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The most interesting picture show in town isn't happening where you would expect it to be. Even before Japan has mastered the difficulties associated with making active matrix flat panel displays, the Koreans are rushing to set up production lines for this process critical product.

Roughly speaking, Korea is getting ready to do to the Japanese in flat panel displays what it did to them in the memory market. Except that this time, it is not even allowing Japan to enjoy a period of dominance.

There are a few things about this which are ironic. Firstly it must induce a sense of *déjà vu* to the former US memory manufacturers. Secondly the Koreans, in the shape of Hyundai, are setting up a flat panel plant in Silicon Valley's very own backyard in Oakland, California. Thirdly, their success, when it comes, will be based on advanced chemical vapour deposition equipment of US origin while the Japanese struggle with development of home grown processing technology.

Generally speaking, America isn't

terribly worried about being left out of the flat panel party even though, eventually, the value of the market for high quality TV and computer flat panel screens will compare with that of the silicon inside the machines themselves. The US computer companies, who have mostly given up any pretence of commodity component manufacture, would like nothing more than to see the Japanese and the Koreans slicing each others' margins to death on these high priced items.

There is another group of people who will also experience *déjà vu* among other emotions. Former workers at the now defunct Liverpool company Rytrak would be able to tell you about a world beating low pressure chemical vapour deposition process which could place high quality polysilicon on glass substrates up to 14in diagonal. And this was back in 1989. Set up by former employees of the deeply unimaginative GEC, Rytrak simply starved to death for want of a \$2m in long term investment. The Japanese and Koreans now feel that it is worth spending £billions just four years later.

TV through the letterbox

Am I the only person in the world to be irritated by broadcasters' attempts to condition us to letterbox pictures? While there is good argument for transmitting films originally made for the cinema with a black band above and below the picture, there is absolutely none for producing dedicated TV programmes in this format.

The combination of arrogance and stupidity among broadcasters is breathtaking. Having unsuccessfully attempted to foist MAC based pictures on an indifferent population, these twerps have now seized upon 16:9 as an excuse to make us buy more sets. No one has asked me if I am unhappy with my 4:3 picture and I don't

suppose that they have asked anyone else either.

Well, just for the record, I am happy with my 625 line, 4:3 picture. It's not perfect but it is good enough for me. I am prepared to wait for the new extended PAL systems which should allow optional widescreen without the element of compulsion. I have no intention of buying another set simply because Philips and Thomson perceive 16:9 pictures to be good for new business and have the financial clout to influence our softheaded broadcasters.

These wretched people would do far better to concern themselves with the content of the picture than its shape. Frank Ogden.

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Digital TV system attracts US interest

A four-into-one digital compression TV system developed by Britain's National Transcommunications (NTL) has attracted orders in America before its formal launch at the Las Vegas industry fair NAB in April.

System 2000 is "the first commercial implementation of the MPEG standard in broadcasting" according to NTL, the privatised transmission arm of ITV. Its initial markets will be for distribution to cable head-ends, plus satellite news-gathering and backhaul (bringing in outside broadcast material), where the primary consideration is saving transponder costs.

However, as NTL chief executive Dr John Forrest explained during a demonstration at its Crawley Court HQ, it is an "open system" and could well become the basis of a multi-channel direct-to-home service within a couple of years.

At four TV channels to a transponder, System 2000 removes more than 95% of the data from a studio standard 216Mbit/s picture to produce, with audio and error-

protection, a final 8.48Mbit/s bitstream.

In the demonstration this signal, and a 4Mbit/s signal, which would permit eight channels per transponder, were uplinked via Eutelsat to a 1.2m dish, and then shown side-by-side on two monitors. The same source, a tape including footage of horse-racing, a notably tricky subject for compression, and part of the opening sequence from *Silence of the Lambs*, was used for both. The only noticeable differences were some "crawling" around the edges of the credit-title lettering on the 4Mbit/s picture, and a slightly longer delay in the more heavily-compressed signal reaching the screen. Picture quality was virtually indistinguishable – 4Mbit/s was visibly inferior to the critical eye, but would not be noticeably so in general viewing. It was in any case better – at a horizontal resolution of 352 – than VHS video. The system can be upgraded for HDTV, at around 12Mbit/s.

