead of finding the global convergence point.

As a network learns from the sample data applied to its inputs, it gradually reduces the overall error for each pass until a set limit is reached or the error cannot be reduced further.

A convenient way to determine the network error is to calculate the sum of the squared errors of the output nodes:

$$E_n = 1/2 \sum (y_k - d_k)^2$$

where E_n is the calculated error for the network (ie how close is the network to convergence?). y_k is the actual output and d_k is the desired output.

Subscripts i, j, k refer to the input, hidden and output nodes respectively. So, a

performance index is derived by first finding the errors at each output node ($y_k - d_k$), squaring the result and adding these values to provide an overall error measurement for the network. The value can then be used to determine the adjustment required for each weight.

Neurons are commonly compared to switches, but a unit step function (ie a sharp transition between 0 and 1) is rarely used. Instead, a function known as the sigmoid function, with well defined rise and fall times, is employed to represent the neuron's behaviour.

$$y = \frac{1}{1 + e^{-x}}$$

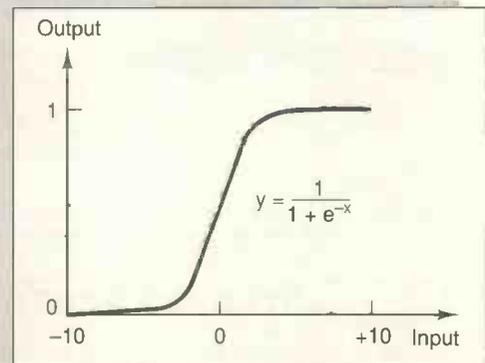
shown diagrammatically in the figure.

First step to reducing the error in the network is to find the localised gradient produced by individual weights. Starting with the weights located between the hidden and output nodes, the derivative of the network error with respect to these weights is calculated:

$$\frac{\delta E_n}{\delta W_{jk}} = y_j y_k (1 - y_k) \beta_k$$

where $\beta_k = (y_k - d_k)$ is the error at the output of an output node (ie difference between actual and desired output, and $y_k(1 - y_k)$ is the derivative of the sigmoid function.

Using this method and moving backwards through the network from the output to the input, we can determine the effect that each weight has on the overall network error. Contribution to the network error of the weights located between the input and hidden nodes can be found from calculating the derivative as follows:



The electrical impulses are generated by neurons when their inputs reach a critical threshold level, but to adjust both the weights and the trigger levels associated with each neuron within a computer simulation is a difficult task. One solution³ (called biasing) changes the threshold into an additional weight connected to the node (neuron) which is driven by a node that always fires. The extra weight has the value of $-T$, where T is the threshold level. Biasing has been implemented in the example program.

An obvious improvement would be to input data from the keyboard during calculation, generating an output from the network dependent on the saved weight values. But the example has purposely been kept as small and simple as possible for illustration.

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5. "How To Win The Pools By Really Trying", G B Stone, 1982.

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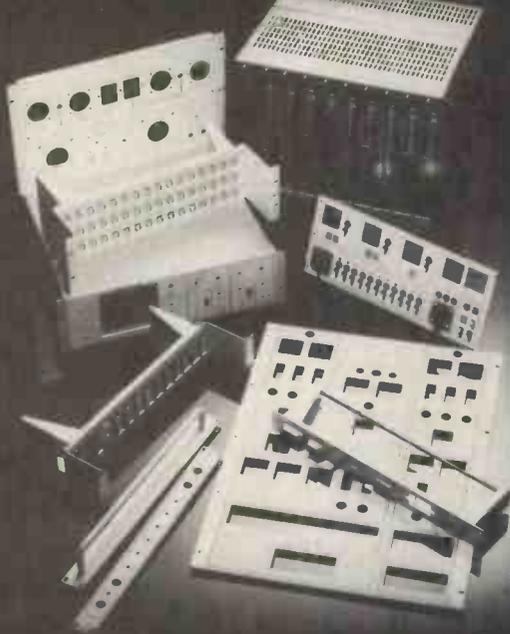
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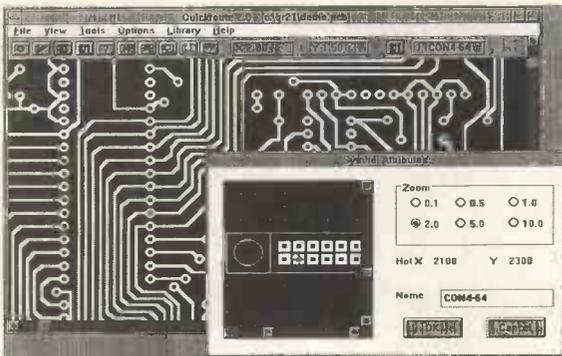
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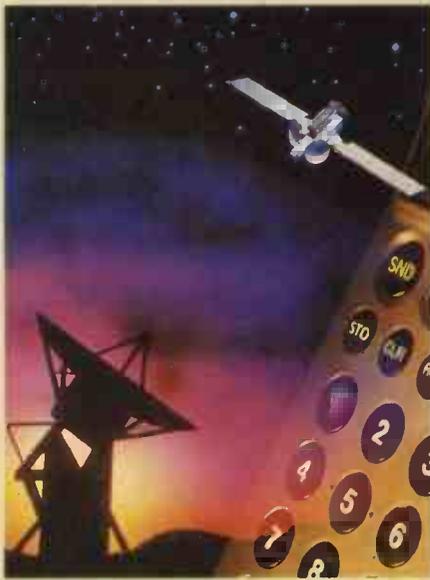
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UHF technology for the cordless revolution

2: systems on a chip



Ian White reports on how higher levels of UHF IC integration are drastically reducing the number of external components required for a complete RF system.

The transmitter and receiver form only one part of a cordless communication system, and their objective is to be as transparent as possible – the user wants all the freedom of cordless operation but none of the limitations. While the overall performance of the receiver may be limited by carrier/interference ratio or other considerations essentially related to system capacity, the receiver's basic RF performance must not intrude. This requires adequate sensitivity to weak signals, combined with freedom from strong-signal overload effects, particularly third-order intermodulation in systems which use strong signals on regularly-spaced channels. The front-end power gain must be well-defined, and may possibly also need to be controllable. All of this must be accomplished within strict limits on power consumption from a single +3V or +5V DC rail, with further power-down facility for standby.

Two converging trends are currently influencing UHF integrated circuits. Lower-cost GaAs ICs, which until recently were regarded as a specialised microwave technology, are now appearing in the RF and first mixer stages of UHF mass-market products. At the same time the complex silicon ICs which already include all the lower-frequency functions are seeking to absorb the rest of the receiver as well. The frontier between these partitions is constantly shifting, leading to product diversity.

Single-function MMICs (microwave monolithic ICs) have been available for some years. For example the wideband 50Ω RF amplifiers and mixers pioneered by Avantek are well established in a wide variety of applications including receiver front ends and low-power transmitters and exciters. However, consumer cordless communications offer a unique combination of closely-defined needs and a huge potential market, encouraging the development and mass-production of UHF ICs with a higher level of integration.

Complete RF systems with the minimum of external components is the goal: for example the higher level of integration allows the inclusion of on-chip buffer amplifiers with single-ended 50Ω inputs and outputs.

Integrated receiver front-ends

Typical examples of integrated front-ends are the Motorola *MRFIC* devices mentioned last

month, and the National Semiconductors *LMX2216B* (Fig. 1). This bipolar/cmos device includes a low-noise RF amplifier (LNA) and mixer in bipolar technology, together with cmos DC and supervisory circuits. The device is specified for use between 100MHz and 2.0GHz and makes provision for inserting a signal-frequency filter between the LNA and the mixer. All ports are single-ended; the LNA input/output and mixer input and local oscillator ports are all matched to the standard 50Ω impedance, which also suits the available signal-frequency filters, while the mixer has a 200Ω output impedance which is more suitable for IF filters in the 100MHz region. As a result the only other external RF components required are DC blocking capacitors. Since the LNA and mixer are internally compensated for broadband performance and the signal and IF filters will probably be pre-aligned, the customary tune-up of RF systems can be reduced or even eliminated. In a typical transceiver system, including all necessary filtering and transmit/receive switching ahead of the LNA, the *LMX2216B* delivers a noise figure of typically 9.5dB combined with a very adequate dynamic range for strong signals – and all from a +3V, 6mA supply with power-down to 10μA on standby.

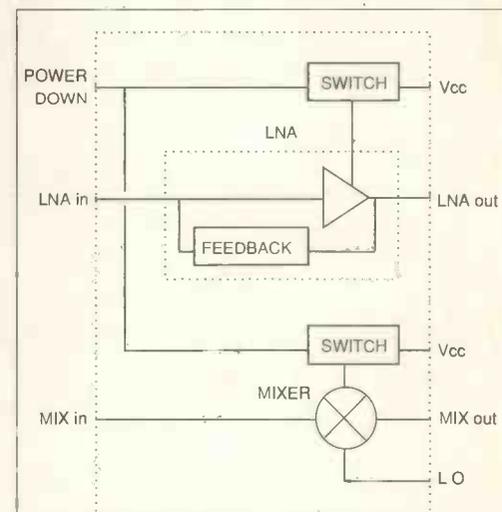


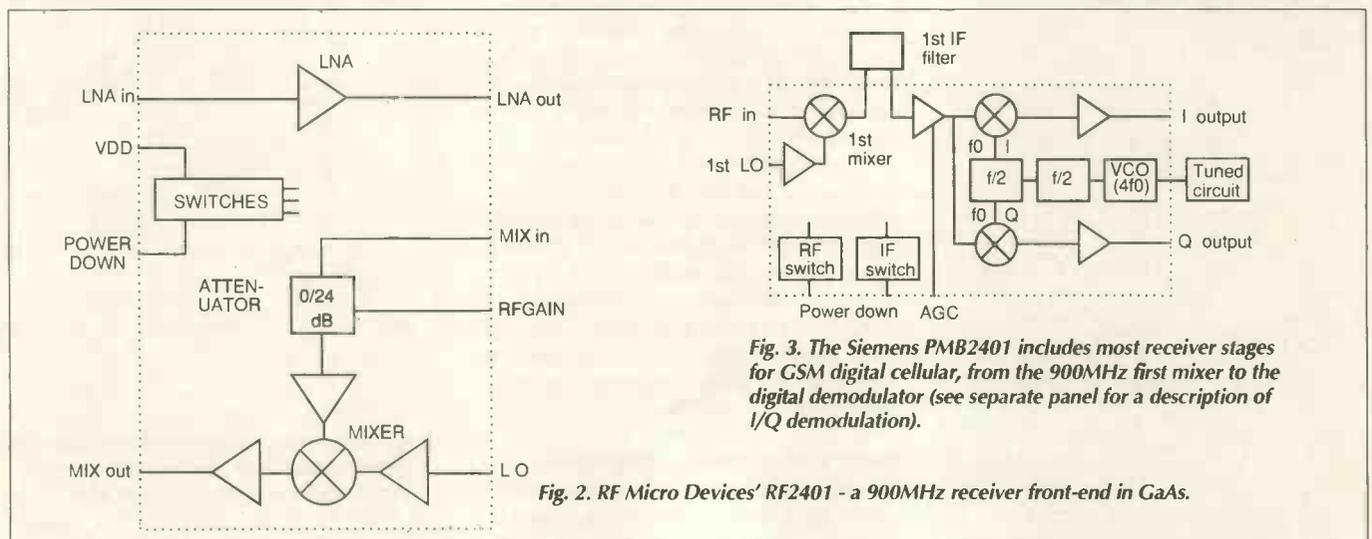
Fig. 1. The National Semiconductor *LMX2216B* includes a low noise amplifier and mixer for many uses between 100MHz and 2GHz.

	GSM	DCS-1800	NADC	JDC	Qualcomm CDMA	CT-2
	Groupe Speciale Mobile	PCN Digital cellular system	North American digital cellular	Japanese digital cellular	Code division multiple access	cordless telephony
Geography	Europe	Europe	North America	Japan	U.S.	UK(Telepoint)
Service	1991	1992-1993	1991-1992	1991-1993	1992-1993	1989
Frequency Range	935-960MHz 890-915MHz	1.7-1.9GHz	824-849MHz 869-894MHz	810-826MHz 940-956MHz 1429-1441MHz 1447-1489MHz 1501-1513MHz	824-849MHz 869-894MHz	864-868MHz
Data Structure	TDMA	TDMA	TDMA	TDMA	CDMA	TDMA
Channels/frequency	8-16	8-16	3-6	3-6	118	1
Modulation	GMSK	GMSK	P1/4 DQPSK	QPSK	Two-level GFSK OQPSK	PI/4 DQPSK $\pm 14.4/25.2$ kHz deviation
Speech CODEC	REL P-LTP 13kbits/s	REL P-LTP 13kbits/s	VSELP 8kbits/s	VSELP 8kbits/s	9550 bps 32kbits/s	ADPCM 32kbits/s
Mobile Output Power	3.7 mW to 20W	250mW to 2W	2.2mW to 6W		0.2 to 6.3 W	1 mW to 10mW
Modulation Data Rate	270.833kbits ± 15.5 Hz	270.833kbits	48.6kbits	42kbits	1.2288 Mblts/s	72kbits/s
Bit duration	3.692 μ s					
Frequency accuracy	50ppm					
Mobile carrier	13.5 Hz					
Filter	0.3 Gaussian	0.3 Gaussian	cosine, r=0.35	cosine, r=0.35		0.5 Gaussian
Channel Spacing	200kHz	200kHz	30kHz	25kHz	1.23MHz	100kHz
Number of Channels	124	3000-6000	832	1600	10	40
Frame length	4.615 ms					
Time slots	8	8	6	6		2(TDMA) 40(FDMA)
Time slot length	0.677 ms		6.66 ms			
Burst length	156 1/4 bits		324 bits			
Power range	20 mW - 20 W					
Estimated subscribers by 2000	<20 million					

Gallium arsenide technology is capable of minimal noise levels at frequencies into the microwave region. Discrete fets and single-stage MMICs are commonplace, but until recently the high development costs of GaAs ICs have limited the scope for larger-scale integration. Many earlier GaAs devices were also difficult to use, requiring dual-rail DC supplies which are out of the question for

cordless equipment. Once again the mass market has provided the impetus for new developments. One of the first manufacturers to offer complete GaAs front ends has been RF Micro Devices, with devices such as the *RF2401* (Fig. 2). This chip provides essentially the same functions as the National Semiconductor and Motorola devices, with the addition of a switchable 24dB attenuator for

use when signals are strong. Placed between the LNA and the mixer, the attenuator protects the vulnerable mixer from strong signals both on the wanted channel and elsewhere within the passband of the signal-frequency filter. In a similar 900MHz front end to that used for the *LMX2216B*, the GaAs device would give a lower noise figure of about 6.5dB and rather better dynamic range, though this is achieved



CT-3	DECT Digital European cordless telephony	JCT Japan digital cordless phone	TETRA Trans European trunked radio	TFTS Terrestrial flight telephone system	ERMES European radio message system	POGSAC (Cityruf) pager system
cordless telephony Europe	Europe	Japan	Philips- Cambridge 1994?			Europe
1990	1992-1993	1991-1992				
800-1000MHz depending on country 1800MHz future	1.88-1.9GHz	1.9GHz	450MHz	1693 to 1694MHz 1625 to 1626MHz	169.4 to 169.8MHz	450 to 570MHz
TDMA	TDMA 124-84	TDMA	TDMA			
P1/4 DQPSK	P1/4 DQPSK ±259kHz ±317kHz deviation	4FSK	FSK. FFSK		±1.6625kHz ±4.6875kHz	±4.0kHz
ADPCM 32kbits/s	ADPCM 32kbits/s 250 mW	ADPCM 10 mW				
640kbits/s	1.152 Mbits/s 361µs	384kbits/s	36kbits/s	44kbits/s	6.25kbits/s	1200 bits/s 512 bits/s
0.5 Gaussian 1MHz 8	0.5 Gaussian 1.726MHz 132 9934 bit 10ms 24 0.417 ms 424 bits 250mW	cosine,r=0.5 300kHz 2x12	cosine,r=0.4	cosine r=0.4. 4	Bessel(10) 3.9kHz .25kHz	
		6.5-13 million				

Source: Rohde & Schwarz

at the expense of higher power consumption, +5V at 30mA.

The 6-10dB noise figures of the *LMX2216B* and *RF2401* front ends are typical of many other products, and comfortably meet the system specifications for receiver sensitivity. For example, the sensitivity specification for a 1.8GHz DECT receiver represents a noise figure of 13-14dB; this is mainly because the communication system is designed to operate in areas of high environmental RF noise, which makes internal receiver noise much less of an issue.

As far as cordless telecomms are concerned, it is clear that historical problems of achieving a low-enough noise figure at UHF have been vanquished.

Higher integration

Alongside these developments in UHF technology are new ranges of ICs for signal processing, control and logic functions. These devices are complex and their low power, high-density silicon processing will operate at VHF. They are capable of taking much of the RF subsystem into the main chip-set, leaving little for specialised UHF circuitry. It is normal to find the second mixer and final IF integrated with baseband processing functions,

and recent offerings are absorbing the first mixer too.

The Siemens *PMB2401* is an example of a single IC containing an almost complete receiver for the 900MHz GSM digital cellular system (Fig. 3). The IC includes all the active stages in the signal path from the UHF first mixer to the digital baseband outputs; the only extra components required at 900MHz are the transmit/receive switch, LNA and filters. As with all such ICs, an external VCO is required for the first frequency conversion, though the *PMB2401* does include the second VCO and digital phase shifting to generate the I/Q digital output signals required for further baseband processing (see separate panel).

Siemens has adopted a different interface philosophy from some other manufacturers, using push-pull inputs and outputs wherever possible. The 5V *PMB2401* has two separate control lines to allow power-down from a maximum current consumption of 20-30mA to a standby value of about 5mA. Other ICs in the same family are available with analogue rather than digital outputs.

The new digital cordless telephone system proposed for the USA and Canada is designed to run in parallel with the existing analogue system. The IS-54 communication standard

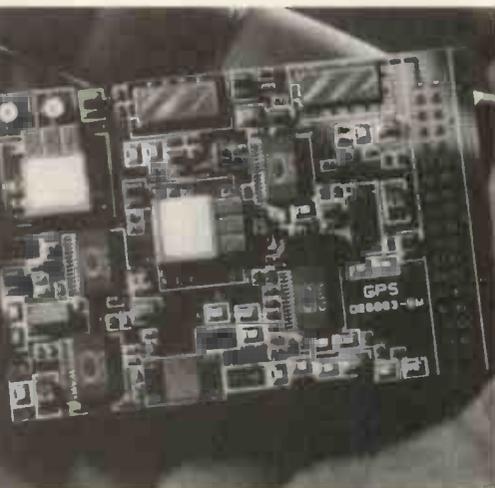
requires new silicon to possess both digital and analogue capability. Accordingly AT&T have announced a dual-mode chip-set which includes the *W2005* receiver.

Beginning with the first mixer, the device provides similar functions to the *PMB2401* plus a conventional analogue output. The entire chip-set operates from a +5V supply; the *W2005* requires 17mA reducing to less than 100µA on standby.

Transmitters

RF output power requirements range from typically 10mW for CT2 to a few hundred milliwatts for a compact hand-held unit operating from a 3V or 5V supply. However, many systems such as GSM and DECT also require the mobile transmitter to reduce its power in defined steps if the signal strength at full power is more than adequate. Design considerations fall into two areas, the exciter and the power amplifier, and for various reasons it is better to discuss the power amplifier first.

RF power relates directly to battery consumption, and should not be wasted. However, this maxim encounters considerable difficulties when the transmitter is connected to a real-life antenna. In many cases the antenna is a helically-shortened monopole operating



Hand on the future. Complete wireless LAN system on a PCMCIA card. Devices such as this DE6003 radio modem G-P Semiconductor include all the elements to turn a standard laptop computer into a cordless mainframe terminal.

against an ill-defined capacitance between the radio and the hand of its user rather than a proper groundplane. Shortened antennas have narrower bandwidths than their full-sized counterparts so, even in the best of circumstances, there are difficulties in maintaining an impedance match across the entire bandwidth of a multi-channel system. But shortened antennas are also sensitive to detuning by their surroundings – again including the user. As a result the power transfer from the transmitter to the antenna will be highly variable and inef-

ficient, and output stages must be capable of absorbing considerable amounts of reflected power at all conceivable phase angles. Sometimes a circulator is used to protect the transmitter from the reflected power, but this is an undesirable complication in low-cost handheld units.

Since battery power is at a premium, the facility for stepped power reduction is never implemented at the output of the transmitter. Instead the input signal to the final amplifier is reduced in the exciter stages, where power levels are lower and signal attenuation wastes less power in absolute terms.

One approach is to monitor the power output and use a power-levelling control loop to maintain the correct value. This is particularly suitable for the class-C power amplifiers normally used for FM transmission, because it compensates for the non-linearity in the input/output power relationship. However, the newer generation of UHF power amplifier ICs use a different mode of operation: these devices possess a linear transfer characteristic such that a reduction of, say, 12dB in input power will result in a 12dB attenuation in the output signal.

Since power amplifier ICs have typically 20-30dB power gain, the requirement for the exciter is to provide a band-filtered signal at a level of about 1mW, which can also be attenuated on demand.

Synthesisers and exciters

The exciter in all UHF cordless transmitters is based on a frequency synthesiser, which phase-locks to a stable reference frequency.

The actual multiplication factor is defined by a series of frequency dividers whose division ratio is controlled by the channel-select logic, and the synthesiser must generally shift channels when changing between transmit and receive.

To generate a frequency-modulated or phase-modulated signal, the normal method is to modulate the VCO directly with the baseband signal (Fig. 4). This is only possible if the bandwidth of the synthesiser's feedback loop is significantly less than the modulating frequency (a wideband feedback loop would treat the modulation as an error signal and remove it). However, there are problems if fast frequency hopping is required, because a small loop bandwidth will also result in a long settling time before the VCO output recovers its phase-coherence after a channel change.

National Semiconductor has tackled these problems head-on in its transmitter proposals for the Dect system. A synthesiser for Dect must have a settling time of 400µs maximum and the necessary loop bandwidth could interfere with the modulation process. The NS system opens the phase-locked loop for the 400µs duration of the transmitted data burst. The LMX2320 synthesiser and LMX2410 baseband processor together provide the VCO with a buffered control voltage which remains at its last-corrected value for the duration of the transmitted pulse. Even at 1.8GHz, the frequency drift of the unstabilised VCO in this brief period can be orders of magnitude better than required by the Dect standards, and the synthesiser loop bandwidth places no restrictions on the modulation bandwidth.

QPSK AND RECEPTION

Quadrature phase-shift keying is a highly accurate method of applying a phase shift to a carrier. The same principle can also be used in a receiver to extract the baseband phase shift information from the incoming phase-modulated signal. QPSK is very similar to the phasing method of generating a single-sideband voice transmission in that it requires in-phase and quadrature (Q – ie shifted in phase by 90°) signals at both the modulating and the carrier frequency.

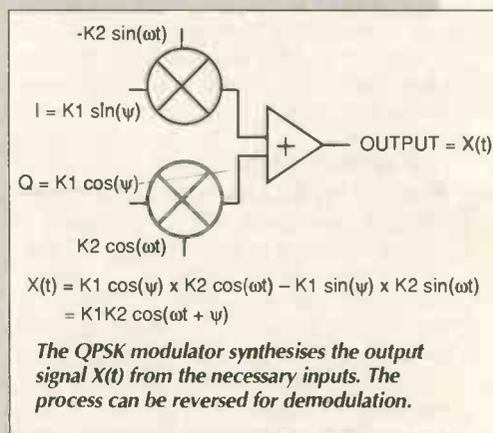
As shown in the diagram, the method of applying a phase shift ψ to a carrier of frequency radian/sec is to generate the phase angle $(\omega t + \psi)$ using the well-known trigonometrical identity involving the sines and cosines of the angles ψ and (ωt) . The formulas are shown in the diagram. A QPSK modulator is simply a small analogue computer which synthesises the necessary functions: multiplication is achieved by balanced modulators, negation by a 180° phase shift ψ , and the resulting signals are summed to produce the phase-shifted output signal.

The I and Q modulating signals can

easily be generated in a digital modulation system, using rom lookup tables if necessary for the sines and cosines and a digital-to-analogue converter. For UHF applications, a variety of ICs are available to accomplish the RF phase-shifting and summing of the modulated signals. An accurate RF phase shift of 90° is generated by starting from a signal at twice or four times the required output frequency; this may be generated either externally in the frequency synthesiser or by an on-chip frequency multiplier as used in the Siemens PMB2200 QPSK modulator (Fig. 6). A frequency doubler followed by a pair of 2:1 frequency dividers, triggered by positive- and negative-going edges respectively, will produce the required 90° phase shift at the output frequency. The RF and modulation signals are then applied in I and Q pairs to two identical balanced modulators and the summed output is a single-sideband QPSK signal.

The equations relating the output phase shift to the I/Q modulating signals are completely reversible, and thus apply equally well to the

demodulation of QPSK signals. Applying a phase-modulated signal to the port shown as OUTPUT in the diagram will generate I and Q output signals proportional to $\sin \psi$ and $\cos \psi$ respectively. This demodulation method is used at the first IF of 35-100MHz in the Siemens PMB2401 receiver IC (Fig. 3). The equivalent of further bandpass filtering is achieved by lowpass filters at the I and Q data outputs.



A further complication with direct modulation of a VCO which operates at the final output frequency is that when the power amplifier is switched on, a certain amount of RF power leaks everywhere around the transceiver and causes frequency-pulling of the VCO. While frequency-pulling can be minimised by careful screening and filtering, IC manufacturers are persuading designers towards an up-converting exciter architecture which places the VCO on an entirely different frequency and thus eliminates the problem. For example, Motorola's *MRFIC1803* GaAs up-converter for 1.8GHz Dect applications (Fig. 5) includes a buffered balanced mixer and a variable output attenuator with a 20dB control range.

Digital communication systems can use a different method of modulation from conventional frequency-shifting: quadrature phase-shift keying or QPSK (see panel). The necessary frequency dividers and matched pairs of balanced modulators are ideal candidates for an IC, and several manufacturers offer quadrature modulators for frequencies up to at least 1GHz. For example, the Siemens *PMB2200* GSM modulator (Fig. 6) includes an on-chip frequency doubler to 1.8GHz, the two frequency dividers to generate 900MHz I/Q signals, the matched pair of balanced modulators and output summing network, and a buffer amplifier. In common with the remainder of the Siemens GSM chip-set, the *PMB2200* uses differential input and output signals both internally and externally.

On the horizon

The next step ought to be the placing of all components within a single package. This is already beginning to happen. Miniaturisation of the GEC Plessey *DE6003* 2.4GHz data transceiver featured in last month's article will not stop at the present size of 75mm x 50mm. Current plans are to encapsulate almost the entire radio system into a single hybrid package based on a silicon wafer which contains the external resistors and capacitors in monolithic form. The remaining functions will be implemented in the same four ICs as at present, but mounted on the wafer as flip-chips leaving enough room inside the package for low-profile SAW filters. This in turn will create space to bring the spread-spectrum communications protocol controller on to the main board. The result will be a complete 2.4GHz wireless LAN adapter inside an industry-standard PCMCIA card which plugs straight into a notepad PC.