As a full MPEG system, it can interface

with any other system designed to the standard, and has already been successfully matched with equipment developed by Scientific Atlanta, with which NTL has marketing and development links.

"We were able to make our equipment work with Scientific Atlanta's within a day. All eight million bits agreed," was how Mike Windram of NTL's advanced products division put it. Both systems are "as MPEG 2 as possible", pending finalisation this year of this broadcast version of the standard. NTL uses its own ICs for the encoder, and the C-Cube CL950 chip in the decoder. But System 2000 has no formal similarity with Spectre, the terrestrial digital system being developed by NTL under contract for the ITC, which is designed to fit additional digital channels around an existing analogue programme, and within the same spectrum band.

At present a System 2000 decoder costs around £4000, and an uplink £75,000 per channel, considerably cheaper than the £1 to 3 million annual rental currently being asked for a transponder. NTL has no plans to develop a domestic model itself, but is confident that, with an open industry standard, a consumer goods manufacturer could bring one to the market at a realistic price – less than £400.

Peter Willis

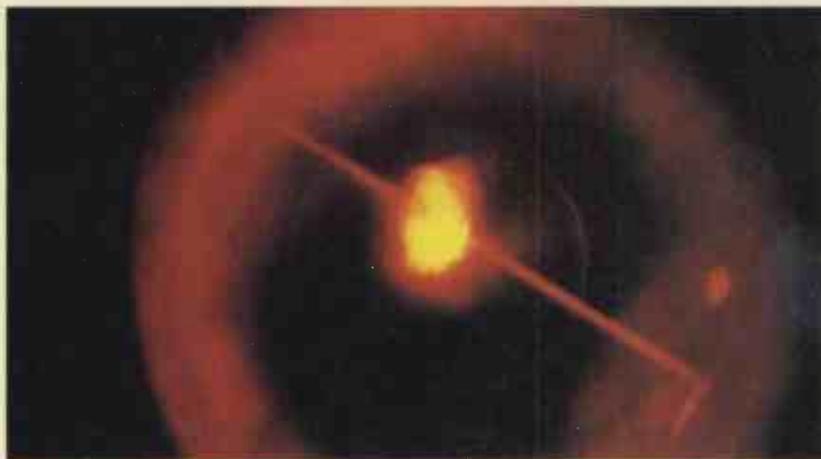
Super-cool trap brings new light

And from the darkness came light. The bright yellow spot at the centre of the photograph contains about one billion sodium atoms caught in a unique trap. The red line comprises atoms passing near the trap, some of which will also be ensnared.

This trap, built at the University of Rochester in New York, uses laser light. At room temperature atoms vibrate at about 500m/s. By hitting these atoms with powerful jets of photons, these atoms can be slowed down to speeds of only 1cm/s. And, as temperature is really

only a measure of the speed atoms vibrate, the temperature drops to within three millionths of a degree of absolute zero.

A super-cool atom puffs up to 1000 times larger than before, becoming bigger than some transistors and behaving like a fuzzy cloud. Researchers believe that these big atoms could be used for what they term atom optics. They are looking to build atom microscopes, interferometers, and even lasers using these super-cool atoms.



Radiation limits for radiocomms

The National Radiological Protection Board has included new limits on pulsed and RF radiations in the 2GHz band in new proposals for guidelines on exposure to non-ionising radiation. Recent public concern over the safety of increased mobile telephone use has resulted from an uncertainty over the effect of low power pulsed radiation. According to the NRPB: "Pulsed microwave and RF radiations can interact with tissue to produce effects which are different from those caused by continuous wave radiation."