Looking further ahead, one can forecast a divergence in the level and availability of RF technology for UHF to SHF. High-performance ultra-miniature packaged radio sub-systems will appear on the market, operating on higher frequencies where there are more channels available, and prices will fall as production technology improves. But these products will only be available for the few large-volume applications which can repay the development costs.

Although cordless RF technology will become more accessible to electronic design-

ers catering for the mass markets, it needs a competent RF design team to provide ready-made radio modules which a non-specialist designer can use with confidence. These skills will become concentrated into fewer hands accustomed to designing at the wafer-scale

level rather than with PC boards. That specialisation of skills may have ominous consequences for small-scale innovation, which is one of RF technology's major strengths.

At least, it has been up to now.

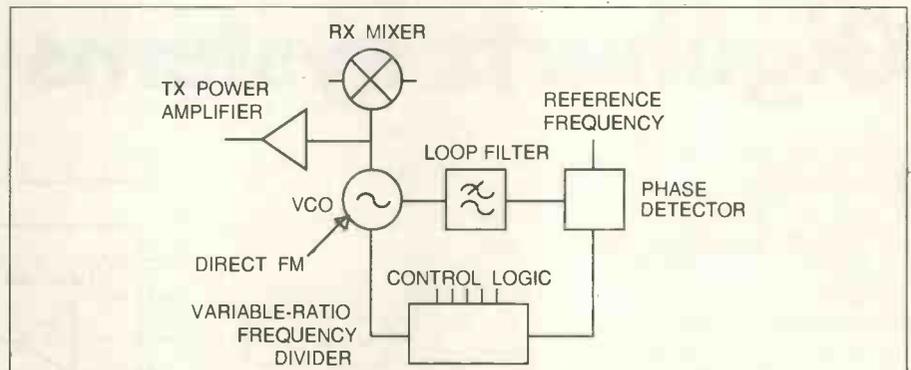


Fig. 4. Direct modulation of the voltage-controlled oscillator at the final output frequency may not be as simple as it seems.

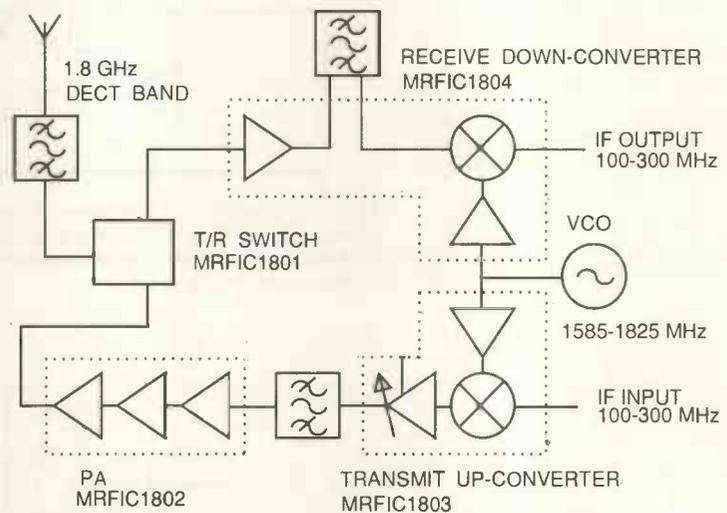


Fig. 5. Motorola's line-up for a 1.8GHz Dect system uses GaAs ICs throughout, with an up-converting transmitter. Note that the *MRFIC1803* also controls the power output from the *MRFIC1802* linear amplifier.

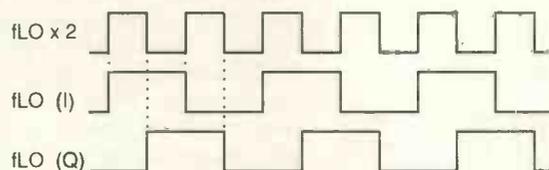
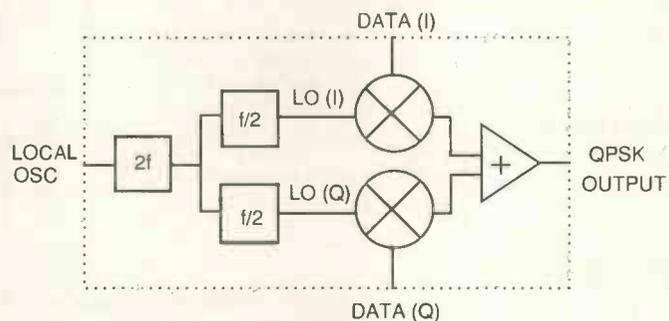


Fig. 6. The Siemens *PMB2200* QPSK modulator for GSM uses a frequency-doubler and two 2:1 dividers to generate 90° phase shifts at 900MHz.

Martin Eccles examines the application details of the new National Semiconductor UHF chip set.

Gigahertz systems on a chip

Intended for use with the European digital (dect) standard, a new silicon chip set provides most of the operating functions for low-power communications transceivers operating up to 2GHz.

Using a National Semiconductor proprietary bimos process these ICs represent the first silicon-based transceiver chip set to operate at over 1GHz. The ICs are also relatively cheap to manufacture. All are capable of operating from a single 3V supply. Compared with existing transceiver designs, the chip set is said to offer the potential for reducing component count by up to 66%.

In common with other bimos technologies, the process used lends itself to mixed-signal analogue and digital circuitry. As a result, functions such as RF amplification and digital signal encoding can potentially exist on the same chip. Of the four new devices, one is an amplifier/mixer, one is an IF receiver, one is a frequency synthesizer and one is a baseband processor.

Amplifier/mixer operates to 2GHz

Down conversion and flat gain over a 0.1 to 2GHz range are provided by the LMX2216 low-noise amplifier mixer, Fig. 1. With a 3V supply, it needs typically 6.2mA for the amplifier section and 8.9mA for the mixer. In total this equates to just over 45mW. A doubly-balanced Gilbert-cell mixer provides local-oscillator to RF isolation of 30dB while cancelling second-order distortion products. There are separate power supply pins for the amplifier and mixer. As a result, consumption can be minimised if only one section is needed.

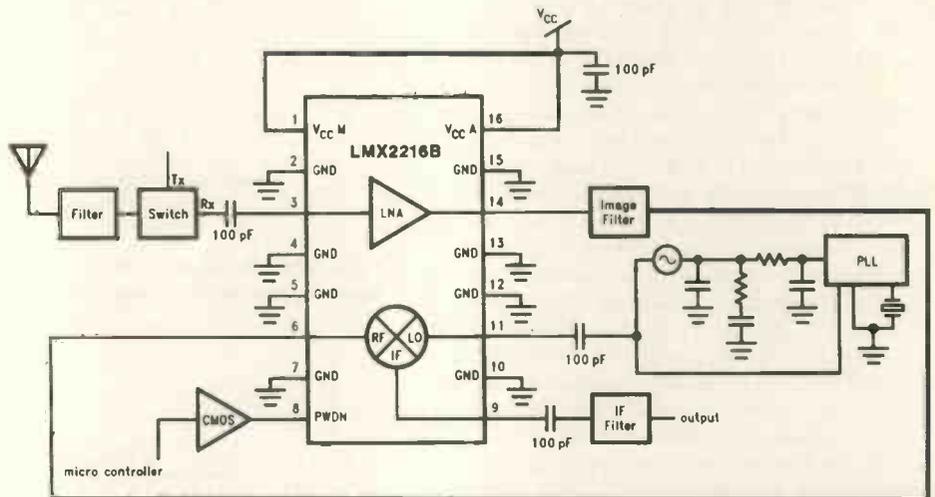


Fig. 1. In this radio-receiver front end, a new 2GHz low-noise amplifier/mixer IC is fed by a PLL local oscillator but in other applications a free-running oscillator could well suffice. Intermediate-frequency output would typically be filtered by a channel-select filter ready for further down conversion or demodulation.

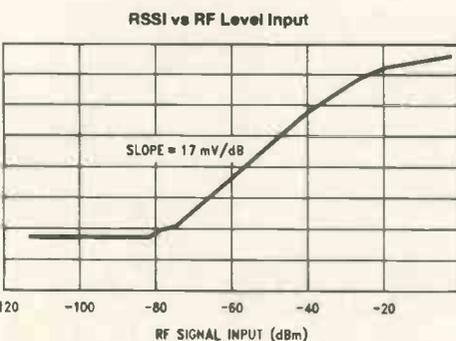


Fig. 3. Received signal strength indication built into the LMX2216 IF receiver. RSSI is important in hand-held communications equipment. When input falls below a useful level, the system can be powered down automatically to maximise battery life. The indication could also be used to select the channel with the strongest signal.

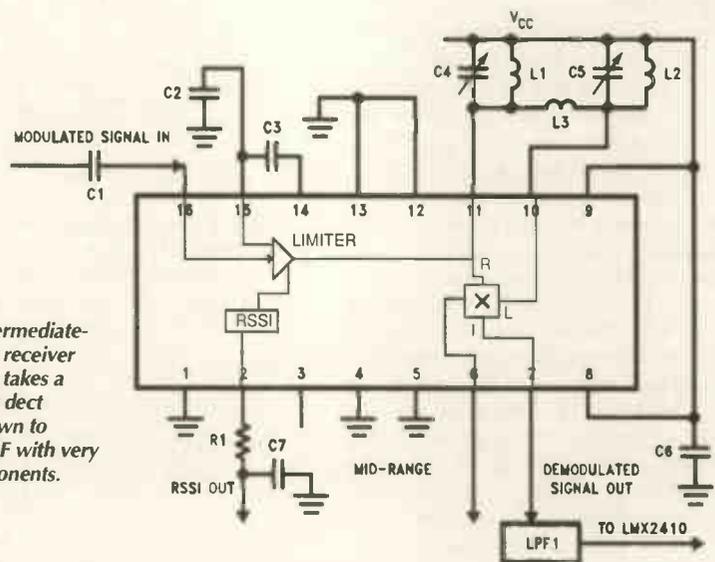


Fig. 2. Intermediate-frequency receiver and mixer takes a 1890MHz dect carrier down to 110MHz IF with very few components.

Further, power-down circuitry reduces current drain to less than 10µA.

Input biasing is on chip; the LMX2216 needs few external components apart from blocking capacitors. Only eight pins are needed for all amplifier and mixer i/o signals, power supplies and power-down control. A further eight strategically positioned ground pins enhance operation at higher frequencies by reducing lead-impedance effects.

All i/o apart from the 200Ω mixer output port is single ended and matched to 50Ω. Relative to discrete circuits, the device needs

little or no tuning due to its monolithic construction.

Single IC processes 110MHz IF

Currently, most monolithic IF demodulators rely on multiple IF stages and demodulate up to 455kHz. The NS LMX2240 is a five-stage single-chip solution capable of operating at typically 110MHz with an RF sensitivity of around -72dBm, Fig. 2.

Besides combining mixer, amplifier, filter and oscillator functions, the 2240 further reduces component count since it incorporates

signal-strength indicator circuitry. Gain of the limiter amplifier is 70dB. Limiter input, at 150Ω, is matched to typical surface-acoustic-wave filters operating at 110MHz. Limiter output impedance is 100Ω for driving an external quadrature tank circuit and the IC's internal discriminator.

For communications equipment relying on cellular operation, good signal-strength indication is important. Sensitive indication circuitry allows an intelligent receiver to reduce power consumption immediately when signal strength falls below an intelligible level. Operating over a 70dB range, the signal-strength indicator circuit of the 2240 provides a logarithmic output from 0.3 to 1.8V. Sensitivity of the signal-strength section is -82dBm, Fig. 3.

Tx/Rx PLL for dect telephones

When combined with a single voltage-controlled oscillator circuit, the LM3220 phase-locked loop frequency synthesizer caters for both transmit and receive requirements of a DECT radio, Fig. 4.

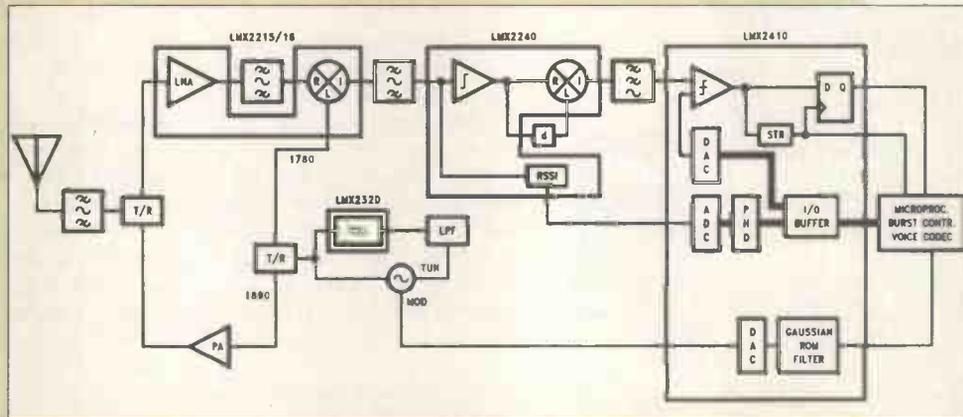
This digital PLL is claimed to be the lowest power frequency synthesizer IC on the market capable of operating up to 2GHz. It also incorporates a proprietary phase comparator and internally-balanced charge pump for fast channel switching. While operating from a 3V supply, the LM3220 has a current consumption of only 12mA. In sleep mode, consumption falls to 500μA. Prescaling is selectable for either 64/65 or 128/129 modulus division via a simple three-wire control interface involving serial data, clock and enable signals.

To accommodate frequency scanning and narrow-band operation, the device has an on-board analogue switch. The purpose of this switch is to decrease the loop-filter time con-

Dect or CT2?

Compared with the current CT2 standard, the European cordless telecomms dect standard is claimed to offer a nearly seventeen-fold increase in communications traffic. Data shows that the European dect standard is capable of handling 10,000 erlangs/km² while CT2 only caters for 600 erlangs/km².

At 1.9GHz, the operating frequency of dect is only just over double that of CT2 but bandwidth is increased five times. Unlike CT2, dect accommodates simultaneous voice/data encryption and seamless handover. It also caters for ISDN, GSM and X25.



Combining small size, low power consumption and high integration, these four new ICs from National Semiconductor could be the first step towards bringing the cost of a dect handset below \$30 by 1995. Only one PLL is needed for both transmit and receive functions.

stant, allowing the VCO to adjust to its new frequency more quickly. This is achieved by paralleling another filter stage.

Output of the charge pump normally passes via D₀ but when LE is high, charge-pump output also becomes available at BISW. Filter

stage LPF2 is effective only while the switch is closed (in scanning mode).

Baseband processing

The last building block in the series is the LM2410 baseband processor. This device provides an interface between the RF section of a dect radio system and the digital portion. It handles both transmit and receive functions. Digital pulse shaping for either direct VCO or quadrature modulation is provided in the transmitter section. On the receiver side, digital-to-analogue conversion is carried out and clocking information recovered. There is also an a-to-d converter with peak-hold detector to allow the received signal strength to be read by a microcontroller.

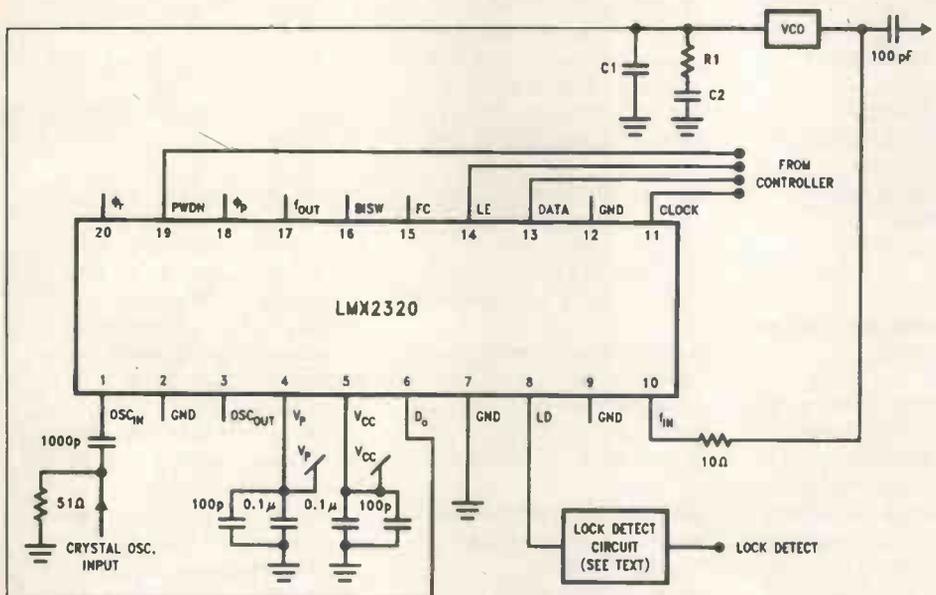


Fig. 4. One digital PLL for 1.9GHz dect equipment handles both transmit and receive functions, reducing component count. Its three-wire programmable prescaler can be set for 64/65 or 128/129 division.

Typical Loop Filter Component Values

Ref. Freq.	C1	R1	C2
300 kHz	390 pF	10 kΩ	0.1 μF
1.728 MHz	330 pF	10 kΩ	0.022 μF

Note: (K_V = 11 MHz/V)

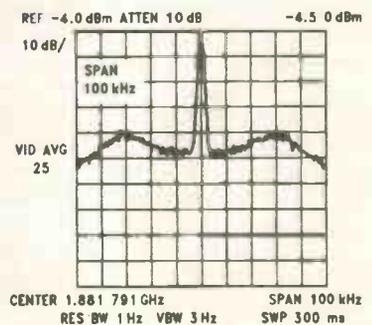
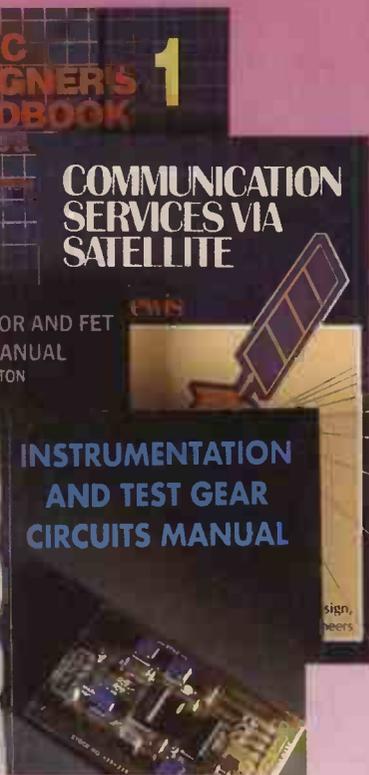


Fig. 6. Performance of the LM3220 PLL. This is a typical phase-noise plot with an offset of 10kHz, resulting in -77dBc/Hz. Reference frequency is 1.728MHz.

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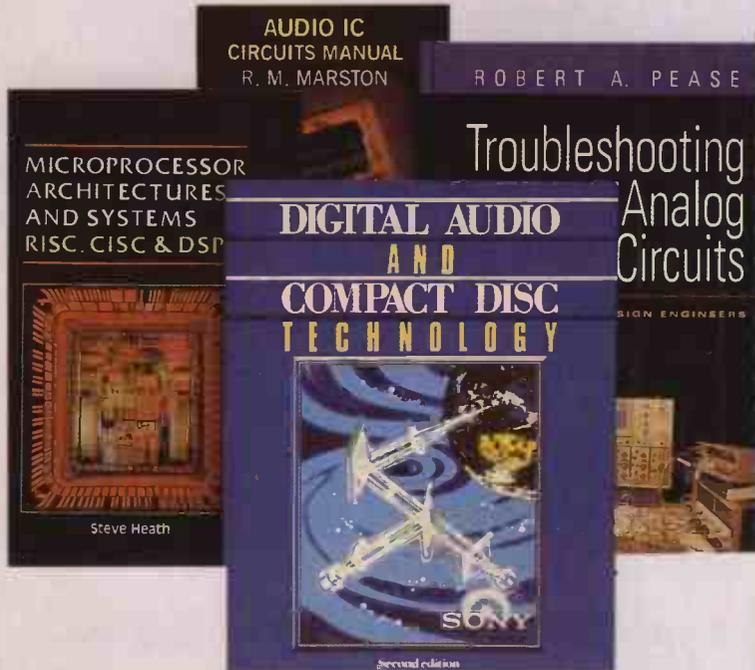
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Marconi's magnetic domain that stretches into the ether

The influential Marconi magnetic detector relied on the movement of magnetic domain walls – years before such phenomena were understood. T H O'Dell builds a real detector and shows why Marconi's co-scientists were so confused.

Magnetic detector manufactured in quantity by the Marconi Company from 1902 and widely used for about 20 years.

Before the advent of valves, Marconi's magnetic detector was *the* radio receiver. In its time, it excited much speculation about how exactly it worked, and since then many theories have been propounded. The only way to find the true mechanism seems to be to give a real device a thorough testing. But the search for a solution must first begin by going back to Marconi's patent for the device: British Patent No. 10245, filed on May 3, 1902. Marconi begins by stating that the:

"invention is based upon the discovery that a ...magnetic material which is not sensibly affected by high frequency oscillations ...under ordinary conditions becomes sensitive to them when placed in a ... moving magnetic field".

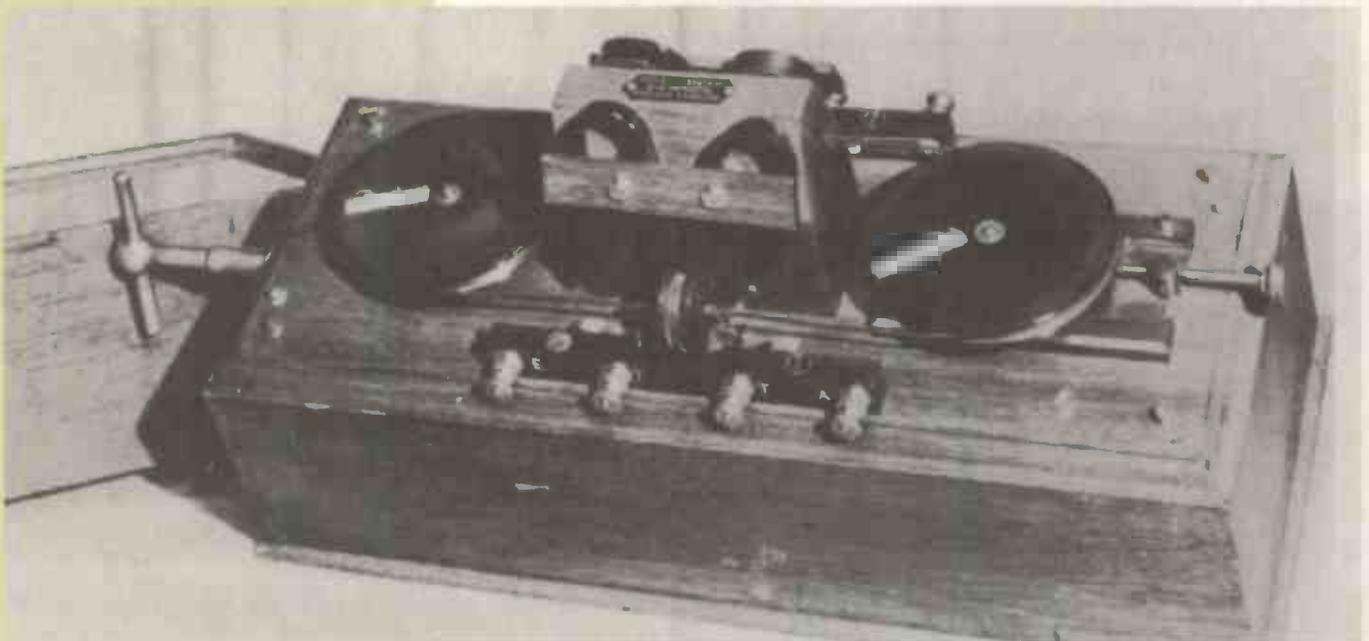
His meaning becomes clear when we look at the diagram in his patent specification which gives the details of the device. It shows a loop (a) of "hard drawn iron wires", made to move endlessly through two coils by a clockwork drive. Inner coil (b), a single layer winding, is connected between aerial and earth through a tuning filter. Outer coil (c), with a large number of turns of fine wire, is connected directly to the headphones.

Operated in this way, the magnetic detector was a sensitive receiver for CW signals transmitted by damped spark transmitters working in the long wavelength band that was used for marine wireless telegraphy at the beginning of the century.

The important observation to make about the magnetic detector is the pair of permanent magnets arranged around the two coils. These provide the "moving magnetic field" that Marconi stipulated in his patent specification, and he chose to make the magnetic material move through the field. Note that Marconi also clearly specified that the two magnets (d) had "like poles together". Later authors⁴ often became confused about this point and placed the magnets in very unlikely positions!

The two magnets placed according to Marconi instructions produce a quadrupole field that causes the iron rope to divide into two magnetic domains magnetised tail to tail. The effect happens because the rope is under tension and iron is a material with positive magnetostriction. So tension causes the rope to have an easy direction of magnetisation along its axis.

When the rope is made to move to the right, the magnetic domain wall separating the two domains also moves to the right until it finds itself in a magnetic field strong enough to



cause it to move continuously backwards at the same speed as the rope is moving forwards. This is its equilibrium position. Relative to the coils around the rope, the domain wall is stationary and no signal will be induced in either coil.

But suppose a short pulse of current is made to flow in the input coil (b). If the pulse polarity causes a magnetic field along the axis of the iron rope that adds to the polarity of the quadrupole field around the moving domain wall, then the backward velocity of the wall is increased and it flicks back. An input pulse of the opposite polarity will not have much effect because the quadrupole field is increasing in intensity as we move rightwards and the input pulse would simply cause the wall to move slightly to the right.

Sensitivity to the polarity of input pulse explains the detector's action. When the input to the magnetic detector is a damped sine wave, of the type sent out by an early twentieth century spark transmitter, the detector responds to the first half cycle of the damped

wave train with the correct polarity. The domain wall flicks backwards, and a large signal should be induced in the many-turn coil (c) to produce a click in the headphones. The click will be repeated at whatever repetition rate the spark transmitter may have, reproducing the dots and dashes of morse code, just like a crystal detector.

But unlike the crystal the magnetic detector is, in principle, an amplifier. Its power supply, the clockwork mechanical input power driving the iron wire rope, enables it to produce more electrical output power from the coil driving the headphones than the radio frequency input power supplied to the input coil. In contrast, a crystal detector introduces a considerable loss between RF input power and audio output power.

New construction

As a theoretical explanation of how this archaic device works, all this speculation may be of some help. But far more satisfactory would be some experimental evidence.

Place in history

Writing recently in *EW + WW*¹, Stanley Wood pointed out that Marconi's magnetic detector, the "Maggie", was one of the most important kinds of radio receiver of its time: from 1902, when it was introduced by the Marconi Company, to around 1918, when valves began to become widely used. Keith Geddes has made the same point in his Science Museum Booklet *Guglielmo Marconi: 1874-1937*, published to commemorate Marconi's centenary²: the magnetic detector "was in widespread use for nearly twenty years". The device was certainly manufactured in quantity as the one on display in the Science Museum in South Kensington bears the serial number 725.

The Marconigraph, the journal which became *Wireless World* and is now *Electronics World + Wireless World*, tells us that the famous ship Titanic had a Marconi magnetic detector as its only receiver³.

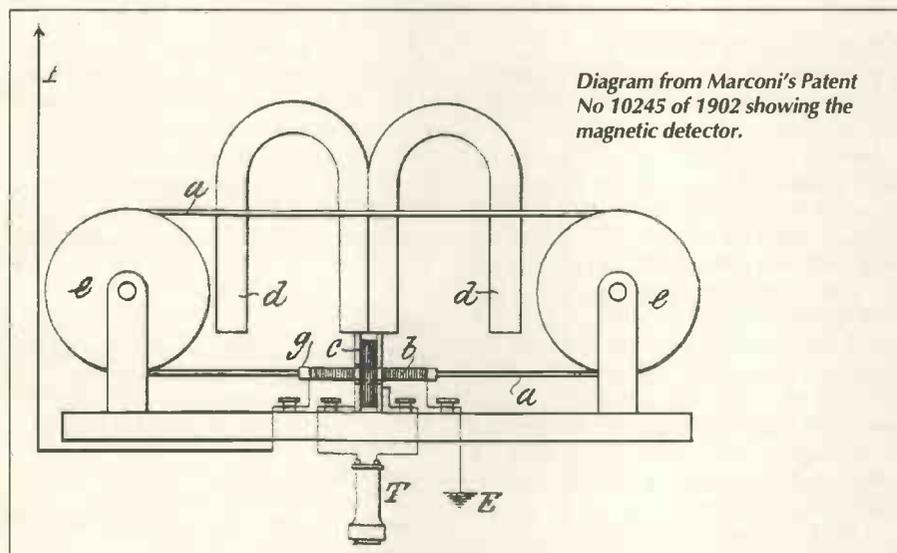
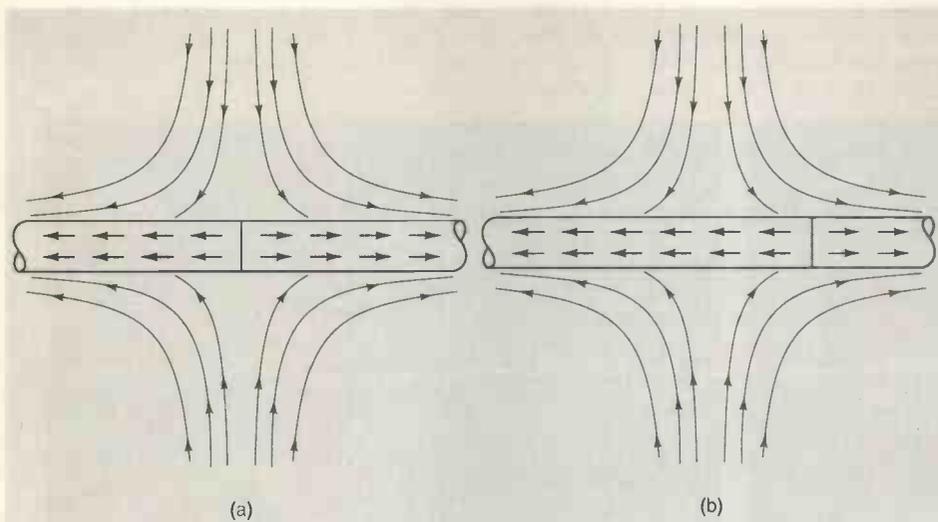


Diagram from Marconi's Patent No 10245 of 1902 showing the magnetic detector.



Magnetic field produced by the permanent magnets in the detector causes the iron wire rope to split into two magnetic domains. When the rope is stationary the situation is as shown in (a). When the rope moves to the right the domain wall moves to the right also (b).

Surviving magnetic detectors in museums are not useable for experimentation because of their very fragile state – mainly through corrosion in the iron wires making up the rope cores. The alternative is to construct a new magnetic detector.

The wire rope needed is made from seven strands of 200 μ m piano wire, put in tension around the two pulley wheels and made to move at about 10cm/s by a simple motor drive. The two coils are set around the moving rope.

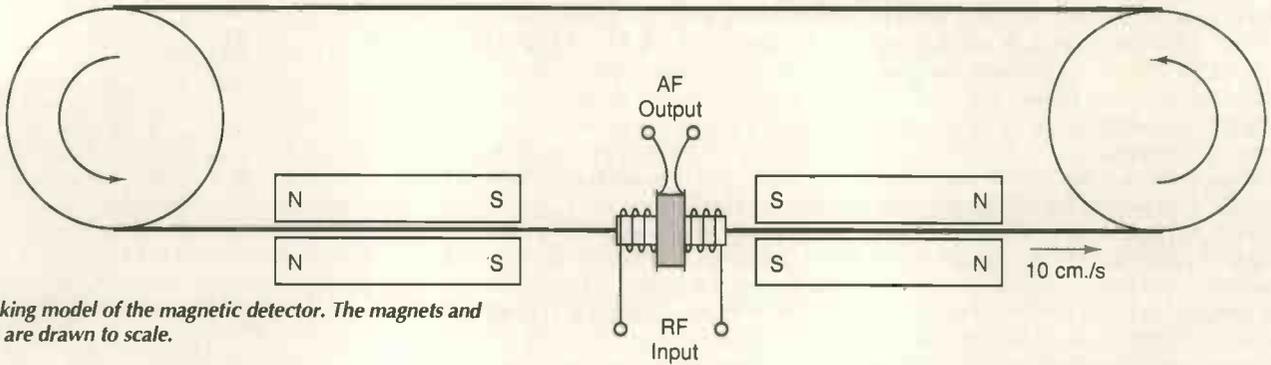
The "RF input" coil is a layer of 50T on a ceramic tube 2.5mm OD and 10mm long. A single layer of Mylar film is put over the coil. On top of the Mylar goes a 200T winding of 40 μ m insulated copper wire over a 2mm length at the centre of the first coil, to serve as the "AF Output" coil. The quadrupole magnetic field is provided by clamping four reed operating magnets (RS 349-052) in place.

Experimental results

First experimental test of the model was made with no connection to the "RF input" coil, and with the "AF output" coil connected to the high impedance input of an oscilloscope of 10mV/div sensitivity. The position of the quadrupole magnets was then adjusted to give maximum noise output.

As would be expected from the theoretical explanation given above, a sharp maximum in the noise output is obtained when the "AF output" coil is moved away from the centre of the magnet system so that it is a few millimetres off-centre in the direction of motion. The "AF output" coil is now centred over the domain wall as it moves backwards through the moving magnetic core and the noise output is due to fluctuations in its velocity arising from unavoidable inhomogeneities in the wire.

The second experimental test was made by connecting the "RF input" coil to a pulse gen-



Working model of the magnetic detector. The magnets and coils are drawn to scale.

erator and applying a 10µs pulse of either positive or negative polarity. Such an input pulse should simulate the first half cycle of a 50kHz damped spark transmitter signal, the kind of input for which the Marconi magnetic detector was intended.

Where the direction of the current in the "RF input" coil is such that it produces a magnetic field that adds to the quadrupole field around the moving domain wall, driving it backwards against the direction of motion of the wire rope, a large induced signal can be observed in the "AF Output" coil during the time that the domain wall is moving inside the "AF Output" coil. Amplitude of the output pulse begins to fall after the first 5µs.

Increasing the input pulse length to 20µs confirms that the domain wall is driven out of the "AF output" coil by that time.

Using a double pulse as an input shows that the magnetic detector takes a few milliseconds to recover once the input pulse is over because the domain wall can only return to its equilibrium position at the 10cm/s speed that the wire rope is made to move. Increasing the speed of the wire rope reduces the time needed for recovery.

Applying an input current pulse that drives

Results of this second experiment. In both cases the lower trace shows the input current to the "RF input" coil, on a scale of 10mA/div, while the upper trace shows the output from the "AF output" coil, on a scale of 20mV/div. Time base is 5µs/div. The detector only responds to a positive input pulse as shown (a). A negative input (b) has negligible effect.

the domain wall further in the direction of motion results, during the input pulse, in a clear decrease in noise from the "AF output" coil. The effect is expected because the wall is moved only a very small distance in the direction of motion by this sign of input pulse. So there is negligible output signal though it will be held more firmly because the gradient of the quadrupole field will be greater in this new position, explaining why there is a drop in noise level.

Science lagging technology

Two conclusions to be drawn from this investigation of the magnetic detector that could have implications for us today.

Marconi was making use of effects in magnetic materials that depend on the motion of magnetic domain walls. Yet the existence of domains was not known in 1902. It was not until 1907⁵ that Pierre Weiss first put forward the idea of magnetic domains; 1919 when Barkhausen found experimental evidence for their existence, and not until the 1930s that the first clear experimental work on moving magnetic domain walls was published by Sixtus and Tonks. Marconi's important step forward in technology must have been made in advance of the scientific establishment: science did not lead technology in this case – it was left far behind!

A second conclusion looks to be that the scientific establishment can deny the opportunity to advance its understanding. Scientific papers published in an attempt to explain the operation of Marconi's magnetic detector at

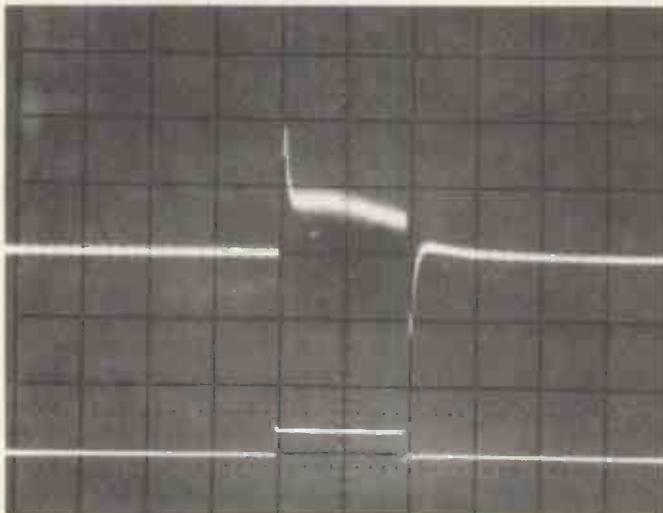
the time, were either reductionist or simply nonsense.

For example, Professor Ernest Rutherford, then at McGill University in Canada, wrote that the magnetic detector was: "well known to physicists". Professor J A Fleming, of University College London, in a paper he read to the Royal Society, said that the RF input to the detector "probably" brought about "a momentary release of the molecules of iron from the constraint (or viscosity) in which they are ordinarily held".

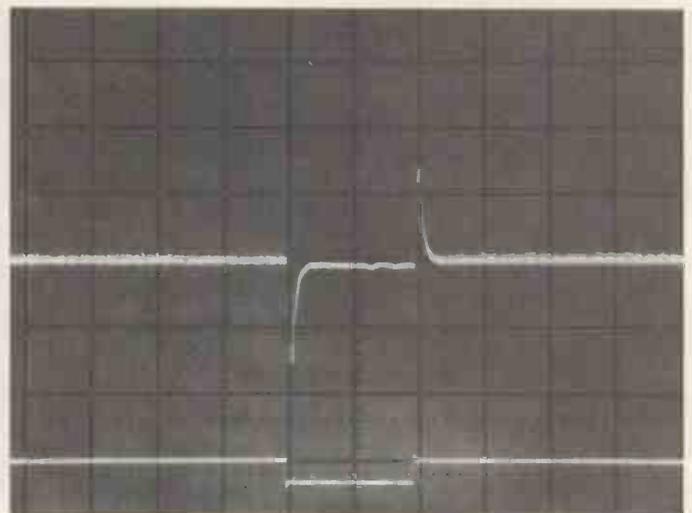
In contrast, Marconi himself was almost certainly responsible for a very sensible account of how his detector worked, which was written for the British Admiralty. The explanation was not released to the public until 1950, but the text has been republished recently as an appendix to a review, published in Italy, of early work on Marconi's magnetic detector⁶.

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a)



b)

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Intelligent battery management

Using experience gained from fast charging circuits for communications equipment, Philips has produced six ICs dedicated to charging and monitoring either nickel cadmium or nickel metal-hydride cells and batteries. These devices are collectively outlined in a booklet *ICs for NiCd and NiMH battery charging management*.

Fast chargers bring charge times down to 30 minutes or less from anything up to 16 hours. They are certainly not new but they have traditionally been enigmatic, expensive and complex. All safe fast charging circuits have needed involved discrete-component protection mechanisms. Combining fast charging with high efficiency has involved the use of switch-mode technology on top of complex overcharge and faulty battery safeguards.

The six battery management ICs fall into three categories. One category, comprising the *TEA1100* and *TEA1088*, provides charge

monitoring and control dependent on battery characteristics. Both are switch-mode chargers offering variable $-\Delta V$, cell temperature, voltage threshold and charge-time sensing. Both also provide pulsed trickle charging following full charge detection to keep cells in good condition. They differ in that one is optimised for either mains isolated or non-isolated charging and one is designed for non mains isolated charging. For readers not familiar with $-\Delta V$, it represents the falling voltage that a cell exhibits if you attempt to charge it further after it has reached its fully charged state.

In the second category is the *TEA1090* self-oscillating power supply. This device is designed to operate directly from rectified mains at up to 450V DC with 750V transients. Requiring very few external components, it has a built-in mos power switch capable of charging at up to 1.2A, and it is very efficient since both primary

and secondary currents of the oscillator transformer flow in to the battery.

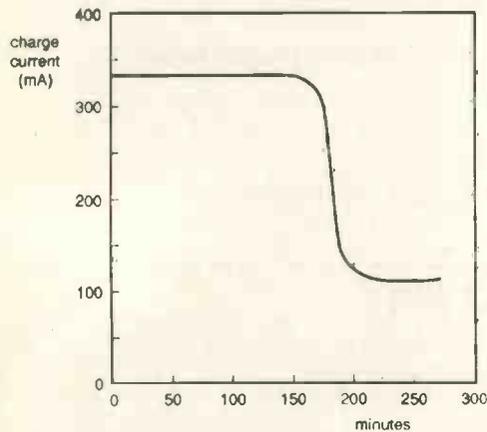
Three ICs in the third and final category provide charge monitoring and control. Two have outputs for LC or LED displays indicating charge status in 20% steps to 100%, together with an output controlling full-charge/trickle switching. These are the *PCA1329* optimised for static loads such as cordless telephones and the more highly integrated *SAA1500* for dynamic loads presented by camcorders, notebooks and cellular radio equipment.

Metal Hydride batteries

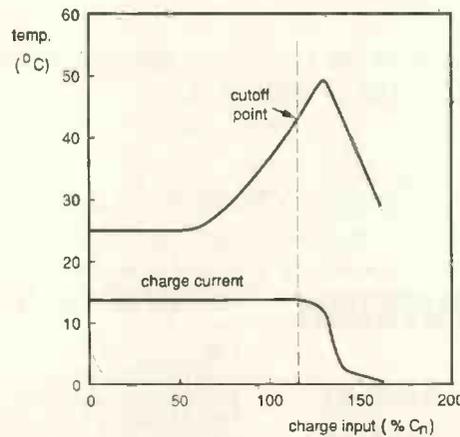
At present, small-cell rechargeable batteries in portable electronic and electrical equipment are usually NiCd or the new NiMH (nickel metal-hydride) technology introduced into the market last year. The main advantages of this new type of battery over the NiCd types are:

- 30% to 50% more energy per unit volume
- Compatible with existing NiCd charging systems
- Absence of poisonous cadmium makes them safer for the environment
- No memory effect; i.e. they don't have to be completely discharged to allow recharging to full capacity.

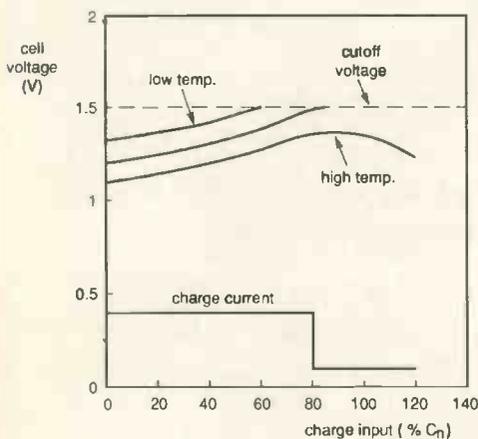
It seems likely that NiMH batteries will replace NiCd types within the next decade. However, they currently cost twice as much as NiCd batteries, are not yet generally available in large quantities and charger systems for NiMH cells must incorporate stringent electronic control to avoid danger of explosion due to overcharging, or limitation of cycle life due to prolonged trickle charging.



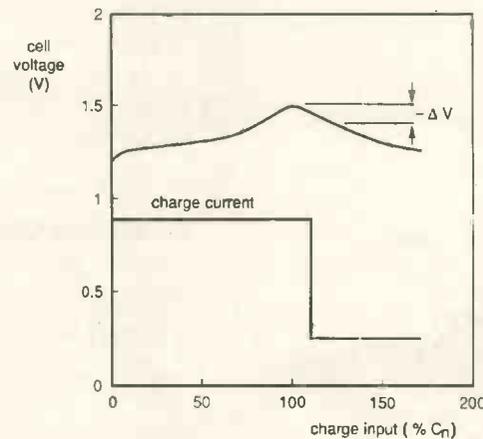
(a)



(b)



(c)



(d)

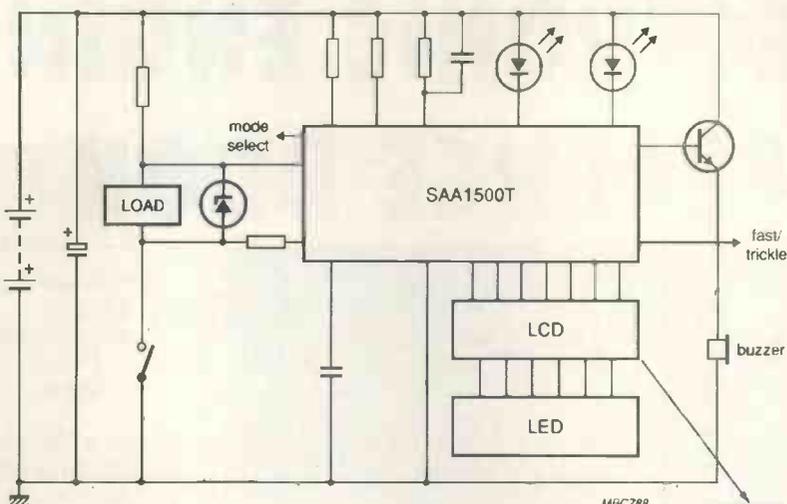
When is it charged?

Various parameters can be monitored to determine when a nickel cadmium cell is fully charged, namely charge time, temperature rise, voltage threshold and $-\Delta V$. The simplest method, shown in (a), is to switch off the charging current after a preset charging period under control of a timer. As shown in (b), the charging current can also be switched off at the maximum permitted cell temperature (45°C for example). Figure (c) shows the charging current switched off when the cell voltage reaches the fully-charged level (1.5 V). Finally (d) shows the charging current being switched off after sensing the negative slope of the cell voltage ($-\Delta V$) that occurs after the cell is fully charged.

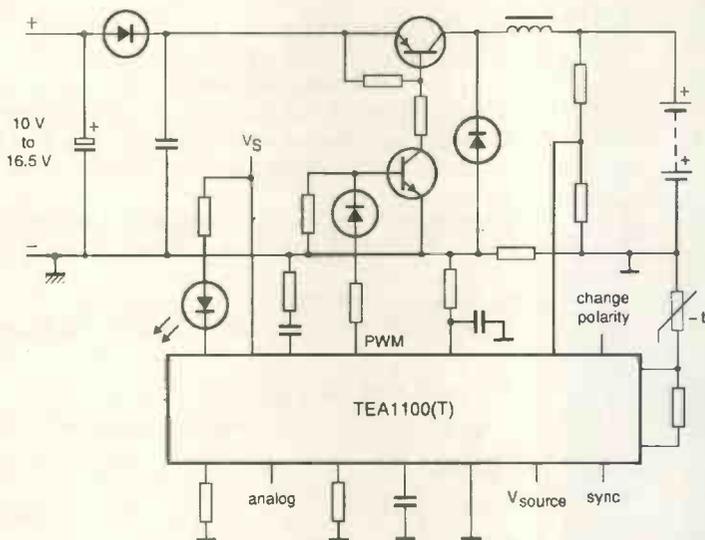
The *TEA1041* is the final IC in the third category. This is a battery low indicator with adjustable threshold. It filters against transient battery loading and provides both low and critically low indications. Also outlined in the booklet are NiCd charge characteristics, the pros and cons of NiMH cells and design requirements for efficient fast chargers.

Philips Semiconductors, PO Box 65, Philips House, Torrington Place, London WC1E 7HD. Telephone 071 436 4144.

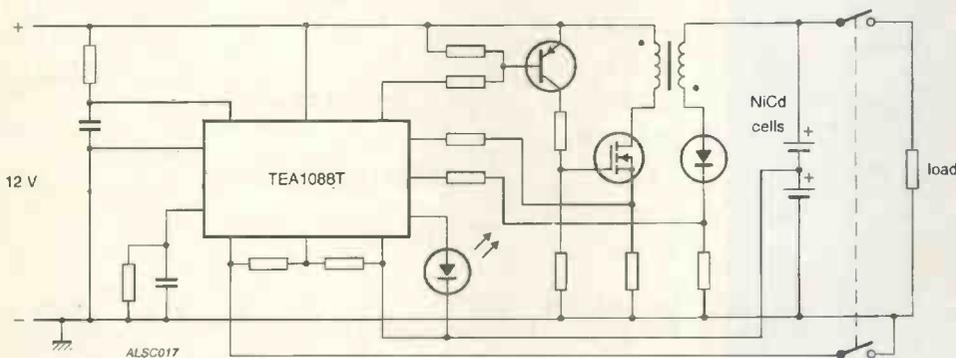
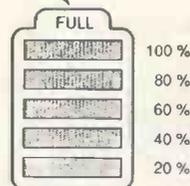
Primarily a remaining energy indicator for dynamic loads, the *SAA1500* also has an output for switching between fast and trickle charge current levels.



MBC788



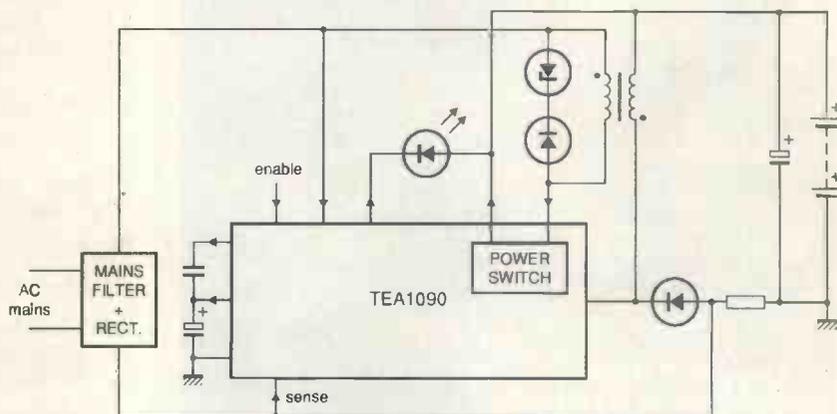
Left, Switch mode charger for NiCd or NiMH cells operates in forward conversion mode and provides comprehensive protection against faulty batteries. To determine when the battery is fully charged, the *TEA1100(T)* calculates/detects the $-\Delta V$ by measuring the battery voltage while the charge current is inhibited, and by using a sample and-hold circuit for digital filtering. An adjustable timer is also available to switch the charging current after a set period. To prevent damage to the battery, the circuit also allows the battery temperature to be monitored with thermistor, and a limit to be set with hysteresis) above which only trickle charging is allowed.



ALSC017

Below, Highly efficient and requiring few components, this mains charger for NiCd cells has a built-in power mosfet. It can provide up to 1.2A when battery voltage is less than 2.5V from a 450V supply rail.

Above, *TEA1088T* monitor and control circuit for non mains isolated SMPS battery charging systems. The device is being used in this application to monitor the battery and control an SMPS flyback converter. The fully-charged condition is detected by monitoring the $-\Delta V$ of the battery without drawing current. An adjustable timer is also available for determining the fully charged condition. After detection of the fully-charged condition, the circuit switches from fast charging to pulsating trickle charging by controlling the direct drive to an SMPS switching transistor in the power module. In the discharge mode, a battery voltage low-level detector provides an output for driving an LED or a buzzer. The device also includes dynamic limiting of SMPS transformer primary current.



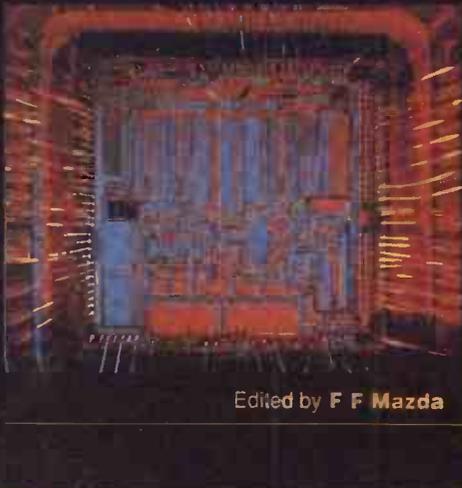
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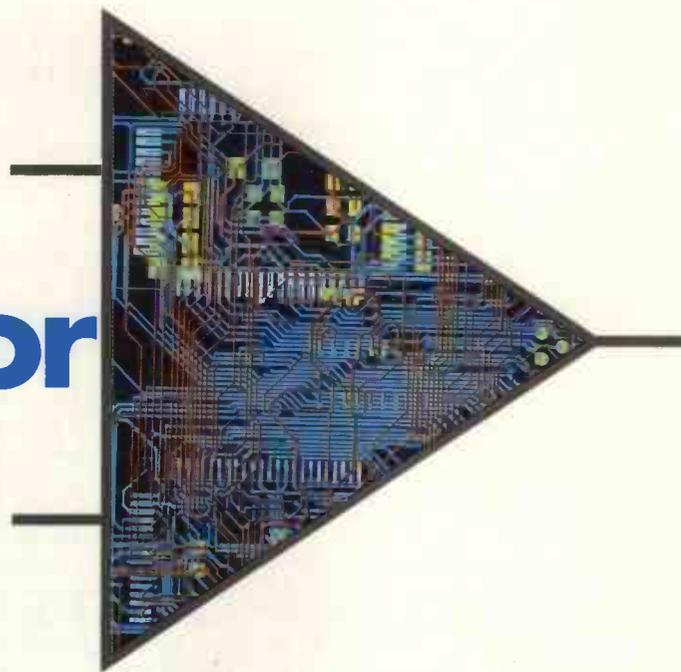
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Ruling principal for balancing op amps



Calculating component values around a multi-input summing amplifier can be laborious. David Gibson shows how sticking to the rules can help relieve the tedium.