Pulsed radiation can under certain conditions be audible to the recipient. This, said the NRPB, can be avoided at 2.45GHz if, in any 30µs interval the specific energy absorption in the head is less than 10mJ/kg, corresponding to an incident energy density of 280mJ/sq m.

Intel chip too expensive, too slow?

Motorola says it will undercut Intel's *Pentium* chip prices substantially with its *PowerPC* chips and will soon start offering first samples of its 68060 microprocessor.

A 50MHz *PowerPC 601* microprocessor will cost \$280 and the 66MHz version just \$374 in volume. This compares with Intel chip prices ranging from \$750 to \$950. The chips will be available in volume quantities in the third quarter of 1993. Motorola hopes that its aggressive prices will attract customers wishing to build systems to compete with *PowerPC* systems from IBM

and Apple.

The 68060 is a more powerful version of its 68040 microprocessor. Sources close to Motorola say that the microprocessor will be available in sample quantities in the third quarter of 1993 with volume shipments beginning in early 1994.

So far, Apple is the only systems manufacturer planning to use the 68060 in new products scheduled for introduction in early 1994. Pricing has not yet been set, but is expected to be around \$500 for the 50MHz version. There will also be a 66MHz version.

NCR plans a wireless world

After three years of successful operation in the US, wireless lans from NCR are being introduced around the world. The firm has introduced the product in some European countries in the last month and plans to spread it to other countries on the continent during the summer.

There has also just been a launch in Japan, and further launches are planned this year for Australia, the Far East, and the Middle East.

Called *Wavelan*, the product uses spread spectrum technology to link computers by

radio waves rather than cabling. They will operate at frequencies around 2.5GHz depending on local conditions and frequency allocations.

NCR first started working on this project some six to seven years ago because it saw a change in social trends towards greater mobility.

Dr Khaled Marrei, an NCR vice president, said: "People are no longer accepting being tethered to terminals, telephones, and work areas. Offices, work areas, and service locations are continuously changing.

Organisational structure is changing based on skills and not physical locations."

One of the problems with this is that computers and point of sale machines tend to be limited in mobility because of the cabling. The idea behind wireless lans is to link them using the radio waves. This means they can be moved anywhere in an office or work area without tearing up carpets and ceilings to move the wiring.

The product comes in the form of a plug-in 16-bit AT or MCA card with a radio antenna attached. A PCMCIA version is planned for later this year. The system is fully compatible with Ethernet or token ring networks and can be linked to existing wired networks using the firm's Wavepoint access bridge.

In an open office the transmission range is between 150 and 180m. If there are partitions (with the antenna below the partition level) this drops to 30 to 60m. In closed offices with walls then 15 to 30m is the range. And in offices with concrete walls the range drops to less than 10m.

The spread spectrum technology used has its origins in the 1940s when it was used for secure military communications. The transmitter diffuses its radio signals over a range of frequencies making it difficult to detect, jam, or monitor. The NCR product has extra security by way of user-codeable encryption chips on the *Wavelan* card.

Silicon fluid valve

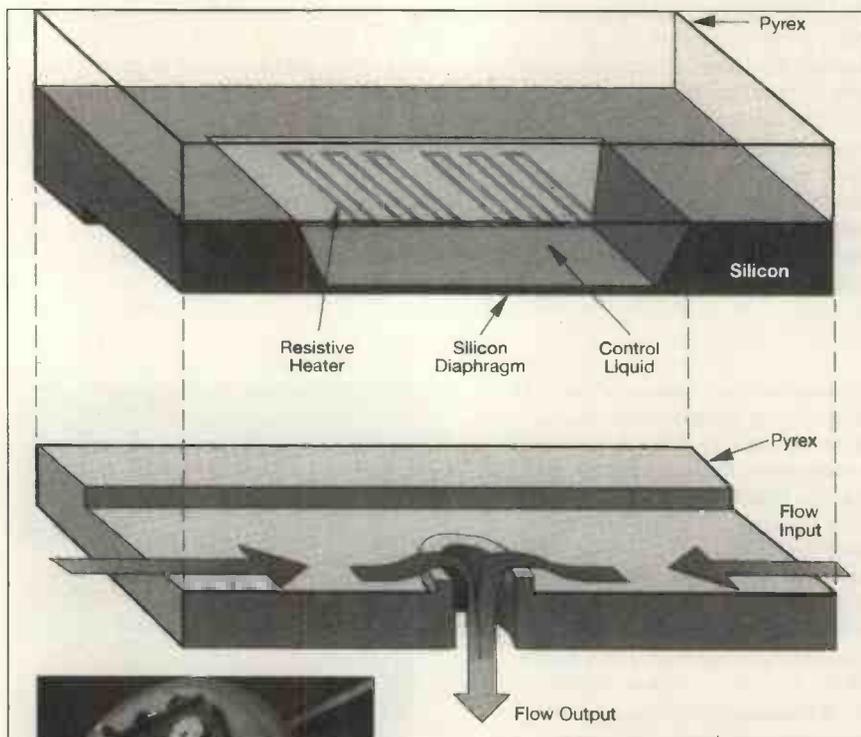
Redwood Microsystems is shipping commercial quantities of silicon valves built using thermopneumatic technology developed at Stanford University.

Redwood can make an electrically controlled fluid valve using semiconductor etching techniques which is smaller, more accurate and potentially much cheaper than mechanical alternatives, according to Redwood's president Perry Constantine. Primary market will be instrumentation for process control and food production.

In the *Fluistor* valve, a well is etched out of a wafer of silicon leaving only a thin membrane at the bottom to act as a diaphragm. The well is filled with a control liquid. A piece of Pyrex to which metal heating elements have been fixed is then sandwiched on top of the silicon so that the wires are in contact with the control liquid. This is then attached to a second Pyrex wafer underneath the silicon which contains channels and holes through which the fluid to be controlled can pass.

As current flows through the heating elements, the control liquid expands and forces the silicon membrane downwards restricting flow through the channels in the Pyrex. Wires sandwiched between the channel layer and the silicon can inform the control circuitry when the valve is fully closed so that the power can be turned off.

Deflection of the silicon varies directly



with applied current making the valve suitable for precision work. Deflection can vary between 0 and 50µm and the elastic properties of silicon means it returns to its original position when the fluid is cooled.
Redwood Microsystems
415-617 1200

Russian semiconductors to get cheap boost

Russia is to pump 120bn roubles (£120m) of state aid into its electronics industry this year with priority going to programmes assisting the conversion of military production into commercial production. Russian silicon costs just one fifth of the Western equivalent.

"The programme has been approved by President Yeltsin and the Supreme Soviet in the budget for 1993," said Anatoly Andreev, Head of the State Committee on the Defence Industry, announcing the decision at a recent conference in Zelenograd, near Moscow, organised by East/West Electronics of Sevenoaks.

"We regard electronics as a priority," said Andreev, "and particularly microelectronics where the potential of the Russian market is enormous... unlimited. For instance in Russia the number of chips owned per head of population is eight compared to 200 per head in Japan and America.

"In microelectronics we must do and do, and manufacture and manufacture," added Andreev.

"We have a powerful scientific structure

and we have to put it to the service of the microelectronics industry which is only 60% loaded". Asked if he was expecting to get some of Andreev's 120bn roubles, Valery Dshkhunyan, director of Angstrom, one of Russia's largest chip companies, said: "In May or June there may be some money transferred to us. Last year there was no government investment. This year, in spite of the difficult situation, the government is beginning to realise that electronics is a key industry where government support is necessary. At least the government realises our difficulties. Up to now we've been mostly relying on low interest rate credits and co-operation with customers."

Andreev defined the scale of Dshkhunyan's task: "In 1992 80% of microelectronics output went into military applications." In the West, the equivalent figure is 3.4%.

One of many surprising revelations arising from the conference was that Russian semiconductor companies can make a 4in wafer for \$35.