There is clearly a need for a simple method for calculating the component values around a multi-input summing amplifier. To this end I have developed an impedance balancing rule for op amps that can be applied where an op-amp sums two or more signals using both the inputs, or where a bias voltage is applied to the positive terminal.

The rule is novel – I have not seen it quoted previously – and at any rate is certainly extremely useful.

Calculating component values can be tedious.

Figure 1 is an example of a “difficult” design where three signals need to be summed and filtered. Only the first step – calculating the input resistor for U_B – is easy.

The problem with the analysis is that the voltage at the positive input terminal needs to be calculated in advance, leading to some messy algebra. Fortunately, the new rule cuts out much of the algebra.

The impedance balancing rule:

“The gain from any input of a summing amplifier is simply the feedback impedance divided by that input impedance, provided that the sum of the admittances connected to the inverting input of the op amp is equal to the sum of the admittances connected to the non-inverting input. Given this condition, amplifier bias currents will not contribute an error”.

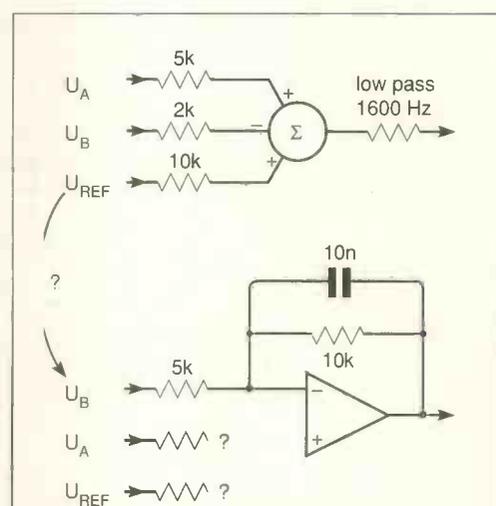


Fig. 1. Obtaining cancellation of bias currents in a summing amplifier requires a matching of admittances.

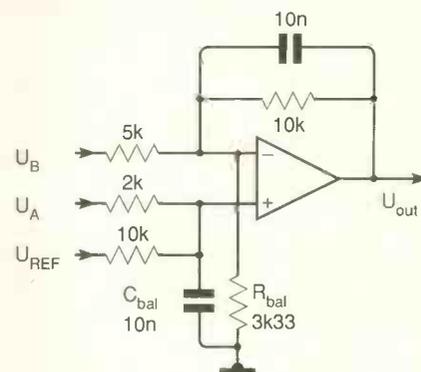


Fig. 2. A practical implementation of Fig. 1. Components R_{bal} and C_{bal} are included to balance the impedances and lead to the correct gains from each input.

The impedance balancing rule should perhaps really be called admittance balancing – but admittance is simply the reciprocal of impedance. The impedances do not have to be resistive, but must take into account the signal source impedances.

Feedback impedance is treated in the same way as the input impedances. Applying the rule to Fig. 1. could not be simpler. The input resistances are immediately known (Fig. 2) and all that remains is to “balance” the impedance at each input by connecting a capacitor C_{bal} from the positive input to ground, and a resistor R_{bal} from the negative input to ground. Ground is treated as just another input, but being at 0V it contributes nothing to the output.

The values required for $C_{bal} = C = 10\text{nF}$ and $R_{bal} = 3.33\text{k}\Omega$. Notice that there are no unique values for R_{bal} and C_{bal} – a $5\text{k}\Omega$ resistor from

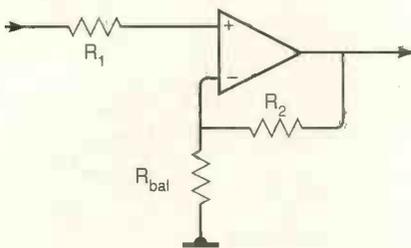


Fig. 3. The impedance balancing rule applied to a non-inverting amplifier. $G = R_2/R_1$. $R_1 = R_b || R_2$ is added to balance the impedances. This arrangement cancels out bias current effects.

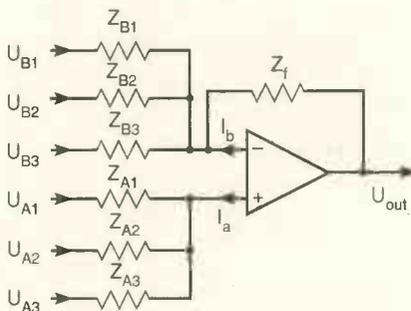


Fig. 4. Impedance balancing rule. The gain from each input is given by Z_f/Z_{in} provided that the impedance balancing condition is met, as described in the text.

the positive input to ground and a $2\text{k}\Omega$ resistor from the negative input to ground would provide a clearer representation of the required condition.

Effect of unbalancing

The easiest way to achieve a balance is to make sure that the same value of component is connected to both inputs. If this is not possible then some compromise in component values may be necessary, or it may be decided to omit a high value balancing component.

The effect of this “unbalancing” can be quantified as:

“If the impedances are not balanced then the gains of all the channels connected to the positive input must be increased by a factor equal to the ratio of the sum of the admittances connected to the non-inverting input of the op amp divided by the sum of the admittances connected to the non-inverting input, provided that the amplifier input currents (bias and offset) are negligible”.

Bias currents

One effect of the rule is that the input currents to the amplifier see equal impedances and so contribute the minimum to offset voltage. Applying the rule to a single input non-inverting amplifier leads, by an unfamiliar route, to the familiar configuration for minimum bias current effect shown in Fig. 3. In this case the gain is given by R_2/R_1 and R_1 is chosen to be $R_{bal} || R_2$ to fit the impedance balancing rule.

Analysis

Analysis is straightforward, though the notation may be a bit confusing. Referring to Fig. 4, the amplifier input currents are denoted by I_a, I_b , the input voltages and impedances for the non-inverting input by U_{ai}, Z_{ai} ; those for the inverting input as U_{bi}, Z_{bi} ; and the voltage

at the amplifier input terminals (usual “ideal” assumptions) by U_x .

The currents at each op amp input are summed:

$$I_a + \frac{U_{a1} - U_x}{Z_{a1}} + \frac{U_{a2} - U_x}{Z_{a2}} + \dots \quad (1)$$

$$\text{and } I_b + \frac{U_{b1} - U_x}{Z_{b1}} + \frac{U_{b2} - U_x}{Z_{b2}} + \dots + \frac{U_{out} - U_x}{Z_f} \quad (2)$$

re - arranging to obtain U_x :

$$U_x \cdot \sum_i \frac{1}{Z_{ai}} = I_a + \sum_i \frac{U_{ai}}{Z_{ai}} \quad (3)$$

$$\text{and } U_x \cdot \sum_i \frac{1}{Z_{bi}} + \frac{U_x}{Z_f} = I_b + \sum_i \frac{U_{bi}}{Z_{bi}} + \frac{U_{out}}{Z_f} \quad (4)$$

U_x can now be eliminated from (3) and (4), and U_{out} written as:

$$\frac{U_{out}}{Z_f} = K \cdot \sum_i \frac{U_{ai}}{Z_{ai}} - \sum_i \frac{U_{bi}}{Z_{bi}} \quad (5)$$

where

$$K = \frac{\sum_i \frac{1}{Z_{bi}} + \frac{1}{Z_f} + I_b}{\sum_i \frac{1}{Z_{ai}} + I_a} \quad (6)$$

The impedance balancing rule requires K to be 1. Equation (6) shows that the condition for this is that the reciprocal sum of the impedances connected to the inverting and non-inverting inputs are equal. In addition, I_a must be equal to I_b . For bias currents this is true by definition, so apart from the contribution from offset current, and provided that the impedance balancing condition is met, the amplifier input currents do not contribute to gain errors, and the gain can be written as in (5) by setting $K = 1$. If the circuit is unbalanced then (5) still describes the rule, but with the weighting factor, K , defined as in (6) and in the accompanying text. ■

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LETTERS

Living in the neutral zone

Roger Castle-Smith (*EW + WW*, July) commented that there is no rational reason for swapping the live and neutral leads to an amplifier; in fact there is a grain of truth in this piece of folklore.

Where transformer windings are wound on the same former, as they will be on a toroidal, they are capacitively (as well as magnetically) coupled; and the strongest coupling is between adjacent layers of wire. This means that the way the leads are connected has an effect on the electrostatic coupling of hum and mains noise. This is to some extent true even of singly screened toroidals.

The best arrangement is to place the primary on the inside with the neutral covering (and so shielding) the live. The secondary should then have its two centre tap leads taken from next to the primary and from the final outside winding. When these two wires are earthed the secondary is shielded from the primary and the rest of the electronics from the secondary.

Of course the shielding is imperfect and hum can be magnetically coupled, but there can still be a worthwhile improvement on sensitive instrumentation which often has a preferred live/neutral arrangement.

However, I am sceptical that it would have a detectable effect on modern audio where there is something seriously wrong if one can hear any noise or hum at normal distances from the speakers.

RA Woolley
London

Overlook THD at your peril

In a typical non-inverting op amp configuration, the baseline low-frequency THD level is generally dictated by the predominantly second harmonic common-mode gain non-linearity of the input differential stage. This doesn't vary greatly with feedback, being slightly higher for low gain settings with their higher common mode signal levels for a given output. The published performance of the AD845 typifies this behaviour.

Ian Hickman appears to have

overlooked this source of THD in verifying the claimed THD performance of the Burr-Brown OPA2604 (*EW + WW*, Design Brief, July 1992).

His Fig. 1a test circuit, drawn from the Burr-Brown data sheet, effectively lumps in the $x+1$ common mode non-linearity with the x 100 differential non-linearity and noise, and interprets the $x+1$ (non-inverting) THD to be 1/100th of the measured THD.

This in effect devalues the common-mode baseline by 40dB. Indeed, the Burr-Brown data sheet shows the baseline non-inverting THD plus noise at 0.0003%, 0.003%, and 0.02% for signal gains of $+1$, $+10$, and $+100$, respectively, at a fixed output of 3.5V RMS.

However, on rescaling the extreme case $G = 1V/V$ up to x 100, and comparing the circuit end measurements with the standard $G = 100V/V$, it can be seen that THD plus noise is up 50% at 0.03% from 0.02%, while the only effective parameter change is a 100 times larger common mode signal.

The noise gain of both is 101 and, assuming a zero contribution from the AP test set, the typical $11nV/\sqrt{Hz}$ voltage noise of the OPA2604 contributes 0.01% to the THD plus noise when measured with the 80kHz three-pole filter specified. Hickman used a 20kHz BW filter giving 6dB lower noise or 0.005%.

If the generator noise in the AP test set can be neglected, and the readings are to two significant figures, it could be concluded that between 0.016% and 0.028% THD is present due to common-mode non-linearity. A preferable test method would use the harmonic/DSP facility on the AP test set to evaluate and sum individual harmonics and plot the resultant THD without wideband noise masking the results or having to resort to spacious scaling methods.

Greg M Ball
Coolangatta Australia

Friendly request

With reference to John Linsley Hood's article "An integrated audio amplifier" (*EW + WW*, June), I noticed that the CCT nominates on output Z were 8 Ω .

As commercial speaker combinations have a range from 4 to 16 Ω , are there any figures correlating Z, output power, and THD over this range?

It would be appreciated if a range of output impedances, power output, and THDs were quoted by somebody obviously more friendly with this particular amplifier.

BH Shardlow
Bracknell Berks

BH Shardlow was one of a number of readers making a similar point. Part of the problem was caused by a production error which meant that the following table was missed out when the article was printed - Editor.

P_{out}	TR_1	VA	C_7/C_{107}
30W	30-0-30	75	50V
40W	40-0-40	100	64V
50W	45-0-45	120	80V

Radio request

I am carrying out a survey of broadcast and LW transmitters operated in Europe between 1939 and 1945, and the power grid existing at that time.

It is my thesis to indicate that the non-use of fundamental harmonics for loops and positioning of aircraft before introduction of H2S, G-H, and Oboe was intentional.

Radio Luxembourg, 250kW, Swiss and Swedish signals were never used by RAF navigators for running fixes on 11 to 12h trips to central Europe and Italy.

As an RAF navigator, 106 Squadron, Metheringham, I am concerned. Any information regarding frequencies, field-strength patterns, and harmonic analysis that your readers can supply will be vital to my analysis.

I wish to prove that the entire generation and transmission of electrical power could have been destroyed by DH Mosquitoes in three nights in the fall of 1943 using the harmonics of the power lines and broadcast transmitters.

George A. Collins
P O Box 11 Lynden Ontario
LOR 1T0 Canada

A matter of facts

I wonder whether those who feel it worth while to spend a lot of money on some of more dubious beliefs

Faulty Fawly

I could not agree more with you over the attempted imposition on us of a wide screen format (*EW + WW*, Comment, June). One suspects if a poll were held, the pro-wide screen lobby would lose heavily.

Draw a parallel with Cinemascope, where to some extent there were some benefits of the larger format, and look what happened to that.

French and German television regularly transmit in D2-Mac programmes in 16:9 format but, with the limited amount of original material available, usually old Olympics coverage, they resort to ancient conventional format films and videos (I saw a Fawly Towers episode in it on German TV).

The snag is, a little is lost from the top and bottom. Why do they do it?

Reg Williamson
Kingsgrove Staffs

about audio equipment do the same with their video equipment

A video system, having a much wider bandwidth than an audio one (at least, in respect of terminal signals at each end) and with transient response of greater importance, should offer a splendid proving ground for some of these ideas. In addition, the output transducer, the CRT, gives a remarkably comprehensive display of consequences to the eyes; human senses comparable to ears to say the least.

I have hesitated to raise this issue lest it trigger a spurious market in unduly expensive gold plated antennas and high loss, that is directional, feeders. But if it arises I hereby stake a claim to a modest royalty on sales!

More seriously though, I deplore the replacement of much of the objective testing of review equipment by personal comment in some of the audio magazines. Fact rather than opinion is so absent these days that reviews often fail to report even basic information such as whether a tape recorder has a microphone input facility.

RH Pearson
Bourne Lincs

Mixing Rs and ohms

Referring to my circuit idea "Preset on time for battery equipment" (*EW + WW*, April) and the letter "Saving batteries" from Gaines Crook (July), the circuit referred to by Crook was the "editorial alternative" to my idea. The 10R referred to is 10Ω, not ten times R. It limits the inrush current, to prevent damage to the capacitor or switch contacts.

The values of R and C have to be chosen for the circuit operating period. Capacitor, C, will charge through the 10Ω resistor and its voltage will almost reach that of the

supply. When it discharges (through resistor R) its voltage will decay exponentially. The current to the load will switch off when the capacitor voltage is approximately equal to half the supply voltage.

Resorting to a little math: $V_c = V_s e^{-t/cr}$. switching will occur at $V_c = 0.5V_s$. Therefore $0.5 = e^{-t/cr}$ or $\log_e 0.5 = -t/cr = -0.693$. Simplifying this we get $t = 0.693cr$.

I decided to use a 555 timer in preference to the editorial alternative. It is physically smaller and cheaper (and I had some at the time). I also required a led indicator; this could be provided in the alternative circuit by using the

unused inverters. The editorial alternative requires the battery voltage to be applied directly across the cmos IC. Its quiescent current is only about 3nA, probably less than the transistor leakage current.

My circuit was part of an alarm system that used a radio transmitter. The momentary switch was in fact a trembler switch. Low power consumption during inactivity was necessary because the intended application was to protect semi-fixtures, such as oil paintings. By attaching the circuit to the painting's frame, any movement of the painting would result in an alarm being transmitted.

This alarm system was not developed further, but readers may like a project. I intended to use a lithium battery because of their long shelf life. The intended radio transmitters were the devices from Radiometrix. These devices use an unlicensed band.

Steve Winder
Ipswich

Corns of ears

Hot Carrier need not worry about the problem of retrofitting RF suppression to existing hearing aids (*EW + WW*, White Noise, June), at least as far as NHS users are concerned.

The government has already taken action that will negate the problem - it has changed the material used for the ear mould to one that causes corns, sweating, and nocturnal discharges, so they are not being used.

To the savings on remoulds and batteries is now added the benefit of not having to retrofit RF suppression, definitely an excuse for a champagne party at the DHS.

PJ Crawley
Basingstoke

Make the earth move and win \$\$\$\$

As long as you guys are humouring outlandish debates about the question of the earth's motion, here's my two cents in favour of Amnon Goldberg's fixed earth position (*EW + WW*, April). I think he's right - the earth is central and stationary in a rotating cosmos.

I am offering \$1000 for proof that the earth moves. The proof must be direct, observable, physical, natural, repeatable, unambiguous, and comprehensive. The award is

divided into \$500 for scientific positive proof that the earth orbits the sun, and \$500 for proof that the earth rotates on its axis.

Does Euan Orr (*EW + WW*, June) have anything more than pigs and porpoises or rocket-ridiculing rhetoric to offer against the idea? A lot could depend on this question about the earth's status in space.

RG Elmdorf
Bairdford
Pennsylvania

Central arguement

If Euan Orr believes in relativity (June), then he is obliged to accept that the possibility that earth could be the centre of the universe, as could the moon, Pluto, Alpha Centauri, or Arcturus! "The geocentric paradigm is at least as good as anybody else's!" (Sir Fred Hoyle).

If he does not accept relativity, then how does he explain the zero-velocity result of the Michelson-Morley experiment, while the Michelson-Gale experiment gives a positive result? The most straightforward explanation is that the earth really is located at the centre of the universe in an absolute sense, and that it's the rest of the universe that's doing the moving, as the Bible states!

Amnon Goldberg
London

Music to my body

With reference to Elizabeth Davies' article "The healing face of electromagnetic fields" (*EW + WW*, April), this is not new knowledge. A group I study with in Nottingham know that the whole human body and its various components are tuned to a cosmic keyboard, therefore harmonic frequencies are beneficial and inharmonic ones are not.

Audio frequencies also can, if used at the correct rate, stimulate the pancreas, pineal, thyroid, lymph system, and so on.

Just as a wine glass will shatter at a high C sung by a person, other materials also have a natural resonating frequency. Going back to fundamentals, it is known that a conductor carrying electricity has a certain concentric field around it. Lock this into a variable frequency generator and you have the ideal tool to put requisite pulsing into an electromagnetic field to give the necessary healing to bone, tissue, or whatever system is being treated.

R Gough
Newark Notts

Catt calls

I take it that nobody took up my challenge in the June issue, and that the only response was the letter from Brian Clement in the July issue, which had no discernable bearing on the "Catt Anomaly" (*EW + WW*, September 1987).

All those with accreditation in electromagnetic theory, that is who earn salary or royalty or Nobel Prize on the back of it, keep their heads down, as usual. Could their students have a go at them and put something in writing?

I will give £50 to the first student who gets a reader in electromagnetism or equivalent to comment in writing on the Catt Anomaly. The editor of this magazine will judge (*Not if I can help it - Ed.*) the matter of whether the comment is a serious contribution.

Ivor Catt
St Albans Herts

Putting the Catt out

I am pleased to see that Ivor Catt is as irrepressibly anomalous as ever (*EW + WW*, June). Now he is challenging accredited experts to defend classical electromagnetic theory against his Anomaly, and offering to pay them as well, though mostly with other peoples' money, it seems.

This really is all a bit of a joke isn't it? A wave can propagate at a characteristic speed in a medium without any physical entity involved in the propagation having to move at anything like the same speed as the wave itself. This is as true of electrons in transmission lines as it is of little bits of vibrating spring or drops of ocean sloshing about.

Catt's Anomaly rests on the opposite assumption, that for an electromagnetic wave to propagate down a transmission line at the speed of light it is necessary that electrons move at or faster than the speed of light. Such an assumption is entirely groundless. Groundless assumptions are no challenge to anything.

In correspondence following publication of Catt's 1987 article I illustrated this point using an (admittedly very non-linear) analogy involving a row of pennies, simple enough for an intelligent child to understand. The best Catt could do by way of reply was to challenge the origin of the analogy, thereby ducking the issue. Well, now you know. I invented it myself, but that is completely irrelevant to the validity of the argument.

Electromagnetic theory does not need defending against the Catt Anomaly. Defenders do not need to be accredited or even fully understand the theory. It does not even matter whether the theory is right or not. All that has to be shown is that Catt has no case whatever to answer.

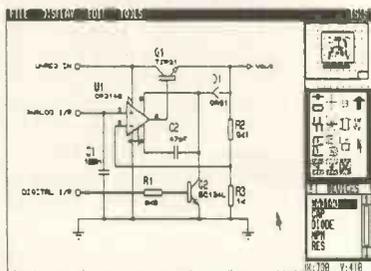
Perhaps this is why the accredited experts can't be bothered. I'll have a go if you like, I could do with the dosh - assuming, that is, the defence fund ever manages to exceed the original £100 plus interest.

Alan Robinson
Holgate Yorks

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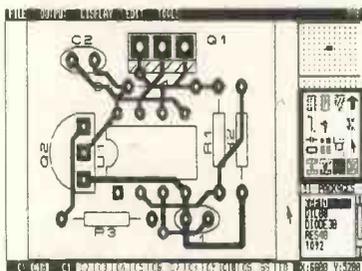
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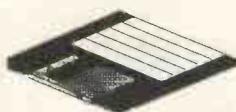
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Discrete active devices

Video transistors. Bipolar transistors in both polarities are available from ITT in leaded and SM form. Maximum collector saturation voltage is 0.6V (npn) and 0.8V (p-n-p), both at 30mA collector current and 5mA in the base. ITT Semiconductors, 0932 336116.

SM, RF transistor. Features of the *MMBR521L* small-signal p-n-p transistor from Motorola include 70mA maximum current and a noise figure at 1GHz of 2.5dB. It is in an SOT-23 package and is intended for use in high-gain, low-noise, small-signal amplifiers to 2.5GHz. Motorola Ltd, 0908 614614.

Digital signal processor

100ns DSP. Designed for high-volume, multi-media use, Zilog's *Z89320* is a low-cost digital signal processor, running at 10MHz with 16-bit single-cycle instructions, 100ns multiply/accumulate time and modified Harvard architecture. There are 4Kword of masked rom and two banks of 256word data ram; 24bit ALU, accumulator and shifter, 16-bit i/o port; six-level stack and three vectored interrupts. Gothic Crellon Ltd, 0734 788878.

40MHz DSP. At 3.3V, Motorola's *DSP56L002* 24-bit general-purpose digital signal processor uses a third of the power (165mW) of the *DSP56002* and is software compatible with it. The new design performs 20 million instructions/s with a 50ns cycle time from a 40MHz clock. Motorola Ltd, 0908 614614.

33Mips DSP processors. Claimed to be the fastest available, NEC's μ *PD7701x* family of general-purpose digital signal processors have a 30ns instruction cycle time and a 3-stage pipeline to give 33Mips performance, carrying out up to eight operations in parallel. The instruction set is similar to C. The devices have two serial interfaces running at 16Mb/s, four i/o ports and a parallel interface. Development tools running under Windows include an editor, assembler, linker, debugger, simulator and emulator. NEC Electronics UK Ltd, 0908 691133.

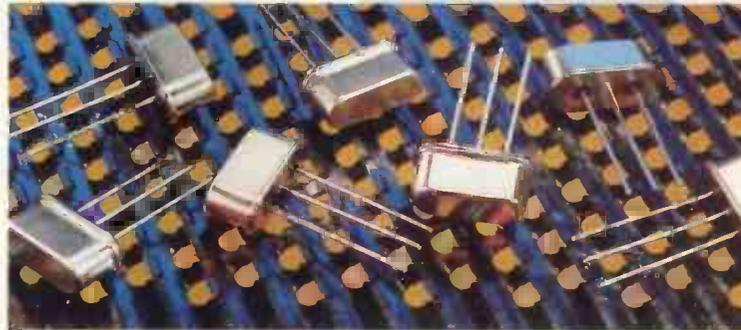
Linear integrated circuits

Quad op-amp. Settling time of less than 200ns to within 0.01% and a slew rate of 170V/ μ s are features of Analog's *OP-467* quad op-amp, which imposes a current drain of less than 10mA from \pm 5V to \pm 15V supplies. Bandwidth is 28MHz, offset better than 500 μ V and capacitive load handling 1600pF. Analog Devices Ltd, 0932 253320.

FM demodulator. GPS has the *SL1461* wideband PLL FM demodulator for use in satellite television tuners. Although packed in the small, surface-mounted SO16 form, it contains all the elements needed for an 800MHz system except external oscillators sustaining network and loop feedback components. High input sensitivity is combined with 1% distortion on video parameters and there is AFC, which can be disabled. GEC Plessey Semiconductors, 0793 518582.

850MHz buffer. A closed-loop buffer for video and high-speed communications, the *HFA1113* by Harris offers 850MHz bandwidth, 2050V/ μ s slewing, 11ns settling time to within 0.1% and 0.075dB gain flatness to 200MHz. It possesses a programmable output clamp to protect subsequent devices and programmable gains of +2, +1 and -1, the feedback R being internal. Gain error is 0.01V/V at unity gain. Harris Semiconductor(UK) Ltd,

50V/ μ s op-amps. Both low-power devices, the *LT1201* dual and *LT1202* quad 11MHz op-amps slew at 50V/ μ s but draw only 1mA per amplifier with supplies of +5V to \pm 15V. *LT1200* is a single op-amp version introduced last year. Maximum input offset is 1mV and settling time to within 1% of a



10V step is 430ns. Linear Technology (UK) Ltd, 0932 765688.

Dual op-amps. Recent additions to TI's Excalibur family are the *TLE2227/2237* dual op-amps, which exhibit 100 μ V offset and 13/50MHz performance. Open-loop gain is 153dB and noise voltages are 2.5nV/root Hz at 1kHz and 3.3nV/root Hz at 10Hz. Deep saturation in the output stage is prevented and overdrive recovery is rapid. Texas Instruments, 0234 223252.

Logic building blocks

Fast, low-voltage cmos. Working over a range of 1.3V-3.6V, the Philips HLL high-speed, low-power, logic and LV-HCMOS high-speed families have push-pull outputs swinging rail-to-rail, small over and under-swing and latch-up-free operation. HLL runs at twice the speed of Fast bipolar logic with a 2.5ns delay. Inputs can exceed supply up to 5.5V and the elements can be used as level shifters in mixed-supply systems. Gothic Crellon Ltd, 0734 788878.

Peripheral controller. *ZIP Z80182* by Zilog is an intelligent peripheral controller with standby modes for power saving and connection to PCs with no extra circuitry; it will bridge Apple and IBM machines. Frequency range is 16-20MHz and the two ESCC channels have built-in baud-rate generators, FMO encoding and DPLLs. The 24 parallel i/o lines include a printer interface, extra modem controls and display. Gothic Crellon Ltd, 0734 788878.

Graphics controller. At low power, Hawke's *VIDC20* high-performance video controller for LCD and CRT screens is a single-chip solution for displays in almost any system from workstations to games. It supports VGA, super VGA and XGA with up to 16million colours and includes a

Clock oscillator. IQD has an HCMOS clock oscillator in a crystal holder that takes up only 18% of the space needed by a 14-pin dip package, measuring 5mm high and 11.05mm by 2.65mm. Frequency range of the *IQXO-32* is 1MHz-70MHz with stabilities of \pm 100ppm or \pm 50ppm at 70°C. International Quartz Devices, 0460 77155.

hardware cursor, programmable pixel rates and a sound system. Hawke Components Ltd, 0256 880800.

Fastest bicmos. Claimed to be the fastest bicmos logic available, the ABCT family from National exhibits propagation delays of 3.6ns and is intended for 33MHz-plus computing. Bus disruptions are eliminated and datastreams protected during live insertion. From 0V to 5.5V, high impedance is guaranteed over full power up/down cycles. National Semiconductor, 01049 89-903 39 02.

Frequency synthesis. Analog Devices *AD7008* direct digital synthesiser provides usable analogue outputs at frequencies up to 20MHz at a claimed accuracy of one in four billion. Frequency, phase and amplitude modulation is controlled by the microprocessor, the amplitude modulator having two multipliers with sine and cosine values fed from rom. Quadrature data from the rom provides an SSB signal. Polar Electronics Ltd, 0525 377093.

Memory chips

70ns erasable prom. 3.3V, UV-erasable cmos proms from Micro Call, the *WS57LV291C-70T/90T* have maximum read access times of 70/90ns and a time from chip select to



Oscillators

Audio VCOs. The M2 series of 4-30MHz voltage-controlled oscillators for satellite receivers and digital audio processing is now available in the low-cost D110 10-pin sip package or the D150 ceramic SM package. A single lithium tantalate crystal resonator provides 20 times the frequency variation of quartz at $\pm 800\text{ppm/V}$. Dual and triple versions are available. Fujitsu Microelectronics Ltd, 0628 76100.

valid data out of 20/30ns. They are 2K by 8 devices and run at clock speeds of 12MHz, with no wait states. Micro Call Ltd, 0844 261939.

Microprocessors and controllers

Up-graded microcontroller.

Motorola's 68HC08 is a higher-performance development of the earlier 68HC05, which is object-code compatible with the HC05. CPU is an enhanced 9-bit core running at 8MHz and incorporating 78 new instructions, a 16-bit index register, a 16-bit stack pointer and stack manipulation instructions. Motorola Ltd, 0908 614614.

Mixed-signal ICs.

Audio codec. AD1848K Soundport stereo codec by Analog is a compatible development of the earlier AD1848J, which is designed into the Windows sound system and Compaq Deskpro computers. It provides A-to-D and D-to-A conversion, gain control and mixing of analogue and digital data streams to give CD-like quality on an ISA or EISA motherboard or card. The new version gives 46.5dB gain and attenuation of stereo inputs and 85dB dynamic range. Analog Devices Ltd, 0932 253320.

Vocoder. Qualcomm's Q4400 variable-rate single-chip vocoder is intended for digitised speech encoding in telephone, wireless communications and speech synthesis. It is a full-duplex device using the company's codebook excited linear predictive code for high-quality speech at 4kb/s or 8kb/s fixed or 1-8kb/s for frame-by-frame encoding, automatically adjusted. I/o

is by 8-bit parallel bus to standard microprocessor buses. Chronos Technology Ltd, 0989 85471.

Latchable multiplexers. DG4XX high-speed latchable multiplexers from Siliconix are claimed to offer the lowest power consumption and on-resistance. The 428 is a single-ended, eight-channel multiplexer and the 429 has four differential inputs that can be routed to a common differential output. On-resistance of both devices is 100 Ω and power consumption 2mW. Siliconix/TEMIC, 0344 485757.

Cellular audio interface. Voice-band audio processor by Texas, the TCM320AC39, provides transmit linear or PCM encoding and receive linear or PCM decoding, with transmit and receive filtering for a GSM digital cellular system. It connects directly to an electret microphone, piezo speaker and DSP. Operating power is 40mW; 1.25mW during power-down. Texas Instruments, 0234 223252.

Optical devices

Bright, SM and blue leds. H-P's range of leds is extended by the addition of high-brightness AlInGaP amber and red-orange types in T1 and sub-miniature packages; bright TS-AlGaAs red types on a transparent substrate; low-cost surface-mounted leds in yellow, green, orange and red; and blue leds on 481nm. Hewlett-Packard Ltd, 0344 362277.

Programmable logic arrays

110MHz PLDs. Up-graded speeds for Lattice's PLDs are announced. Clock frequencies now reach 110MHz, with propagation delays down to 10ns. The devices contain up to 8000 gates. Lattice Semiconductor (UK), 0753 830842.

Dense FPGAs. With 13,000 usable gates, Xilinx claims its XC4013 is the world's largest field-programmable gate array. Enhanced integration includes 576 logic resources or configurable logic blocks. Xilinx Ltd, 0932 349401.

Power semiconductors

Light tube chip. A new IC by Micro Linear reduces the power used by fluorescent light tubes by 35%, by supplying them with power at frequencies adjustable between 10kHz and 20kHz. Flicker is also eliminated and the chips contain a light-level sensor to reduce light output from the tube when ambient light increases. A starting circuit preheats the tube filament to extend tube life. Ambar Components Ltd, 0844 261144.

Intelligent power. Harris's CA3277

power IC is a dual 5V linear voltage regulator, one of the sections being enabled, and a number of data buffers. It is intended to preserve data when a main system is shut down, one of the regulators being then turned off. Level-translating buffers allow communication between local and remote controllers. Harris Semiconductor (UK), 0276 686886.

Varistors. New metal-oxide varistors from Harris are claimed to offer the highest peak current-handling rating for the size. The four 14mm and 20mm Series III devices have 9000A single pulse peak current rating and 120J energy rating in the 20mm package. Intended for use in transient-suppression devices, The V130LA20C (130V) up to V174LA20C (175V) are stressed to 100 pulses at 3000A. Harris Semiconductor UK, 0276 686886.

High-side driver. A fully-protected monolithic high-side relay driver switch, the IR6000 from IR has a clamp voltage rating of 72V and repetitive avalanche rating of 100mj; over current, over temperature and open-circuit detection are included. Operation is normal to 150°C. International Rectifier, 0883 713215.

Low-R mosfet. On-resistance of Siliconix's n-channel SMP60N06-08 mosfet is claimed to be half that of other TO-220 devices on the market at 8m Ω at 60V and 60A. An enhancement mode device, it is intended in the main for use in motor control and power conversion in cars and UPSs. Siliconix, 0344 485757.

Analogue/digital waveforms.

Thurlby Thandar offers the 8553, an arbitrary waveform generator providing both analogue and digital output from 0.01Hz to 50MHz. Simple waveforms are keyboard entered, while more exotic ones are composed on a PC and down-loaded via IEEE-488. The built-in digitiser allows signal capture without the need for any further equipment. Thurlby Thandar Instruments, 0480 412451.



Passive components

Coil chips. Although exhibiting an inductance up to 1mH and handling up to 80mA, the LQM32C surface-mounting chip coils by Murata are only 3.5 by 2.5 by 2.5mm. A polymer-based package provides magnetic shielding and allows dense board packing. Resistance lies between 13 Ω and 20 Ω and tolerance is $\pm 20\%$. Murata Electronics (UK) Ltd, 0252 811666.

SM electrolytics. Nichicon's WT range of electrolytic capacitors, which has a range of values from 0.1 μF to 100 μF at 4-50V, measures only 3mm diameter and 5mm in height. Leakage current is 3 μA and, depending on value and voltage, ripple can be up to 60mA. The capacitors are impervious to all the usual anti-solvent fluids. Nichicon (Europe) Ltd, 0276 685393.

EMI filters. Oxley's DLT2 TVS EMI-suppression filters now handle up to 60V working and transients to 83V. One feedthrough component will cope with both low-level conducted interference and high-level transients, using metal-oxide varistor clamps for transients up to 1.5j at up to 250A for 20 μs . Oxley Developments Company Ltd, 0229 52621.

Power resistors. Two smaller sizes of metal-oxide resistor from Welwyn, the MO1/2S 0.5W and MO1S 1W types, operating over -55°C to 235°C, the range now comprising 0.5W-3W types. The range of values is 10 Ω -50k Ω for the 0.5W resistors and 10 Ω -100k Ω for the 1W types, to 5% and 10% tolerance. Flameproof coating will not burn or emit incandescent particles, whatever the temperature. Welwyn Components Ltd, 0670 882181.



Connectors and cabling

Miniature connector. Miniature, circular connectors from Amphenol, the IP67-rated *C091D* range, is an upgraded version of the earlier *C091A* type and is intended for use in more hostile areas where it may encounter dust and temporary immersion in liquids. It is in die-cast aluminium in 1-8, 12 and 14 way versions. Amphenol Ltd, 0227 773200.

Modular connector. When access is difficult or the connector is hidden, the modular connectors introduced by Hypertac ensure first-time mating. The design is based on the existing HLM assemblies and can be supplied in a number of styles. Hypertac Ltd, 081-450 8033.

RF connectors. Conhex (SMB) RF connectors now have new features to give more consistent mating, greater packing density and time protection. No separate spring ring is needed, giving more constant mating/disengagement and allowing wear surfaces to be plated. A square PCB-mounting base is used for close packing and there is a closed entry to sockets to ensure straight mating. ITT Cannon/Sealectro, 0256 473171.

Smallest RF connector. Murata claims its ESC coaxial connector to be the world's smallest surface-mounted RF type, with a height of 3.2mm when mated and a board area

of 11.55 sq. mm. Voltage standing wave ratio up to 3GHz is less than 1.2 and impedance is 50Ω. Murata Electronics (UK) Ltd, 0252 811666.

Hardware

Fuseholders. For type *G* and *D* fuses, Rafi's new holders comply with IEC and VDE regulations. Versions available include illuminated and waterproof types, connections being quick-connect, tin-plated solder, nickel-plated brass or screw terminals. Rafi (GB) Ltd, 0737 778660.

Instrumentation

Cellular test set. Error-free in-car telephone installation every time is claimed for the Triplet *5050E* portable cellular test set, which checks over 40 functions, including registration, paging, SAT frequency measurements including deviation on all three tones, power and a facility for earphone and microphone test. An antenna coupler allows operation when the existing one cannot be removed. Amplicon Liveline Ltd, (Free)0800 525 335.

Digital multimeter. ITT's *Asyc* range of hand-held digital meters are waterproof. They are provided with a bar graph display for rapid reading as well as with the digital readout. Model *MX52* offers RMS, dB and frequency measurements on a 5000-count display to within 0.1%. Amplicon Liveline Ltd, (Free)0800 525 335.



control, and panel instruments. First-time entries are data acquisition boards in the 200 series, including the *PC226* for the PC. Amplicon Liveline Ltd, (Free)0800 525 335.

EMC tutorial. ERA Technology is to hold a tutorial entitled *The EMC Directive: a Progress Report*, for which programme and booking forms are now available. It is to be held at the Lucas Advanced Engineering Centre in Solihull and will review the current position in Britain, with discussion. Speakers from the Commission of the European Communities, DTI, IBM, BSI Testing and ERA will take part. ERA Technology Ltd, 0372 374151 ext.2234

PC data acquisition. A short brochure from Intelligent Instrumentation describes a range of software and hardware for data acquisition on a PC. Intelligent Instrumentation, 0923 896989.

EM compatibility. Educational material from RFI on electromagnetic compatibility is detailed in a new catalogue. Information sheets are available on surge immunity, mains flicker, mains harmonics and conducted RF immunity; and guidance described are on EMC management, design for EMC and EMC testing. The catalogue is free. Radio Frequency Investigation, 0256 851193.

Power supplies

Flexible UPS. *GUPS 2400* by Elgar is a global uninterruptible power supply for loads up to 2400VA, with a universal input for use anywhere in the world. AC input is automatically selected from the 85-140V and 170-280V ranges at any frequency between 45Hz and 450Hz. Internal or

DMM calibrator. All the functions of hand-held digital multimeters can be tested, calibrated and documented by Wavetek's Model *9000* multifunction calibrator. When connected to a printer, it will produce certificates for quality management programs such as ISO9000. The relevant test program is initiated by plugging in a card containing software from a comprehensive library of procedures. Wavetek Datron Division, 0603 404824.

external batteries provide backup of between 5min and 18min at 2400VA and there are versions accepting 24V DC. The units are rack-mounted. Schaffner EMC Ltd, 0734 770070.

Radio communications products

IF-L-band interface. *P700* by Paradise is a fully synthesised 70MHz-to-L-band up/down converter, designed as a companion to the company's *P200* and *P230* satellite modems to form an Intelsat IBS/Eutelsat SMS converter in the 925-1525MHz transmit range and 950-1750MHz for reception. *P700* uses dual conversion with a 125kHz step size. Paradise Datacom, 0376 500340.

SWR bridge. ZRC from R&S is an SWR bridge operating over the frequency range 40kHz-4GHz. Directivity up to 2.5GHz is better than 40dB. Calibration accessories are included, as is an adapter for connection to a test generator without cables. Rohde & Schwarz UK Ltd, 0252 811377.



Optical reflectometer. Tek's *TFS3030 FiberMini* optical time-domain reflectometer offers single or dual wavelength analysis, automated fibre analysis, multiple measurements, control over measurement parameters and a display with infinite zoom. Basic model gives 18dB measurement for a 0.5dB loss up to 55km from the front panel. Tektronix UK Ltd, 0628 486000.

IC identifier. *PC-86* is a pocket tester made by UEI that identifies unknown ICs in TTL 74, CMOS 40 and 45 series and 4164, 41256 and 44256 drams, testing them if required and displaying on a 16-character LCD. Citadel Products Ltd, 081-951 1848.

Literature

Amplicon. International catalogue from Amplicon Liveline is in three sections: industrial data communications, data acquisition and

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

Flat-plate antennas. Single and double-patch flat-plate antennas for 400MHz-16GHz mobile operation are available from Suhner, offering 10-20% bandwidth, 6-12dBi gain, high front-to-back ratio and either horizontal or vertical polarisation. The antennas are suitable for incorporation into building facades. Suhner Electronics Ltd, 0869 244676.

Microwave video link. Several new microwave links by Wood and Douglas carry sound as well as colour video. Working at either 10GHz or 12GHz, this miniature equipment comes in three versions for 10mW, 60mW and 400mW operation; both the more powerful types carry the sound. An external patch antenna with 14dBi can be supplied, but the transmitters have an internal antenna. Supply needed is 12V. Wood and Douglas, 0734 811444.

Switches and relays

Solid-state relays. *PLA110* and *PAA110* solid-state relays by CP Clare are in 6 and 8-pin dip packages and handle load voltages to 400V, peak currents to 150mA and exhibit a 22Ω on resistance, switching in 1ms. CP Clare Corporation, 0460 41771.

Transducers and sensors

Hall sensors. Low power consumption and a 10mA current allow Honeywell's *SS19* and *SS49* miniature Hall-effect position sensors to be used in portable equipment in which no buffering is needed to interface with most circuitry. They produce amplified outputs linear to within 1%. Honeywell Control Systems Ltd, 0344 424555.

Gas sensors. *NAP-5A* and *6A* are improved versions of the Nemoto *NAP-7A* and *3A* hot-wire flammable-gas sensors, offering a 50% improvement in sensitivity and very linear output. The *5A* is for fuel gases such as town and natural gas, and the *6A* is meant for liquid petroleum gas. Quantelec Ltd, 0993 776488.

state sram is available. Loughboro Sound Images Ltd, 0509 231843.

Development and evaluation

Windows FPGA design. Actel offers a Windows 3.1-based interface for its FPGA design systems, Designer and Designer Advantage, at no extra cost. New tools in this new environment include ChipView to show place-and-route results and ChipEdit to allow pre-placing of critical areas. Actel Europe Ltd, 0256 29209.

8051/Z80/68HC11 development. Cactus Logic's IDS/LC 8-bit microcontroller development system is compatible with a wide range of standard compilers, including *2500AD*, Avocet, IAR, Keil and Intermetric. It runs on a PC and operation is similar in appearance to the Turbo C interface. Watch windows, intelligent breakpoints that trigger others, simulated i/o capability and real-time run with zero wait states are all features, as is the capability of executing up to 1000 breakpoints. Great Western Instruments Ltd, 0272 860400.

16-bit micro emulation. PC-based emulators for Hitachi's *H8/500* series of 16-bit microcontrollers are now available, allowing target systems to be debugged at full system speed. Each is an interface card and a personality module for the specific type of micro and takes up two slots in a PC or one slot with the module at the target. 1Mbyte of ram is available to the user. A full set of software is provided. Hitachi Europe Ltd, 0628 585000.

Universal programmer. The Eclipse universal programmer by Stag supports all major categories of memory, including bipolar proms, cmos gals and Flash. *EPU48D*, catering for 8-48 pins and the *EPU84P* for 20-84 pins can both be up-graded to handle 255-pin drive and are ram-expandable to 64Mbit. The unit can be either locally controlled by keypad or run under Windows or dos on a PC. Stag Programmers Ltd, 0707 332148.

Computer peripherals

Keyboard-port interface. Aptec's Keylink plugs into PC and keyboard sockets to form an operator push-button interface, allowing the user's choice of buttons to be connected while retaining the normal keyboard operation. The extra keypad or push-buttons may be up to 10m from the PC. Since the Keylink supports standard reset and test functions, normal boot-up and reset can be performed in the absence of the PC's keyboard. Aptec Ltd, 0706 358362.

Smart keyboard. An IBM-MP extended keyboard by Cherry is meant for use at checkouts, in control and in luggage management, having a built-in bar-code reader and a number of special keys such as a laser key for switching between a



multi-resolution stylus and a scanning gun. Code identification is automatic and a CCD scanner and slot reader may be connected. Cherry Electrical Products, 0582 763100.

Software

Mathematical modelling. Version 1.2 of *VisSim*, an interactive maths modelling program by Adept Scientific, provides cosh, sinh and tanh functions, transfer function and FFTs. Options now include Neural network and C-code generation from block diagrams. A block diagram with unlimited sub-blocks down to component level contain enough information for the program to run a simulation of the process. Adept Scientific Micro Systems 0462 480055

8051 simulator. *Simice-51*, developed by Raisonance, enables faster 8051 software development by debugging code before the target board is available. It supports all 8051 derivatives, including IIC versions, providing high-level language support for *C-51* and *PLM-51*. Variables and registers are dynamically up-dated on screen and speed control is provided. Logicom Communications Ltd, 081 756 1284.

Device programmer. Nohau's *HI-LO ALL-07* device programmer handles up to 256-pin devices and has a high-speed parallel interface, incorporates component test and PLD vectors. There is a 40-pin dip zif socket and modules for high-speed four-gang and eight-gang programming. Communication is by parallel interface card for fast down loading and software is supplied. Nohau UK Ltd, 0962 733140.

UV eprom eraser. Stag's *SE1T* ultraviolet eprom eraser is meant for low-volume work, although it takes up to five devices in up to 32-pin dips. Exposure time of up to an hour is mechanically timed, the device drawer having a safety interlock. Stag Programmers Ltd, 0707 332148.

COMPUTER

Computer board level products

400Mflops DSP. Using LSI's *MDC40T* DSP board, VME and PC cards can be configured with performances of 400Mflops and 2.2 thousand million operations per second (Gops). *MDC40T* uses three C40 comms channels to provide wideband interprocessor communication between two on-board processors; six comms ports remain available. Up to 4Mbyte of zero-wait-



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Bending the laws of resistance

Combine a linear taper with standard fixed-resistance devices and a non-linear potentiometer output can be the result. Richard Bolin and Philip Bullus show how.*

Variable resistor applications, including trimmer potentiometers and precision panel controls, often call for a non-linear output that is generally not available from standard products. Typical examples are voltage-level controls in power supplies, panel or internal adjustments on test or instrumentation equipment, and some audio controls.

Normally non-linear requirements are met by the expensive solution of specifying a customised device. But non-linear output can also be created in a more cost-effective way by using a standard variable-resistance device with a linear taper – ie with a constant rotation/resistance characteristic over the entire range of rotation – in conjunction with fixed resistors.

Figure 1 shows examples of ten circuits producing special tapers using conventional linear variable devices and fixed resistors. The tapers are plotted as rotation versus output voltage as a percentage of input voltage, using a combination of Ohm's law and parallel resistance calculations.

Four of the circuits use standard three-terminal linear potentiometers, either trimmer or panel types: one produces a linear characteristic with a minimum output voltage defined by the value of R_F (Fig. 1a); while two others use the loading effects of resistors R_L to produce nonlinear characteristics (Figs. 1b and 1c).

Degree of non-linearity is determined by the ratio R_L/R_T with very high values (in the limit, the equivalent of R_L being open) producing a linear characteristic, and lower values producing increasingly nonlinear characteristics. Curve concavity or convexity is determined by the configurations of the input and output terminals.

The fourth circuit option (Fig. 1d), uses a standard linear device and produces a characteristic which is non-linear near the ends of the rotation range but nearly linear in the middle portion. As in the preceding two circuits, the degree of non-linearity is determined by ratio R_L/R_T , with high values producing a larger linear portion and a moderate degree of non-linear

arity at each end.

The other six circuits use centre-tapped variable resistances. The first (Fig. 1e) produces a curve that resembles an inverted version of the one in Fig. 1d though neither the degree of non-linearity nor the slope of the linear portion can be varied.

The last five circuits produce various "dual-slope" characteristics. Some have linear portions, while others are totally non-linear. For example, in the circuit of Fig. 1g, the curve comprises two linear sections intersecting at the "knee" of the curve.

This knee will always occur at 50% rotation. But the slope of the first 50% of rotation – and therefore the percentage of maximum output voltage at which the knee occurs – is determined by the value of R_L . Relatively high values will produce a nearly-continuous, nearly-linear characteristic with only a slight knee; relatively low values will produce a shallower slope for the first 50% and will raise the knee. With proper values, the circuit approaches the logarithmic taper of an audio volume control.

To illustrate how different values of R_L affect the linearity, Fig. 2 shows a family of curves plotted for the circuit of Fig. 1h, assuming a value of $10k\Omega$ for R_T and using values of $5k\Omega$, $10k\Omega$ and $100k\Omega$ for R_L . Data points are taken at each 10% of rotation. As can be seen, the $100k\Omega$ value of R_L produces a nearly linear plot.

Before implementing circuit designs of this type, calculations must be made of the maximum current through both fixed and variable resistances, and power ratings for these components chosen accordingly.

Calculation of the loading effects on the output of the voltage source should also be made for a range of settings. ■

*Bourns Electronics.

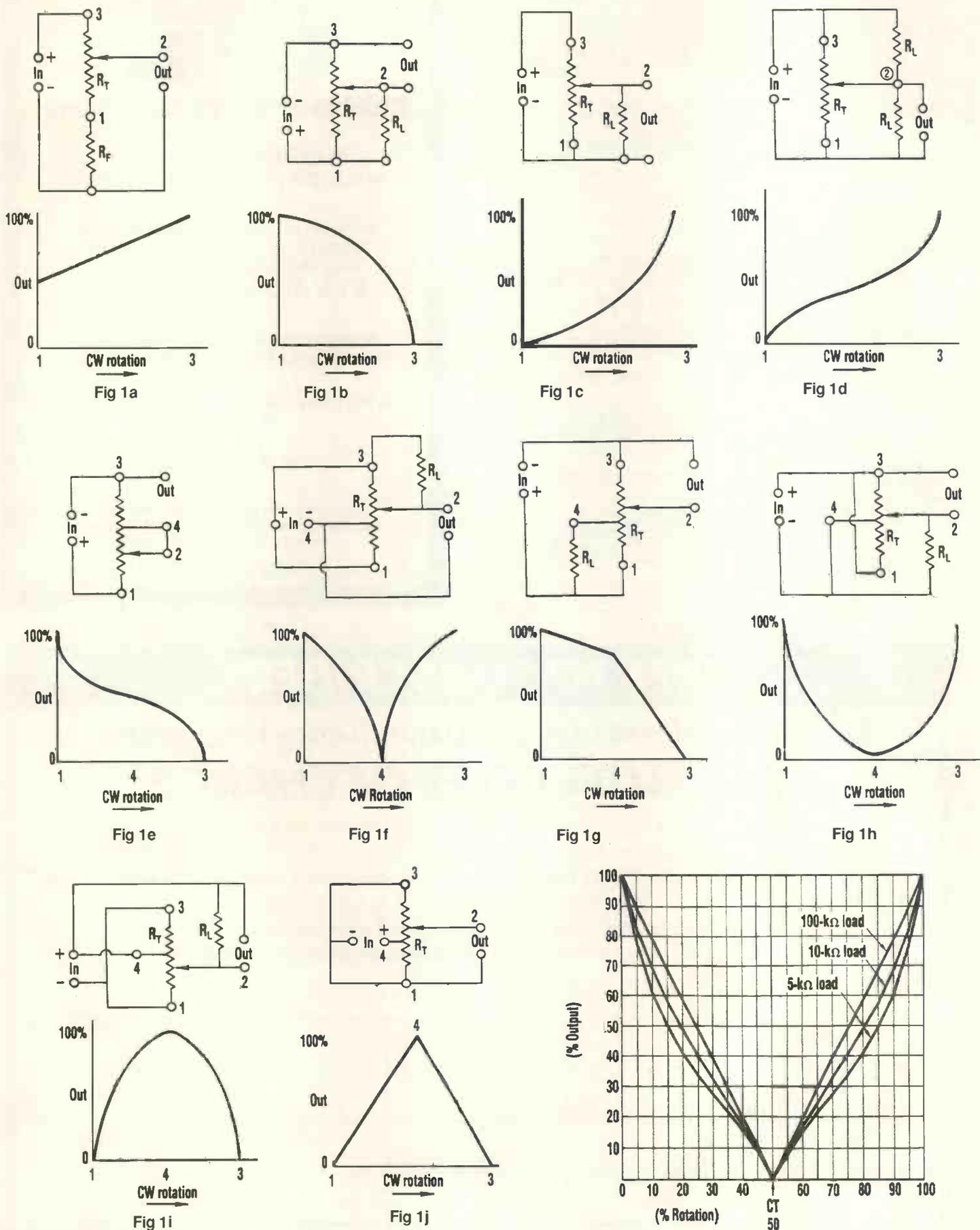


Fig. 1. Typical circuit designs producing non-linear outputs from standard variable and fixed resistances

Fig. 2. Effects of varying R_L on the output linearity

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BUILDING A BRIDGE

to more cost effective testing

Tying up powerful equipment on straightforward testing is an expensive luxury. Ian Hickman shows how the reflection coefficient bridge can form the basis for a serviceable scalar network analyser.

Oscilloscopes, signal generators, spectrum and network analysers are now immensely powerful – but unfortunately they are also very expensive. Often a much simpler piece of test equipment will suffice, such as when measuring whether a particular piece of equipment (such as an antenna) presents a good match to a 50Ω system. If the measurement only has to

be carried out at a fixed spot frequency, then a reflection coefficient bridge is a cheap solution. Since it can operate over a range of frequencies, the bridge can also be used to check performance over a wider bandwidth – though operation does become a little tedious. A suitably constructed reflection coefficient bridge (also known as an SWR Bridge) can operate up to microwave frequen-

GROUNDWORK

Some time ago I designed a simple reflection coefficient bridge using a diode detector. It worked, after a fashion, at low frequencies but became very insensitive at VHF. The transformer avoided capacitive coupling between the primary and secondary (which would have unbalanced the output) by the simple expedient of using diametrically opposite bunched windings on a ferrite ring core. The inevitable result was that at VHF, little of the drive signal reached the bridge due to excessive leakage inductance in the transformer. So an amplifier was added and provision also made to use an oscilloscope as an alternative detector. The old SL560 is an exceedingly useful device, and while not matching up to the performance of the latest current feedback op-amps in some respects, it still equals or surpasses them in others. Unfortunately even the addition of the amplifier did not rescue the performance at VHF.

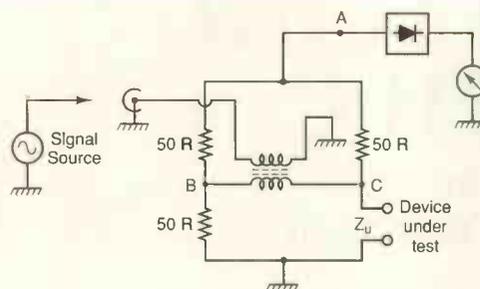
If the load under test Z_u is not a perfect 50Ω match, the reflected voltage is simply the difference between the voltage at point C and the voltage that would be there if Z_u were 50Ω (assuming the voltage at point A is constant). The voltage that would be there is equal to the voltage at B, so a detector monitoring the difference between these two points detects the reflected voltage.

Reflection coefficient ρ is simply the reflected voltage expressed as a fraction between 0 and 1, ie the voltage between C and B, divided by the voltage at B with

respect to ground. The voltage between C and B can be monitored with a diode detector which, if surrounded by resistors high enough to look like an open circuit compared to 50Ω, will not place any significant loading between these points or between either of them and ground.

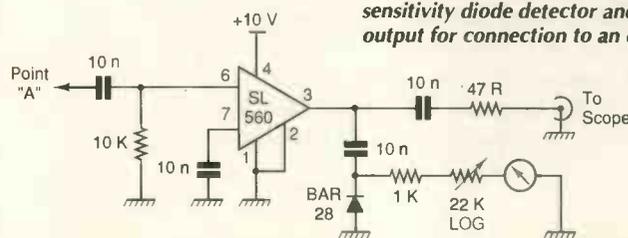
For example, if Z_u is a short circuit, the voltage at C with respect to B will be numerically equal to the voltage at B with respect to ground; a reflection coefficient of

unity. If Z_u is an open circuit, then V_{CB} (the voltage at C with respect to B) will be the same as V_{BC} in both magnitude and sign; a reflection coefficient of +1, as distinct from -1 in the short circuit case. Placing such a floating detector between points B and C means the bridge can be supplied by an unbalanced source driving point A, avoiding the floating bridge source transformer which was the problem with the earlier version.



Basic bridge circuit for use with 50Ω systems. The drive to the bridge is floating (balanced when the bridge is at balance), thanks to the transformer, enabling an unbalanced detector to be used.

(a)



A diode detector is not very sensitive and only large reflections can be detected. An amplifier is added, with a simple adjustable-sensitivity diode detector and a monitor output for connection to an oscilloscope.

cies – eg the Wiltron Model 60N50 with its 5-2000MHz range.

But design of a good radio frequency bridge transformer having low losses while still presenting a truly balanced floating drive to the bridge at all frequencies can be a particularly complicated operation.

After a lot of investigation (see box "Groundwork"), an SL560 became the basis for a circuit which, after a number of iterations, finished up looking like Fig. 1a. For any unknown impedance Z_u connected to the bridge, only voltage V_{CB} needs to be measured, divided by the measured value of V_{BG} , and the value of ρ is determined. Here, division by V_{BG} is implicit, since the measured V_{CB} is a normalised value as the drive voltage is adjusted so that V_{BG} gives full scale meter deflection.

Given ρ , the voltage standing wave ratio simply follows, as $VSWR = (1 + \rho)/(1 - \rho)$. Adjustments to the circuit under test which drive ρ toward zero will give the desirable result of a VSWR close to unity.

The original intention was to use one diode (a low power Schottky signal diode in view of the intended operating frequency of up to VHF) to monitor the voltage between points B and C, and another to monitor V_{BG} . But by using a subminiature slide switch S_1 placed immediately adjacent to the BNC socket accommodating the unknown impedance Z_u , the matched pair is not needed in the final scheme (Fig. 2b). Connecting the cathode of the diode either to point C or to ground G allows the same diode to do both tasks.

As the figure shows, a slight forward bias to improve sensitivity at low input levels is applied using a 10M Ω resistor. The rectified voltage is applied through a 1.5M Ω plus 10nF smoothing circuit to an op-amp circuit driving a 1mA meter with a scale calibrated 0-100.

Input bias current of the op amp used, a Texas Instruments *lincoms* TLC27M2, is so low (0.6pA) that there is no error due to volt drop across the smoothing circuit resistor R_6 . The TLC27M2 is a dual op amp, the other half being used to apply an offset voltage to the meter circuit, derived from R_8 , equal to the forward voltage established across the diode by the forward bias current through R_5 . As the detected voltage can go below ground and pin seven of the op amp has to sink up to 1mA, a negative supply rail is needed.

Since the diode bias current and the R_8 offset voltage are derived from a +10V rail, the familiar LM723 is pressed into service, Fig. 1c. Three leg regulators would have worked, but they don't come in a convenient voltage rating. The +10V chosen for the positive supply rail is about as much as the SL560 will take without getting too hot when its output emitter follower is biased up to drive a couple of volts, peak to peak, or more into the bridge. The latter nominally looks like 50 Ω , but with Z_u at short circuit, it is as low as 33 Ω . A pull-down resistor R_{11} is used to assist the op amp to sink the meter current, so the negative rail is not critical and is derived very simply from a couple of diodes – the ubiquitous 1N4148. This means the whole circuit can be run from a single 15V supply.

Operates well – at first

The circuit at first seems to perform very well. R_8 is set up for zero meter reading with no RF applied to the unit, and the output of a signal generator is connected to the input socket SK₁. Making Z_u a 50 Ω termination, the applied level is adjusted for full scale reading on the meter with S_1 in the SET position. Switching S_1 to MEAS gives a zero reading on the meter as is to be expected, so $\rho = 0$ and $VSWR = 1$. Replace the 50 Ω termination with a short circuit at Z_u (reset the drive at the SET position of

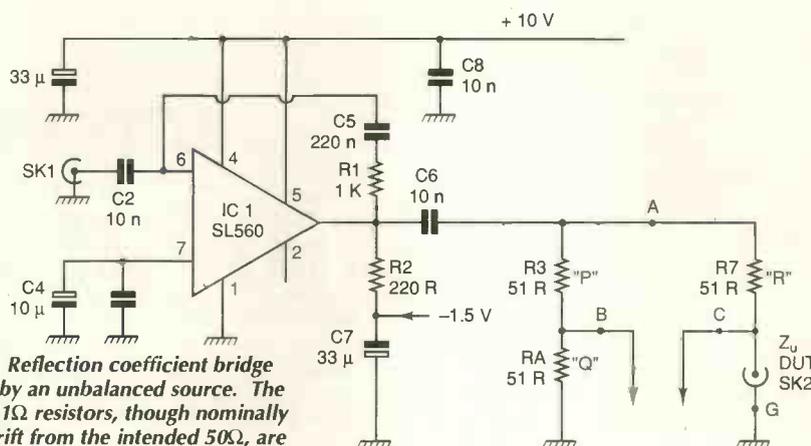
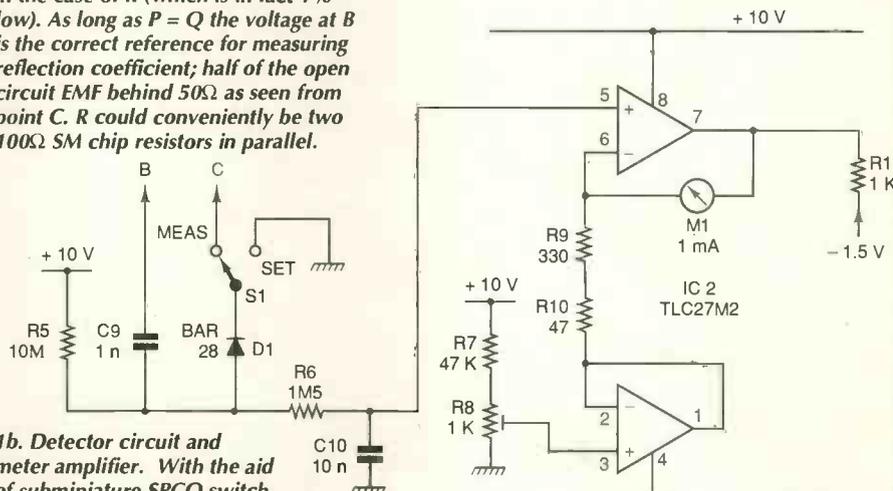
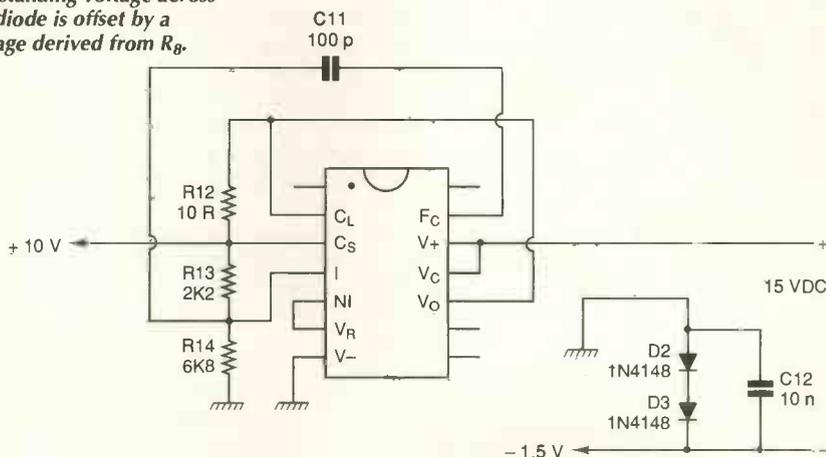


Fig. 1a. Reflection coefficient bridge driven by an unbalanced source. The three 51 Ω resistors, though nominally 2% adrift from the intended 50 Ω , are at least selected for close matching. The exact value is only of significance in the case of R (which is in fact 1% low). As long as $P = Q$ the voltage at B is the correct reference for measuring reflection coefficient; half of the open circuit EMF behind 50 Ω as seen from point C. R could conveniently be two 100 Ω SM chip resistors in parallel.



1b. Detector circuit and meter amplifier. With the aid of subminiature SPCO switch S_1 , the same diode can measure both V_{CB} and V_{BG} . The standing voltage across the diode is offset by a voltage derived from R_8 .



1c. An LM723 supplies +10V stabilised and -1.5V sort-of-stabilised to the circuit.

S_1 – necessary because the point A is not an ideal voltage source), and switch to MEAS to get a full-scale reading on the meter indicating $\rho = 1$, $VSWR = \text{infinity}$.

In fact this could not be otherwise, since with $Z_u = 0\Omega$, in both SET and MEAS positions of S_1 , the voltage at B is being measured with respect to ground. But when making Z_u open circuit, results are not quite so satisfactory with the meter reading somewhere between 5% and 15% over full scale, depending on frequency. Rechecking resistors "P" and "Q" and cross checking

Fig. 2. Showing (exaggerated) how second harmonic in the waveform applied to the diode detector circuit affects DC output. In the SET position of S_1 (Fig. 1b) D_1 DC restores on the positive peaks of the RF, so that the RF is negative-going with respect to ground, producing the detected DC output (left hand waveform). In the MEAS position, exactly the same occurs if Z_u is a short circuit – reflection coefficient $= -1$. If $\rho = +1$ (Z_u open circuit) then amplitude of the RF across D_1 is inverted, so that it DC restores with respect to the other peak of the RF (right hand waveform). As can be seen, if second harmonic is present on the drive waveform in the phase shown, the result is a larger detected output.

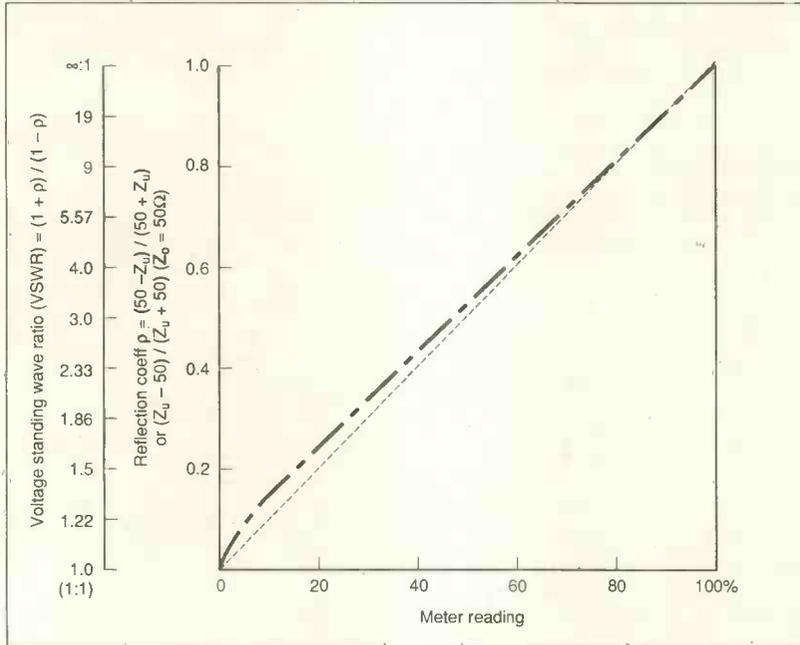
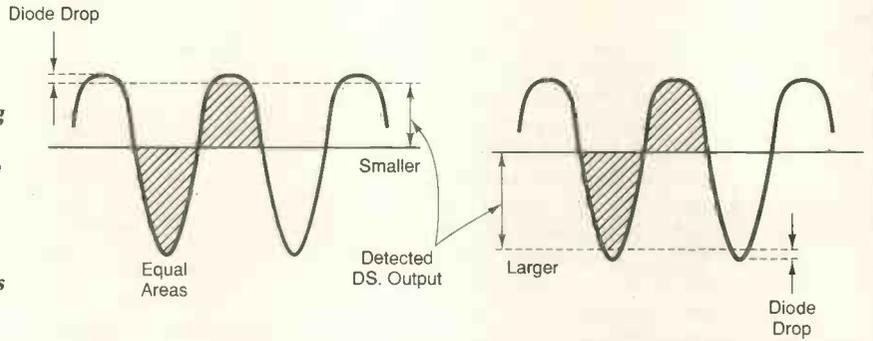


Fig. 3. Plot of reflection coefficient versus meter reading for the circuit of Fig. 1.

with an oscilloscope confirms that the voltage waveform at B is identical to that at A but with just 50% of the amplitude – as it should be.

But the scope reveals that the positive peaks of the waveform are slightly more rounded than the negative ones: not surprisingly with a single ended emitter-follower pushing 2V, peak to peak, into the bridge, some second harmonic is in evidence. To check this is the cause of the discrepancy, a 1:1 inverting transformer is interposed between the output of the SL560 and the bridge. The transformer uses an FX2754 two-hole balun core – a very useful part with its high AI value of 3500nH/turn.

Three turns of 26SWG bifilar wire give a primary inductance of 32µH – a primary reactance of 200Ω at 1MHz. Tests at 1, 3 and 10MHz with the waveform inverted show the reverse effect to that noticed previously with Z_u open circuit – ie the voltage at MEAS is now less than that at SET. So the circuit is modified by removing the short between pins 1 and 2 of the SL560 and connecting an external 220Ω resistor between its output pin, pin 3, and the -1.5V rail. The modifications lengthen the tail of the emitter-follower somewhat and incidentally remove some of the device's internal dissipation. Overall negative feedback is incorporated by adding R_f (DC blocked by C_5) from the output back to the inverting input. Just how much effect the NFB has depends on the output impedance of the signal generator used to supply the drive. Even with my Leader LSG-16, the results are improved so that up to 100MHz, the over-

reading with Z_u open circuit is reduced to 5% worst case and generally less. What more can you expect from such a simple piece of kit?

Low second harmonic

A low second-harmonic content in the signal driving the bridge is necessary because the sort of detector being used here works by DC restoring the signal negative-going with respect to the diode's cathode, and the RC filter R_6, C_{10} picking off the resulting DC level. When the DUT, device under test, terminals are open circuit, the voltage applied to the detector circuit in the MEAS position of S_1 is equal to that applied to it in the SET position; but it is inverted. So the reflected voltage is detected (DC restored) with respect to the other peak, giving rise to a difference in reading if there is second harmonic present (Fig. 2). The reason is that the phasing of the second harmonic relative to the fundamental is such as to sharpen one peak and flatten the other. Shift the phase of the second harmonic by 90° and it would not affect the reading, but the phase will always be as shown in Fig. 2 where the cause is a single ended emitter-follower stage groaning a little under the load.

But, a reactive Z_u will shift the phase of the second harmonic relative to the fundamental, adding further uncertainty to the measurement. There really is no substitute for carrying out measurements at just one frequency at a time – ie with a pure sine wave devoid of harmonics.

Useful performance

Regardless of the limitations noted, this simple piece of kit can prove very useful – provided Z_u is connected to the DUT input before adjusting the RF input level at the SET position of S_1 . Its usefulness is despite the fact that it does not measure actual resistive and reactive components of the load, but only the resultant reflection coefficient, and hence the VSWR.

Of course allowance must be made for the fall in diode sensitivity at low RF voltages, and to quantify this, the normalised reflected voltage is measured for various resistive values of Z_u between 50Ω and zero. Since the bridge measures the reflected voltage directly, the reflection coefficient should simply equal the meter reading, both expressed as percentages (shown by the dashed line in Fig. 3). The chain dotted line is a best fit to the dots, plotted points of meter reading for different reflection coefficients. The four highest points are obtained with 2.2, 4.7, 6.8 and 10Ω resistors – using measured DC actual values rather than nominal values and assuming these unchanged at the 1MHz test frequency used – and so on with various other resistors towards 50Ω. Figure 4a shows the effect of the parallel combination of a miniature 51Ω 2% resistor and a 33pF

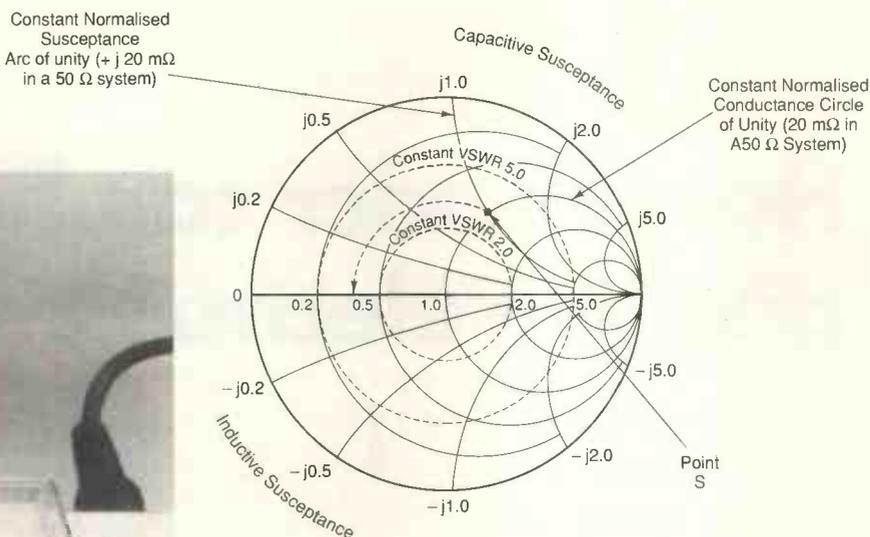
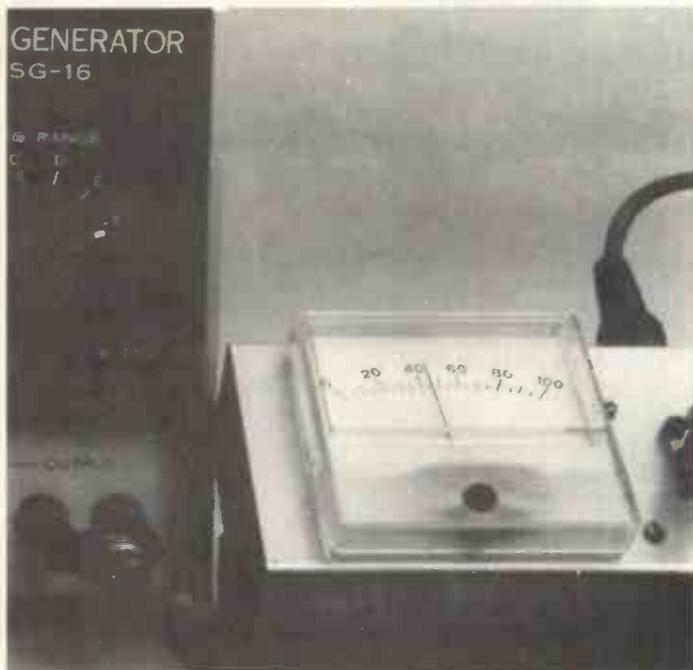


Fig. 4a. Normalised admittances of approximately unity (51Ω and $33pF$) connected in parallel to the bridge, driven at $100MHz$ from a signal generator. By a strange quirk of incompetence, the components are just out of view to the right. 4b. At the test frequency of $100MHz$, the "unknown" DUT Z_u corresponds to the point S on the Smith chart. Following this around to the horizontal diameter of the chart shows the reflection coefficient to be a little over 0.4 , (corresponding to a VSWR of $2.33:1$), compared to the indicated value from Fig. 3 of 0.46 .

2% capacitor at a frequency of $100MHz$, where the reactance of the capacitor is a little under 50Ω .

A BNC lead 3-4cm long is connected to the DUT socket, the two components being connected to the bare ends of the coax. The meter reading is 0.425 indicating a reflection coefficient of 0.46 (Fig. 3). Taking the combination of resistor and capacitor as conductance and susceptance values (since these, in parallel, add directly) corresponds to the point S on the Smith chart, Fig. 4b, near where the normalised conductance line unity crosses the normalised susceptance line of unity. Following round on a constant radius from the chart's centre to the horizontal diameter indicates a reflection coefficient of a shade over 0.4 .

It is a cheat to measure components whose value is already known, but it was done here to show that this simple equipment takes measurements that are at least approximately correct.

Modulus advantage

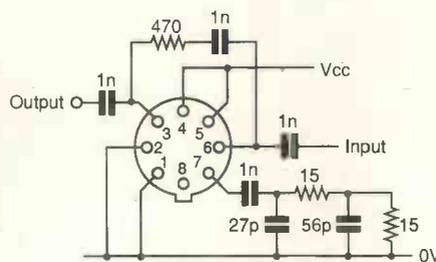
The answer given by the bridge is only the (modulus of) reflection coefficient – which in general is complex – and not the actual resistive and reactive components of Z_u , and this is both a drawback and an advantage. The benefit is that unlike a vector network analyser, a measurement plane does not need to be established. So the short length of coax in Fig. 4a is immaterial to the result. Indeed a longer length would also have no effect, provided only that its loss at the test frequency is negligible. Any loss in the cable will make the measured reflection coefficient smaller, as the reflected voltage is attenuated on its journey back to the bridge from Z_u , by the dissipation in the cable of the energy in the reflected wave.

In use, the reference voltage at B should be reset to full scale at the SET position of S_I each time the frequency is changed. This makes use of the bridge a little tedious when exploring the characteristics of Z_u over any substantial bandwidth. But it can be avoided by using a

signal generator with DC coupled external amplitude modulation, and detecting the voltage at point A (or B) and comparing it with a fixed reference voltage in a difference amplifier.

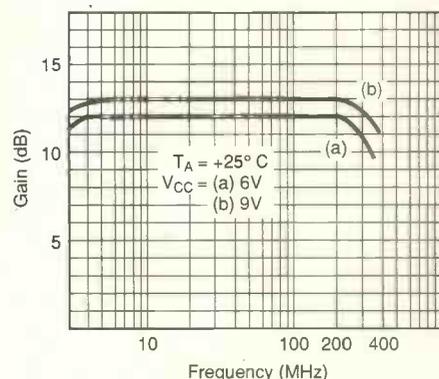
Output from the latter would feed into the sig gen's EXT AM input, controlling the loop to keep the voltage at A fixed. Point A becomes an ideal voltage source which means effectively zero internal source resistance, leaving the voltage at point B undisturbed by changes either in Z_u or frequency. The loop could be fast enough to maintain control as the sig gen is swept, making the whole arrangement into a rudimentary but serviceable scalar network analyser

Fig. 5. The SL560 can be used in a wide variety of ways, trading off between gain, bandwidth, noise figure, input impedance etc. The arrangement shown here provides a gain of $13dB$ with $-1dB$ bandwidth of $300MHz$. (GEC Plessey Semiconductors.)



Gain $13dB$ at $V_{CC} = 9V$
 $-1dB$ at $6MHz$ and $300MHz$

Wide bandwidth amplifier



Frequency response of circuit shown (Typical)

How Fourier can help analyse electronic waveforms

The Fourier series can be useful for analysing oscillators, amplifiers, and filters. In this extract from his book Understand Electrical and Electronics Maths, Owen Bishop explains that the calculations are not as hard as one may first imagine.

In audio and communications applications of electronics, it is often necessary to analyse a signal to find out what frequencies are present and in what proportions. The Fourier series is the key to a powerful analytical technique intended for this purpose. The idea behind the Fourier series is that any function may be thought of as the sum of one or more sine functions of different frequencies and different amplitudes.

Putting this in terms of audio signals, any periodic waveform is made up of one or more sine waves. But the Fourier series is more general than this, allowing mathematically defined functions, such as $y = 3x + 2$ and $y = 2x^3$, to be expressed in terms of sines. Purely random waveforms such as white noise, and the sounds made by many kinds of percussion instrument, cannot be analysed in this way.

When the string of an instrument such as a violin is plucked or bowed it vibrates and produces a sound. The ends of the string are stationary because they are firmly tightened against the body of the instrument at the bridge and by the finger of the player. But the centre of the string swings widely to and fro (Fig. 1a). When it is vibrating in this way the string produces its note of lowest pitch. It is the note of lowest frequency, called the fundamental or first harmonic, and referred to as f . The points where the string is stationary are known as nodes and the point where it is vibrating most widely is known as an antinode.

A string can vibrate in other ways. In Fig. 1b we show it vibrating with nodes at its ends and at its centre. There are two antinodes. The distance between the nodes is half that of Fig. 1a, and the frequency is double, or $2f$. Musicians say it is an octave above the fundamental. It has higher pitch and is called the second harmonic.

In Fig. 1c we see how the string produces its third harmonic, frequency $3f$, vibrating with nodes at its ends and at two evenly spaced locations along its length. There must always be nodes at its ends. In a similar way the string produces fourth, fifth, and even higher harmonics. At any given instant, the string is vibrating in all these ways at the same time producing its fundamental and several higher harmonics. The amplitude of vibration varies from one harmonic to another, with fundamental and lower harmonics usually being the loudest. Some of the harmonics may make the air in the body of the instrument resonate, making those harmonics louder. The result is a particular mix of harmonics which we recognise as the distinctive sound of the violin. As the player varies the length of the string by placing a finger at different positions, the fundamental note is changed and also the collection of harmonics which belong to it.

The same principles apply to producing sound by other musical instruments and human vocal cords. Electronic circuits too, particularly those of oscillators, amplifiers, and filters, may vibrate or resonate at set frequencies and their harmonics. When designing and testing audio and communications circuits, the ability to sort responses to the component waveforms of an electrical signal is very important and it is here that the Fourier series is so useful.

Sine waves

Sine waves, or sinusoidal waves as they are more correctly called, are described by the basic function $y = A \sin \omega t$ in which A is the amplitude, ω is the angular velocity, and t is the independent time variable.

The angular velocity in radians per second is directly related to the frequency f : $\omega = 2\pi f$. So the function is often written in the form $y = A \sin 2\pi f t$. The period of the function (the number of seconds taken for each cycle) is $1/f$. If we take the period to be 2π then $f = 1/2\pi$, and the function is simplified to $y = A \sin t$. All we need to remember is that, using this simplified version, the length of one cycle is 2π seconds.

In Fig. 2a the fundamental frequency is shown as a graph for one whole cycle. The equation for this is $y = A_1 \sin t$, using A_1 instead of plain A as we shall be needing several other amplitudes. In Fig. 2b the second harmonic, having double the frequency, goes through two cycles in the same time. The equation for the second harmonic is $y = A_2 \sin 2t$.

Doubling up the angle before we take the sine means that we fit in two cycles instead of one as t increases from 0 to 2π . A_2 does not necessarily have the same value as A_1 ; the two are related in frequency but may differ in amplitude.

Continuing with this, we find that the third harmonic (Fig. 2c) has the equation $y = A_3 \sin 3t$ and, in general, for the n th harmonic $y = A_n \sin nt$. These are the related harmonics which are

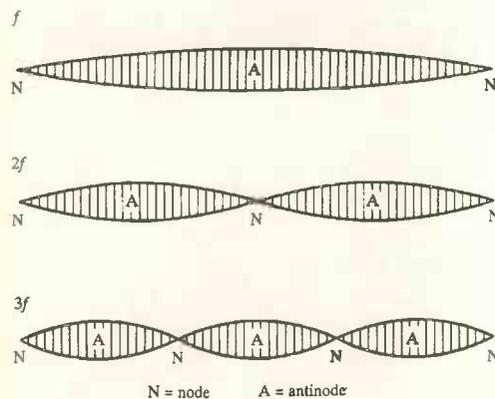


Fig. 1. Three examples of the way a string vibrates: The top shows the fundamental, the middle the second harmonic, and the bottom the third harmonic.

used to build up multifrequency periodic functions, such as those representing the sound of a clarinet. Fig. 2d shows a periodic function built up from the waveform of Figs. 2a and c.

Building the series

A function representing the combination of a fundamental and its harmonics can be written as the sum of the harmonics we have just described:

$$y = A_1 \sin t + A_2 \sin 2t + A_3 \sin 3t \dots + A_n \sin nt + \dots$$

Each of the component signals is present in different proportions according to its amplitude. The signals, although running at frequencies which are multiples of the fundamental, are not necessarily in phase with the fundamental or each other. We allow for this by giving each component a phase angle of its own. Now the series becomes:

$$y = A_1 \sin(t + \phi_1) + A_2 \sin(2t + \phi_2) \dots + A_n \sin(nt + \phi_n) + \dots$$

There is one further addition to the series, especially appropriate to electronic circuits. Very often an alternating signal does not alternate about 0V, but has a constant voltage (a DC voltage) added to or subtracted from it. We incorporate this DC voltage in the function as A_0 :

$$y = A_0 + A_1 \sin(t + \phi_1) \dots + A_n \sin(nt + \phi_n) + \dots$$

This series describes any periodic waveform. It might be wondered how it could possibly describe waves as sharp cornered as square waves and triangular waves, but we shall show later that this can be done. For the present we will take this series and convert it into a form that is easier to use.

Converting the sines

With the exception of the first term, the series consists of terms of the form $A_n \sin(nt + \phi_n)$. This can be expanded using the trig identity $\sin(A + B) \equiv \sin A \cos B + \cos A \sin B$. So:

$$A_n \sin(nt + \phi_n) = A_n (\sin nt \cos \phi_n + \cos nt \sin \phi_n)$$

Rearranging terms and the order of multiplying (which has no effect on the values) gives $(A_n \sin \phi_n) \cos nt + (A_n \cos \phi_n) \sin nt$. The terms in brackets have constant value because amplitude and phase angle are constant. To simplify the expression, replace these terms with constants $a_n = A_n \sin \phi_n$ and $b_n = A_n \cos \phi_n$. The terms of the series become $a_n \cos nt + b_n \sin nt$.

Now we are ready to write the series in its new form, substituting a new constant $a_0/2$ for A_0 , and listing the cosine terms first followed by the sine terms:

$$y = a_0/2 + a_1 \cos t + a_2 \cos 2t \dots + a_n \cos nt \dots + b_1 \sin t + b_2 \sin 2t \dots + b_n \sin nt \dots$$

This is the form in which the Fourier series is most often written. When analysing a waveform the usual aim is to discover the values of the constant coefficients $a_0, a_1, a_2, \dots, b_1, b_2, \dots$.

It may seem as if evaluating a Fourier series is a long and tedious matter, but this is usually not so. Often the series converges rapidly so that it reaches a value close enough to its limiting value after only a few terms. Also it may happen that the series does not contain any cosine terms (all the a coefficients are zero) or any sine terms. If so, the amount of calibration is halved immediately. In this and other ways to be explained later the evaluation of a Fourier series is often a relatively short calculation.

Dirichlet conditions

The Fourier series applies to a function with a period 2π , this being the time for exactly one cycle of the fundamental and for two or more complete cycles of the harmonics. In other words, the domain of the function is 2π . We may choose to calculate the function from any point during the cycle, provided that we take it up to a time exactly 2π seconds later. Since the fundamental repeats itself every 2π seconds, the choice of starting time makes no difference to the result. A suitable choice of starting time may make the calculations easier.

Usually it is best to cover the period 0 to 2π seconds. Occasionally it is easier if we begin at $-\pi$ and end at $+\pi$. Since the period is taken to be 2π , the frequency of the fundamental is 0.16Hz. This does not restrict Fourier analysis to signals at 0.16Hz. In a signal of any frequency the relationship between the fundamental and its harmonics is relative. Given any f we know that the harmonics are $2f, 3f, 4f$, and so on. The value of f itself is not important in the analysis and so we take the period as 2π to simplify the equations.

Not every function can be represented as a Fourier series. For a function to be analysed, it must conform to a number of conditions called the Dirichlet conditions. As long as these conditions apply, the function can be represented as a Fourier series.

The first condition is that the function must have a single defined value for each value in its domain. An equation such as $x^2 + y^2 = r^2$, which represents a circle, gives two values for y

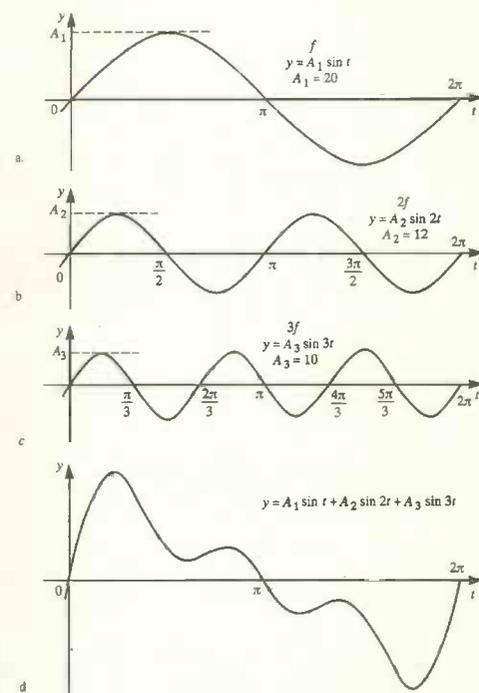


Fig. 2. Vibrations shown as graphs: a) the fundamental; b) second harmonic; c) third harmonic; and d) periodic function built up from a and c.

Definite integrals of sine and cosine functions

Some integrals reduce to either 0 or π when taken over a single cycle, greatly simplifying many of the Fourier calculations. Although the limits used below are 0 to 2π , the integrals have the same values over the limits $-\pi$ to $+\pi$ or any other interval of 2π .

Both n and m are positive integers, and integration is with respect to the time variable t . Here is a worked example

$$\int_0^{2\pi} \sin nt dt = \left(\frac{-\cos nt}{n} \right)_0^{2\pi} = \frac{1}{n} (-\cos 2n\pi + \cos 0) = \frac{1}{n} (-1 + 1) = 0$$

Similarly:

$$\int_0^{2\pi} \cos nt dt = 0$$

$$\int_0^{2\pi} \sin^2 nt dt = \pi \quad \text{given } n \neq 0$$

$$\int_0^{2\pi} \cos^2 nt dt = \pi \quad \text{given } n \neq 0$$

$$\int_0^{2\pi} \sin nt \cos mt = 0$$

$$\int_0^{2\pi} \cos nt \cos mt = 0 \quad \text{given } n \neq m$$

$$\text{or } = \pi \quad \text{given } n = m$$

$$\int_0^{2\pi} \sin nt \sin mt = 0 \quad \text{given } n \neq m$$

$$\text{or } = \pi \quad \text{given } n = m$$

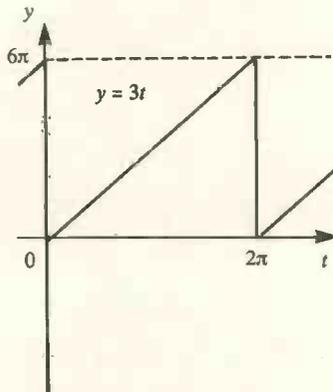


Fig. 3. Calculating the first term in a series using a sawtooth waveform.

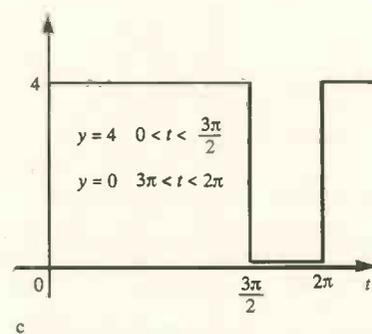
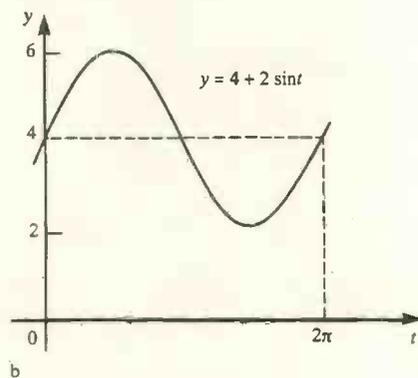
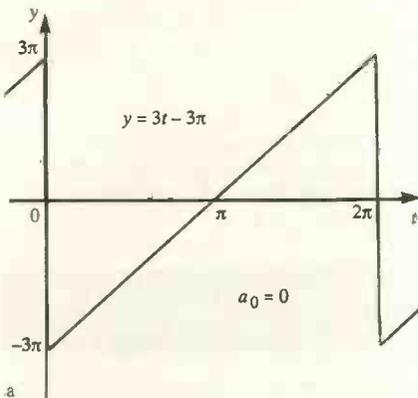


Fig. 4. Functions in which a_0 can be found: a) sawtooth waveform; b) sine wave; and c) square wave.

for each value of x , so it cannot be represented. It is usually not considered as a function.

Secondly, the function must not have any infinite discontinuity of range within its domain. For example, $y = \tan x$ does not fulfil this condition as it jumps between $-\infty$ and $+\infty$ when $x = 0, x = 2\pi, x = 4\pi$, and so on. Similarly, $y = 2/(x + 3)$ makes a similar jump when $x = -3$. But $2/(x - 7)$ has its discontinuity at $x = 7$, which is outside the domain 0 to 2π (x is in radians) and fulfils the condition.

The final condition says that the first and second differentials of the function must be piecewise continuous over the domain.

Most examples taken from electronic circuits conform to these conditions, so this does not normally cause difficulties. If the conditions are all fulfilled, the series is convergent, usually rapidly, and can be evaluated with sufficient precision by taking only the first few terms.

Finding the first term

The first term a_0 is found by integrating the series from 0 to 2π . We will not explain why but will show that it works. We begin by rewriting the series as the first term plus the sum of the cosine and sine terms:

$$y = \frac{a_0}{2} + \sum_{n=1}^{\infty} (a_n \cos nt + b_n \sin nt)$$

in which n is a positive integer. Integrating both sides of this equation:

$$\int_0^{2\pi} y dt = \frac{1}{2} \int_0^{2\pi} a_0 dt + \sum_{n=1}^{\infty} \left(\int_0^{2\pi} a_n \cos nt dt + \int_0^{2\pi} b_n \sin nt dt \right)$$

But $\cos nt$ and $\sin nt$ integrated from 0 to 2π with respect to t are both zero, so:

$$\int_0^{2\pi} y dt = \frac{1}{2} (a_0 t)_0^{2\pi} + 0 = \frac{1}{2} (a_0 2\pi - a_0 0) = a_0 \pi$$

$$\Rightarrow a_0 = \frac{1}{\pi} \int_0^{2\pi} y dt$$

This is the equation used in calculating the first term in the series, the first term being $a_0/2$.

Figure 3 illustrates this calculation with reference to a sawtooth waveform in which a simple ramp function is repeated with period 2π . The function is $y = 3t$ with $0 < t < 2\pi$.

$$a_0 = \frac{1}{\pi} \int_0^{2\pi} 3t dt = \frac{1}{\pi} \left(\frac{3t^2}{2} \right)_0^{2\pi} = \frac{1}{\pi} \left(\frac{3(2\pi)^2}{2} - 0 \right)$$

$$\Rightarrow a_0 = 6\pi$$

From this we obtain the first term of the series $a_0/2 = 3\pi$.

Figure 3 shows that y increases steadily from 0 to 6π during one cycle. Its mean value is 3π , which is the same as that which we have just calculated for the first term. This illustrates the rule that: First term = mean value of y over one cycle.

This rule often makes calculations very much easier. If it is possible to find the mean value of y by drawing a sketch of the function and applying elementary geometry, we can avoid the integration. Figure 4 shows some examples of functions in which a_0 and hence $a_0/2$ can be found by inspecting the sketch. In Fig. 4a the function has the same waveform as that of Fig. 3, but the mean value of y is zero. In terms of voltages, the signal of Fig. 3 can be thought of as the signal of Fig. 4a superimposed on a DC level of $3\pi V$. The first term of the Fourier series represents any DC voltage (or steady current or other physical quantity) that may be present.

Figure 4b shows the function $y = 4 + 2\sin t$ with $0 < t < 2\pi$. Integrating from 0 to 2π :

$$a_0 = \frac{1}{\pi} \int_0^{2\pi} (4 + 2\sin t) dt = \frac{1}{\pi} (4t - 2\cos t)_0^{2\pi}$$

But $\cos 2\pi = \cos 0 = 1$:

$$a_0 = \frac{1}{\pi} [(8\pi - 2) - (0\pi - 2)] = \frac{8\pi}{\pi} = 8$$

From this, the first term in the series is $a_0/2 = 4$. The symmetry of the curve shows that the mean value of y between 0 and 2π is 4 . This could have been obtained without calculation, merely by inspecting the graph, but we calculated it to illustrate the principle involved.

The piecewise function of Fig. 4c is a pulsed waveform. Its value is 4 for three quarters of the time and zero for the remainder. Without further calculation, we can say that $a_0 = 3$ and that the first term is $a_0/2 = 1.5$.

Finding the cosine terms

These are the terms that have $a_1, a_2, \dots, a_n, \dots$ as their coefficients. The equation for a_n is:

$$a_n = \frac{1}{\pi} \int_0^{2\pi} y \cos nt dt$$

As an example, we return to the waveform of Fig. 3, for which $y = 3t$:

$$a_n = \frac{1}{\pi} \int_0^{2\pi} 3t \cos nt dt$$

This is integrated by parts. let $u = 3t$, then $du/dt = 3$. Let $dv/dt = \cos nt$, then $v = (\sin nt)/n$:

$$a_n = \frac{1}{\pi} \left[\left(\frac{3t}{n} \sin nt \right) - \frac{1}{n} \int_0^{2\pi} 3 \sin nt dt \right]$$

But the integral on the right is zero (see box), so $a_n = (6\pi \sin 2n\pi - 6\pi \sin 0)/\pi n$. The value of $\sin 0, \sin 2\pi, \sin 4\pi, \sin 6\pi$, and so on are all zero. So a_n is zero, and all the other a coefficients are zero and therefore there are no cosine terms in the series for this function.

Finding the sine terms

The routine is similar to that for finding the cosine term except that the equation is:

$$b_n = \frac{1}{\pi} \int_0^{2\pi} y \sin nt dt$$

Continuing the example of Fig. 3:

$$b_n = \frac{1}{\pi} \int_0^{2\pi} 3t \sin nt dt$$

Integrating by parts gives:

$$b_n = \frac{1}{\pi} \left[(-3t \cos nt) \Big|_0^{2\pi} + \frac{1}{n} \int_0^{2\pi} 3 \cos nt dt \right]$$

The integral on the right equals zero (see box):

$$b_n = \frac{1}{\pi n} (-6\pi \cos 2\pi + 0\pi \cos 0)$$

The cosine of even multiples of π is 1, so $b_n = -6/n$. Thus the general form of sine terms is $(-6/n) \sin nt$.

The final analysis

We have established that the Fourier series for $y = 3t$ has the following features: $a_0/2 = 3\pi$; there are no cosine terms; and the sine terms have the form $(-6/n) \sin nt$.

Using these results, we write out the series for as many terms as we need:

$$y = 3\pi - (6/1) \sin 1t - (6/2) \sin 2t - (6/3) \sin 3t \dots \text{ This is more simple written: } y = 3\pi - 6 \sin t - 3 \sin 2t - 2 \sin 3t \dots$$

Now to put it to the test. Normally we would accept the analysis as correct, assuming that all the calculations have been worked properly. As this is the first example that we have worked, we will see what happens when this function is plotted as a graph. In Fig. 5, the graph shows the function calculated as far as the fifth sine term, the fifth harmonic. A short computer program was used to calculate the points. Although the waveforms of the harmonics show up on the graph as slight undulations, the overall shape of the curve is the same as that of Fig. 3.

This analysis is sufficiently precise for many purposes but we can extend the calculation to include more terms if required, and obtain an even better approximation to Fig. 3. In Fig. 6 the function is calculated to the 20th harmonic. There are more ripples but they are much smaller. This is a good demonstration of the way in which a sharply pointed function such as the sawtooth wave can be analysed in a series of curvy sine waves.

Fourier coefficients

In some applications we do not need to know the equation for the whole series but only the coefficient of one of the harmonics. Ignoring the sign, this gives the amplitude of the harmonic. It is calculated using the technique already explained. For example, if we need to know the amplitude of the third harmonic of $y = 3t$, we calculate it from $b_n = -6/n$, ignoring the sign. The amplitude of the third harmonic is $6/3 = 2$. More important is the amplitude of the harmonic relative to that of the fundamental.

This is usually expressed as the percentage harmonic. In this example the first harmonic has amplitude 6, so the percentage third harmonic is:

$$\frac{\text{amplitude of third harmonic}}{\text{amplitude of fundamental}} \times 100 = \frac{2}{6} \times 100 = 33.3\% \text{ (1dp)}$$

A graph of the amplitudes of the harmonics is known as the frequency spectrum of a function. Figure 7 shows the frequency spectrum of the waveform of Fig. 3. Compared with spectrums of other functions that we shall look at, this series converges slowly. The coefficient $6/n$ falls more and more slowly as n increases. This accounts for the fact that we need about 20 terms (Fig. 6) to make it approximate reasonably to the sawtooth waveform.

In this example the series comprised only the initial (DC) term and sine terms. In series which have cosine and sine terms it is necessary to sum the corresponding sine and cosine terms for a given harmonic: $a_n \cos nt + b_n \sin nt$.

Multiples of π

If n is an integer, positive or negative and including $n = 0$, then $\sin n\pi = 0$, and $\cos n\pi = 1$ if n is even or -1 if n is odd.

Fourier coefficients

$a_0 = \frac{1}{\pi} \int_0^{2\pi} y dt$; First term $= \frac{a_0}{2}$

Cosine terms: $a_n = \frac{1}{\pi} \int_0^{2\pi} y \cos nt dt$

Sine terms: $b_n = \frac{1}{\pi} \int_0^{2\pi} y \sin nt dt$

Limits of integrals can also be $-\pi$ to $+\pi$.

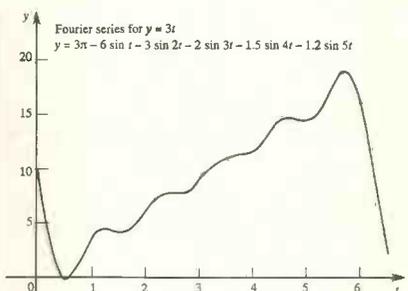


Fig. 5. Function calculated as far as the fifth sine term, fifth harmonic.

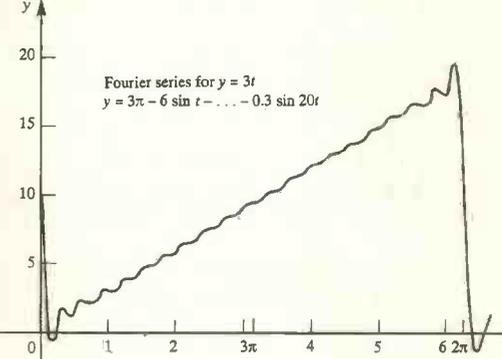
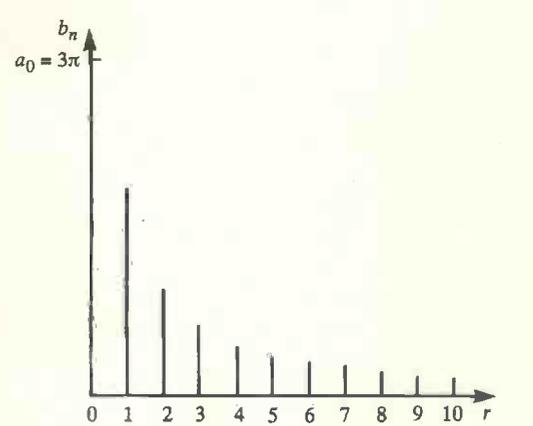


Fig. 6. Same function as in Fig. 5, but calculated to the 20th harmonic.



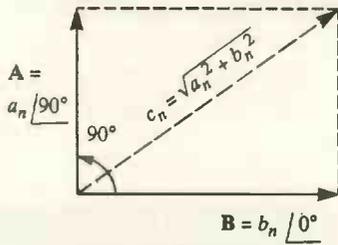


Fig. 8. Phasor representation of cosine and sine terms.

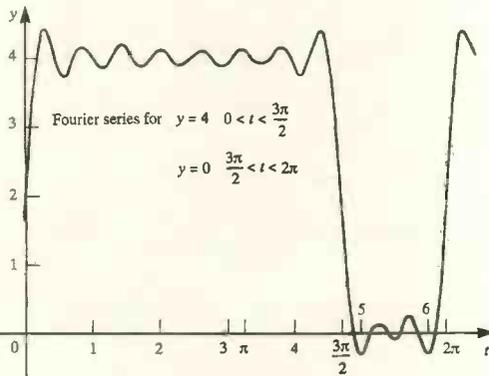


Fig. 9. Plot of the Fourier series for the pulse waveform of Fig. 2c, taken to the 12th harmonic.

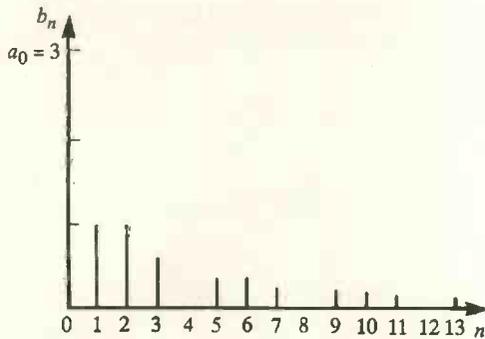


Fig. 10. Frequency spectrum showing the coefficients of the first 13 harmonics.

We cannot simply add a_n to b_n . This is because the two components, being a cosine and sine, are not in phase with each other. But the cosine of an angle equals the sine of an angle plus 90° . Thus the n th harmonic can also be written: $a_n \sin(nt + 90^\circ) + b_n \sin nt$.

In Fig. 8 these two terms are represented by phasors, with the a_n phasor leading the b_n phasor by 90° . From Pythagoras' theorem, the resultant phasor has magnitude:

$$c_n = \sqrt{a_n^2 + b_n^2}$$

An obvious example

The curve of Fig. 4b consists only of a DC component plus the fundamental, as can be seen by glancing at its equation: $y = 4 + 2\sin t$ with $0 < t < 2\pi$.

It is not necessary to use Fourier analysis on such a trivial function, but we will do so to run through the routine and to show how the definite integrals in the box are useful.

It has already been shown that the first term is $a_0/2 = 4$. The formula of a_n gives:

$$\begin{aligned} a_n &= \frac{1}{\pi} \int_0^{2\pi} (4 + 2\sin t) \cos nt \, dt \\ &= \frac{1}{\pi} \left(\int_0^{2\pi} 4 \cos nt \, dt + \frac{1}{\pi} \int_0^{2\pi} 2 \sin t \cos nt \, dt \right) \end{aligned}$$

The equations in the box show that the first integral is zero. The second, in which $m = 1$, is also zero. Thus $a_n = 0$ and there are no cosine terms. The formula for b_n gives:

$$\begin{aligned} b_n &= \frac{1}{\pi} \int_0^{2\pi} (4 + 2\sin t) \sin nt \, dt \\ &= \frac{1}{\pi} \left(\int_0^{2\pi} 4 \sin nt \, dt + \frac{1}{\pi} \int_0^{2\pi} 2 \sin t \cos nt \, dt \right) \end{aligned}$$

The first integral is zero. The value of the second integral depends on the value of n (see box), except that n and m are interchanged). Here $m = 1$, and if $n \neq m$, then the integral has zero value. However, if $n = 1$ the integral has the value π , and $b_1 = 2$.

The series has only one sine term, when $n = 1$. From these calculations we write out the series $y = a_0/2 + b_1 \sin t = 4 + 2\sin t$. This is an exact solution and one that we know already.

A less obvious example

The pulse waveform of Fig. 4c is one of the more complicated functions and illustrates some more of the paths an analysis may take. This is a piecewise function and $y = 4$ with $0 < t < (3\pi/2)$ and $y = 0$ with $(3\pi/2) < t < 2\pi$. It has already been shown that $a_0/2 = 3$.

Integrating for a_n is done separately for each part of the function; the essential point is that the whole period 0 to 2π must be covered:

$$a_n = \frac{1}{\pi} \int_0^{3\pi/2} 4 \cos nt \, dt + \frac{1}{\pi} \int_{3\pi/2}^{2\pi} 0 \cos nt \, dt$$

In this example, the function has zero value during the later part of the phase, so the integral is zero, and we evaluate only the first integral. In other examples, where the function has non-zero value at all stages, we evaluate the corresponding integrals and sum the results.

Continuing with the first integral:

$$a_n = \frac{4}{\pi} \int_0^{3\pi/2} \cos nt \, dt = \frac{4}{\pi n} (\sin nt)_0^{3\pi/2}$$

Note that the equations in the box are of no use in this example as we are integrating from 0 to $3\pi/2$, not from 0 to 2π .

$$a_n = \frac{4}{\pi n} \left(\sin \frac{3n\pi}{2} - \sin 0 \right)$$

There are cosine terms in this series. Ignoring $\sin 0$, which has zero value, the a coefficients are $(4/\pi n) \sin(3n\pi/2)$.

Calculating the value of $\sin(3n\pi/2)$ for $n = 1, 2, 3, \dots$ shows that it cycles through four values: 0, -1, 0, +1, ... This means that there are no odd cosine terms and that the signs of the even terms are alternately negative and positive:

$$\frac{4}{\pi} \left(-\cos t + \frac{\cos 3t}{3} - \frac{\cos 5t}{5} + \frac{\cos 7t}{7} - \dots \right)$$

A similar calculation shows that:

$$b_n = \frac{4}{\pi n} \left(\cos \frac{3n\pi}{2} - \cos 0 \right)$$

in which $\cos 0 = 1$. Calculating b for $n = 1, 2, 3, \dots$ shows that it cycles through four values: 1, 2, 1, 0, ... Terms are all positive but every fourth term is missing, starting with $n = 4$:

$$\frac{4}{\pi} \left(\sin t + \sin 2t + \frac{\sin 3t}{3} + \frac{\sin 5t}{5} + \frac{\sin 6t}{3} \right)$$

Note that the effect of the factor 2 when $n = 2, 5, 10$, and so on. Combining the results of these calculations, the Fourier series is:

$$y = 3 + \frac{4}{\pi} \left(\begin{array}{l} -\cos t + \frac{\cos 3t}{3} - \frac{\cos 5t}{5} + \frac{\cos 7t}{7} - \frac{\cos 9t}{9} + \dots \\ + \sin t + \sin 2t + \frac{\sin 3t}{3} + \frac{\sin 5t}{5} + \frac{\sin 7t}{7} + \dots \end{array} \right)$$

A plot of this (Fig. 9) taken to $n = 12$ shows that it is very close to the original function.

Calculating coefficients

The example above is a series in which most of the harmonics have cosine and sine components. To calculate the amplitude of these harmonics we use the formula given earlier. For example, the seventh harmonic is:

$$\frac{4}{\pi} \left(\frac{\cos 7t}{7} + \frac{\sin 7t}{7} \right) \text{ Coefficients are: } a_7 = b_7 = \frac{4}{7\pi}$$

Applying the formula:

$$c_7 = \sqrt{a_7^2 + b_7^2} = \frac{4}{\pi} \sqrt{\frac{1}{7^2} + \frac{1}{7^2}} = 0.26$$

The coefficients for the first 13 harmonics are shown in the frequency spectrum of Fig. 10. After the third harmonic the series converges fairly rapidly. The 4th, 8th, 12th, ... harmonics are absent.

Symmetry

As we have seen, whole groups of terms are often completely absent from a series. There may be no cosine terms, as in Fig. 11a, or there may be no sine terms, as in Fig. 11b. In other series there may be only terms for which n is odd, as in Fig. 11d.

It can be shown that waveforms of certain types always have the same groups of terms missing from their series. If we know what type the waveform belongs to, we know what groups of terms are absent and so avoid wasting time by trying to calculate them.

The rules for identifying these kinds of waveforms depend on the symmetry of the curve.

In one type of symmetry, the waveform of the first half of the cycle is repeated exactly during the second half, but is inverted. One of the simplest examples is the sine curve (Fig. 12). Other examples are shown in Figs. 4a, 11a, and 11d. This type of symmetry is known as half wave inversion.

One feature is that the mean value of y is zero, so there is no initial term in the series.

Another feature is that the series has no terms for the even values of n .

Figure 4b does not show half wave inversion as both halves are above the t axis; it has a DC component of 4. But if this is removed so that the curve becomes symmetrical about the t axis, the curve then has half wave inversion. Its series has no even terms. Note that shifting a curve to left or right has no effect on its half wave inversion.

The wave of Fig. 2d has a different type of symmetry in that the curve of the second half of the cycle, if rotated about the point $(\pi, 0)$, lies exactly on the curve of the first half. However, this is not half wave inversion; note that this curve has an even term in its series. We will now go on to look at terms of this type.

The two important types of symmetry are shown by odd and even functions. An odd function is symmetrical about the origin. In other words, if the curve is rotated 180° about the origin, it lies on itself. A sine curve is a typical example. Examples shown are in Figs. 2d, 4a, 11a, and 12. An odd function has sine terms only. Also, since the curve must inevitably be symmetrical about the t axis, curves of this type have no initial term.

An even function is one which is symmetrical about the y axis. In other words, the portion to the left of the y axis is the mirror image of the portion to the right. The cosine curve ($y = \cos t$, Fig. 13) is a typical example. Another is shown in Fig. 11b. As might be expected, the series for such curves includes cosine terms but no sine terms. In other words, a_n exists but $b_n = 0$. The initial term a_0 may be absent or present, depending on whether the function is symmetrical about the t axis (Fig. 13) or not (Fig. 4b).

Figure 14 shows a curve which is not symmetrical (so it is not properly included under this heading) but which has the feature known as half wave repetition. This function repeats with a period of π instead of 2π . Functions of this kind may have cosine and sine terms, but only those for even values of n .

Functions may show more than one type of symmetry. The effects of symmetry in eliminating terms are cumulative. For example, Fig. 11a has odd symmetry and half wave inversion. Because of its odd symmetry it has no cosine terms. Because of its half wave inversion it has no DC term and no even terms. The series consists of only odd sine terms.

Rectified AC

Alternating current, as supplied from a generator, consists of a more or less pure sine wave of a single frequency, the fundamental. After it has been rectified by a diode bridge, the

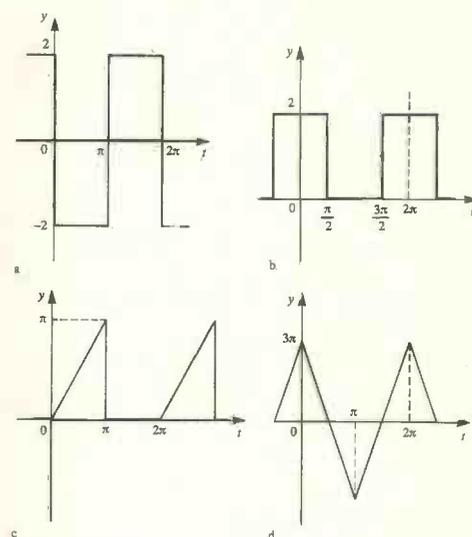


Fig. 11. Various examples of symmetry: a: no cosine terms; b: no sine terms; and d: n is odd.

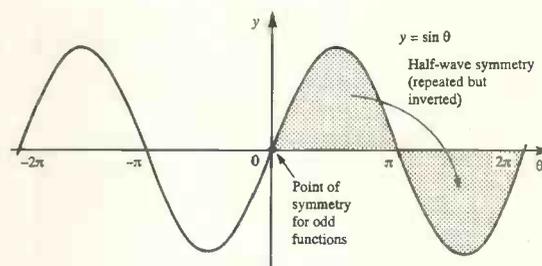


Fig. 12. Typical example of a sine curve.

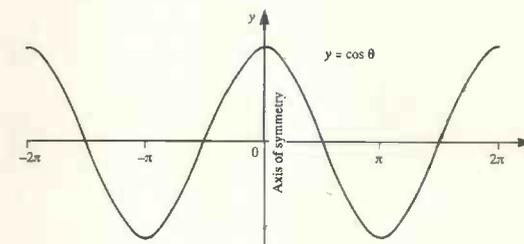


Fig. 13. Typical example of a cosine curve.

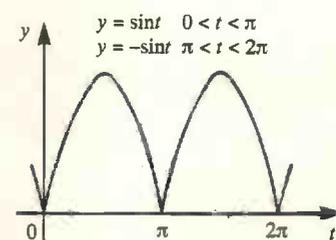


Fig. 14. Example of half wave repetition curve.

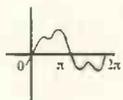
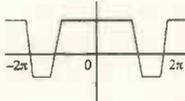
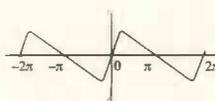
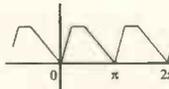
Symmetry		a_0	a_n (cosine terms)	b_n (sine terms)
Type	Example			
Half-wave inversion		No	No even terms	No even terms
Even		-	-	None
Odd		No	None	-
Half-wave repetition		-	No odd terms	No odd terms

Fig. 15. Typical range of symmetrical curves.

Owen Bishop's book "Understand Electrical and Electronics Maths" is aimed at people who find maths difficult, making it hard to grasp electrical and electronics theory. This 256 page book is available by postal application to Lorraine Spindler, EW + WW, Quadrant House, The Quadrant, Sutton, Surrey SM2 5AS. Cost £15.95 including postage and packing. Cheques made payable to Reed Books Services. Credit card orders accepted by phone (081-652 3614). Published by Butterworth-Heinemann.

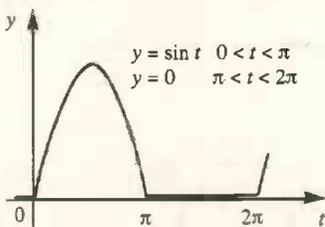


Fig. 16. Series for an AC wave that has been half wave rectified

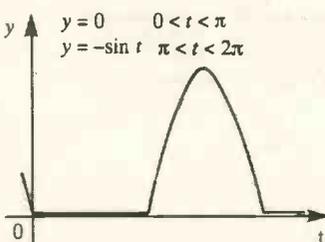


Fig. 17. The series for the second half of the cycle.

waveform is as shown in Fig. 14. The piecewise function is:

$$y = \sin t \text{ with } 0 < t < \pi \text{ and } y = -\sin t \text{ with } \pi < t < 2\pi.$$

Examination of the graph shows that this has half wave repetition and even symmetry. The series therefore has no odd terms and no sine terms, consisting only of a DC term and even cosine terms. Rectification has produced a drastic change in the original pure sine wave.

It is interesting to analyse the two halves of the cycle separately, especially since this gives us the series for an AC wave which has been half wave rectified (Fig. 16). This wave has no symmetry so we shall need to calculate a_0 , a_n , and b_n .

$$a_0 = \frac{1}{\pi} \int_0^\pi \sin t dt = \frac{1}{\pi} (-\cos t)_0^\pi = \frac{2}{\pi}$$

the initial term is: $\frac{a_0}{2} = \frac{1}{\pi}$

$$a_n = \frac{1}{\pi} \int_0^\pi \sin t \cos nt dt$$

In this example, the usual technique of integrating by parts does not work, for it simply produces another integrand containing sine and cosine terms. Instead we rely on the trig identity $2\sin A \cos B \equiv \sin(A+B) + \sin(A-B)$. In this case, put $A = t$ and $B = nt$, and split the integral in two:

$$\begin{aligned} a_n &= \frac{1}{2\pi} \left[\int_0^\pi \sin(1+n)t dt + \int_0^\pi \sin(1-n)t dt \right] \\ &= \frac{-1}{2\pi} \left[\frac{\cos(1+n)t}{1+n} + \frac{\cos(1-n)t}{1-n} \right]_0^\pi \end{aligned}$$

The negative sign from integrating the sines is outside the bracket. Adding the fractions:

$$= \frac{-1}{2\pi} \left[\frac{(1-n)\cos(1+n)t + (1+n)\cos(1-n)t}{(1+n)(1-n)} \right]_0^\pi$$

Multiplying out and then rearranging the terms:

$$= \frac{-1}{2\pi} \left\{ \frac{\cos(1+n)t + \cos(1-n)t - n[\cos(1+n)t - \cos(1-n)t]}{1-n^2} \right\}_0^\pi$$

Now we use two more trig identities to replace the first pair of terms and the last pair of terms: $2\cos A \cos B \equiv \cos(A+B) + \cos(A-B)$; and $2\sin A \sin B \equiv \cos(A+B) - \cos(A-B)$.

$$a_n = \frac{-1}{\pi} \left(\frac{\cos n\pi + 1}{1-n^2} \right)$$

As the sine terms all equal zero:

$$a_n = \frac{-(-\cos n\pi - 1)}{\pi(1-n^2)} = \frac{\cos n\pi + 1}{\pi(1-n^2)}$$

If n is even, then $a_n = 2/\pi(1-n^2)$, and if n is odd $a_n = 0$. However, there is one odd number not yet accounted for. When $n = 1$, the denominator becomes zero, so the expression is indeterminate. We must evaluate this separately:

$$a_1 = \frac{1}{\pi} \int_0^\pi \sin t \cos t dt = \frac{1}{2\pi} \int_0^\pi \sin 2t dt = 0$$

The calculation of b_n is similar to those above, with the result $b_n = 0$. The general expression is indeterminate for $n = 1$. Calculation shows that $b_1 = 1/2$. From this we obtain the series:

$$y = \frac{1}{\pi} \left(1 - \frac{2}{3} \cos 2t - \frac{2}{15} \cos 4t - \frac{2}{35} \cos 6t - \dots + \frac{\pi}{2} \sin t \right)$$

The cosine terms are negative because $(1-n^2)$ is negative when $n > 1$. The series for the second half of the cycle (Fig. 17) gives an almost identical result:

$$y = \frac{1}{\pi} \left(1 - \frac{2}{3} \cos 2t - \frac{2}{15} \cos 4t - \frac{2}{35} \cos 6t - \dots - \frac{\pi}{2} \sin t \right)$$

The only difference is the sign of the last term. When these two series are added to obtain the series for the full wave rectified wave form (Fig. 14), the two sine terms cancel each other, and the other terms are doubled:

$$y = \frac{2}{\pi} \left(1 - \frac{2}{3} \cos 2t - \frac{2}{15} \cos 4t - \frac{2}{35} \cos 6t - \dots \right)$$

This consists of $a_0/2$ and the even cosine terms, as predicted at the start of this discussion. This is a series which converges rapidly, because of n^2 . ■

Prototype technology put to the test

Ferro-electric liquid crystals or FLCs could form the basis for the next generation of computer displays, and eventually may be used for HDTV systems and camcorder electronic viewfinders, if Canon has its way.

FLC displays offer several advantages over conventional LCDs, including higher contrast ratios, wider viewing angles, faster response times and larger screen sizes.

They are made by blending fluorine-containing chiral liquid crystal compounds with low viscosity LC compounds. FLC molecules are long, thin, rod-shaped structures sandwiched between a glass substrate to form a cell. An electric current applied to the cell alters the alignment of the FLCs: a positive voltage allows light to pass through the cell, while a negative voltage blocks it.

The spacing of the glass plates must not exceed $2\mu\text{m}$, but Canon has managed to produce a 380mm FLC display with a uniform cell thickness of just $1.5\mu\text{m}$, $\pm 0.05\mu\text{m}$. Short circuits are an inevitable problem with such a small gap so insulating layers are used to isolate the top and bottom electrodes. Air dampers are also used to support the FLC within an airtight chamber and provide shock resistance.

Contrast ratios are as high as 100:1 and viewing angles as wide as 40 to 50°. The displays also have a fast response time (between $70\mu\text{s}$ and $120\mu\text{s}$ per line) and can handle graphical user interface software like Microsoft Windows – important because of Canon's targeting of the computer sector.

FLCDs on show at the Expo included:

- 380mm monochrome display (1280 x 1024 pixels)
- 1380mm high resolution monochrome display for personal computers (2560 x 2048 pixels)
- 380mm colour display with 16 displayable colours (1280 x 1024 pixels)
- 533mm colour display with 64 displayable colours (1280 x 1024 pixels)
- 610mm monochrome display with 64 grey scale levels (1280 x 1024 pixels)

The next step in computer displays, still video cameras, HDTV and other consumer electronics?

George Cole visits Canon's Tech Expo 93, held in Paris recently, to see where one company's R&D labs are heading.

Visitors to Tech Expo took every opportunity to sample life in tomorrow's office.

The FLCDs look impressive, especially when displaying multiple windows. But, none of the demonstrations featured full screen, moving video and FLCDs are not cheap – Canon sells the 380mm high resolution FLCD in Japan, for around £2000. A colour version would be twice the price.

Electronic photography

Canon is one of the leading supporters of still video technology and was proudly showing off its prototype high definition digital camera. It looks like a bulky 35mm SLR camera and uses conventional camera lenses and a basis (base stored image sensor) chip with 1.3 million pixels.

Most electronic cameras use a CCD array, where the charge is transferred along each element and then amplified. Unfortunately any noise produced during the transfer stage is also amplified. A basis device is a photosensor incorporating a bipolar transistor with each element having its own amplifier to reduce noise.

The camera offers different shooting modes and weighs around 1kg – power is supplied by





From slick demonstrations of digital HDTV through to the chance to see how far we have progressed in the solar charging of batteries. All were on show in Paris.

a NiCd. Each digital image uses around 1.6Mbytes of memory and the prototype uses a 120Mbyte hard disk pack (enough for 75 images) and two IC memory packs of 40Mbyte (25 images) and 80Mbyte (50 images).

The digital camera can be linked up to any workstation with a SCSI port and images transferred into a software package for editing and manipulation. The image quality is sharp and pictures look good when printed out on a colour laser printer. But launch of the digital camera is still several years off and Canon sees it primarily as an office imaging device.

World's smallest and lightest fax is already on sale in Japan.



Digital HDTV

In public, Japanese companies may be supporting the muse analogue HDTV system. But many are also busily developing digital systems. Canon has developed a large digital HDTV system in concert with Japanese broadcasting company NHK. The system uses a high speed camera which records 180 fields/s (three times the standard NTSC rate). The camera has a 0.66in harp (high gain, avalanche rushing amorphous photoconductor) image sensor and a 7Gbyte semi-conductor memory!

The HDTV codec has been developed by the company and the compression algorithm uses discrete cosine transform and variable length coding to reduce the video data stream from 1.2Gbit/s to 120Mbit/s, with no reduction in image quality. The company has also developed an HDTV transmission system for satellite, fibre optic cable and B-ISDN, with transmission speeds between 60-140Mb/s. A high definition video disc system uses jpeg compression to store 1200 still images on a 127mm magneto-optical disk. For home HDTV, Canon has also developed a VCR which stores up to 3h of video, plus four-channel digital audio.

Demonstrations were slick: Canon hopes its system will be at the heart of a number of technologies, but whether the company will be able to sell it on the world market is anyone's guess.

Magneto-optical technology

Magneto-optical (MO) systems have been used for data storage for years, and the technology has also found its way on to the consumer market in the form of Sony's MiniDisc. The new system speeds up the recording process so that MO drive speeds compare favourably with hard disk drives.

MO systems use a disk coated with cobalt

and a rare earth element such as gadolinium. Data is stored in the form of magnetic flux reversals in the magnetic material, with each pole representing a one or zero.

During recording, the disk sits on top of a drive magnet and is bathed in a magnetic field. The data stream modulates a high power recording laser which heats the disk above its Curie point allowing the drive magnet to change a bit's polarity. The modulated laser switches on and off as the disc rotates.

A low power laser is used for playback and the data is read by using the Kerr effect (polarised light twists clockwise or anti-clockwise depending on the magnetic polarity). This system is effective, but relatively slow because each recording session takes three disc rotations (for erasing, recording and verification).

Canon's faster system uses two processes called exchange coupling technology (ECT) and magnetic field modulation. In the latter process, laser power remains constant while the magnetic field is modulated. ECT uses a dual layer MO disk composed of a writing layer and memory layer, which are separate but interact with each other. The writing layer is composed of gadolinium, iron and cobalt, while the memory layer is made of terbium, iron and cobalt.

In recording, the memory layer is heated beyond its Curie point as information is recorded on to the writing layer. During cooling, the memory layer regains its magnetism and the data recorded on the writing layer is transferred on to it. The advantage of this system is that any previous data is over-written and verification is instant – so recording only involves one disk rotation.

The prototype MO system uses a single-sided 3.5in disc which stores up 350Mbyte of data. Recording speed is 2.3Mbyte/s, some 10 times faster than conventional systems.

Whizzy WIS

Wide imaging stereo (WIS) – expanding the stereo field so that listeners do not have to sit in a "hot spot" to hear the full stereo effect – marked Canon's recent move into the audio market.

A prototype system combines digital signal processing with WIS, with the DSP circuitry, electronically shaping the audio signal waveform, is claimed to improve sound quality. The project is being run by the Canon Research Centre Europe, Canon Audio and Essex University. The star office product was the world's smallest and lightest fax. Measuring 297 x 111 x 31mm it weighs just 1.2kg with battery. The mini fax uses a 9m roll of thermal paper and documents up to B4 size can be sent and received. It can be connected to public or mobile phones and a nicad battery provides around 4h of power. The fax is already on sale in Japan for around £1000, and could reach Europe this year. ■

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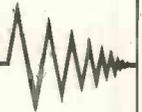
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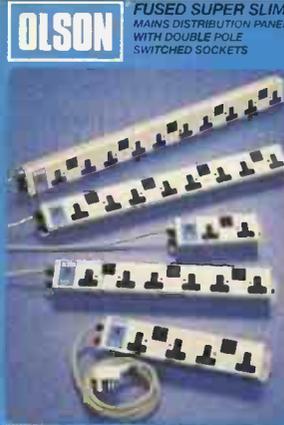


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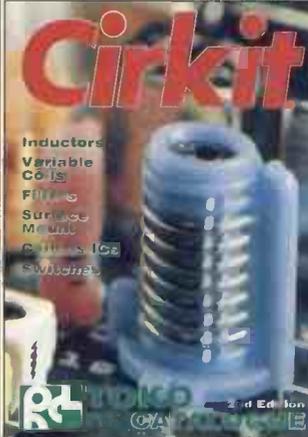
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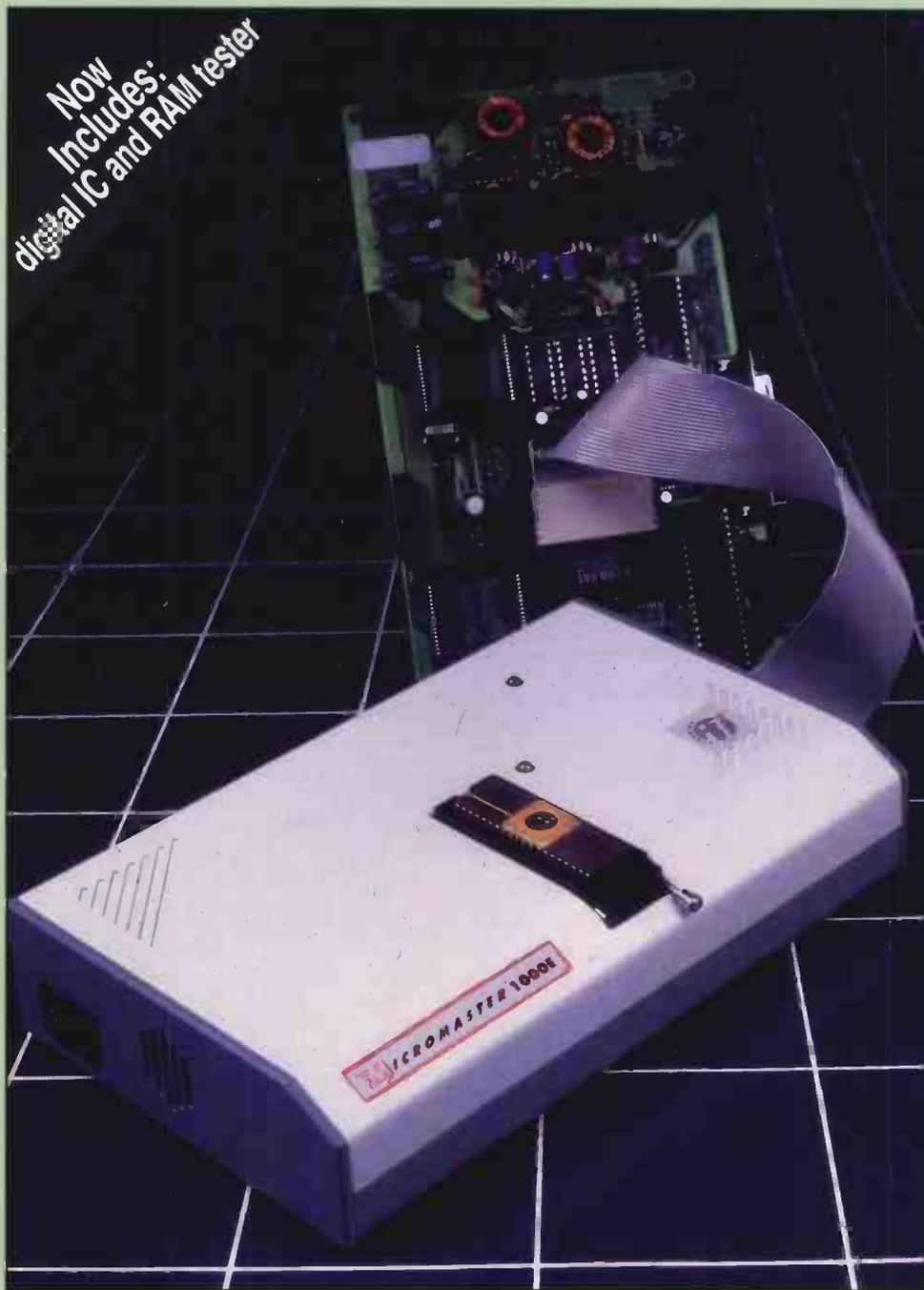
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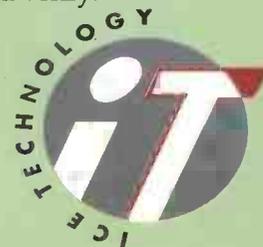
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