Western delegates were asking their

Russian counterparts why, if their costs are so low, their wafer fabs are only half full. The cost in a Western fab for a 4in wafer of equivalent technology would be around \$250.

Some companies are taking advantage of this low production cost. Micron, one of the big chip producers, is selling \$1m worth of 4in wafers to Samsung of Korea every year. According to Micron's Vladimir Necludoff they are charging Samsung \$35 for an unscribed wafer and \$50 for a wafer of diced die.

The chips being sold by Micron to Samsung are calculator and watch chips. The technology involved is 1.5µm cmos.

And Angstrom is making calculator chips, games chips, microcontrollers, and standard cell asics for customers in Hong Kong, Singapore and Taiwan.

"It is not a strategic policy," said Dshkhunyan. "It is designed for survival until we go to a higher level of technology." The Angstrom plan is to move to higher value products incorporating higher technology. **David Manners**

EC initiative boosts UK competitiveness

The international competitiveness of the UK's manufacturing industry will benefit from the success of universities in producing a new breed of manufacturing systems engineer, according to the Engineering Council.

Broadly based courses, introduced in 1988 as part of the council's manufacturing systems engineering initiative, are key to meeting industry's needs for engineers who can work with new manufacturing technologies and automated processes, says the council.

Professor Keith Foster from the council said: "The success of this initiative should certainly change the views of those who have blamed industry's lack of growth on the shortage of people qualified in manufacturing disciplines."

He added: "The courses appear to have attracted students looking for a broader educational base in engineering. Many will take a career in line management in industry which should lead to excellent prospects for moving into top management."

Pictured are the three winners of the 1993 Young Electronic Designer Awards, sponsored by Texas Instruments and Mercury Communications. They are from the left Nicola Hay, Philip Pegden, and Emma Lye.

Hay of Woldingham School in Surrey achieved her success with an electronic device to monitor water contamination in brake fluid, now a requirement of the MOT. Pegden of Tonbridge School designed a computerised quadraphonic sound effects system for theatres. And Lye of Bancroft's School in Essex won her prize with an electronic elbow which tests the temperature of a baby's bath water.

The Engineering Council has just announced a new set of awards to promote inventiveness and originality in young people in engineering design and development. They will run parallel with the council's existing Young Engineer for Britain competition.



How much do you earn?

Average earnings of IEEIE members have increased by 9.28% since 1991, according to the latest survey by the institution. This compares with 7.7% overall for non-manual employees in industry.

The annual salary for highest earners has gone up by £5000 to £39,000 while the median earnings of IEEIE members are at £24,000 with more than 47% of those surveyed earning this amount.

The best pay in 1991 was in broadcasting, telecommunications, and postal services. Today the big money can be found in electricity generation and distribution, and chemical and allied processing.

Some 21% of members have an HND or degree compared with only 18% in 1991.

World's biggest gigaflop

What is expected to be the highest performance supercomputer in the world, the Intel Paragon XP/S, is on schedule for completion this summer.

It is being installed in the Sandia National Laboratories' plant in Albuquerque, New Mexico, and will be used by researchers in the USA for military, scientific, and industrial problems.

When finished its performance will be 140Gflops with 1872 parallel computational nodes each consisting of two i860XP processors. Computer memory will be 38Gbyte.

Arthur Hale, manager of Sandia's parallel computing science department, said: "The computer has performance and memory capabilities that should allow us to solve problems that were previously unsolvable."

HSE investigates new glycol ether miscarriage risks

The Health and Safety Executive is re-examining the risks to women working in the semiconductor manufacturing industry following recent reports of miscarriages associated with work involving exposure to glycol ethers.

The HSE is about to issue information to the industry on the control of exposure.

A recent study by the American Semiconductor Industry Association found a miscarriage rate of 12 to 14% for women

exposed to glycol ethers compared to a normal background rate of 10% in other groups.

The use of glycol ethers in the UK is already covered under the Coshh regulations. But Coshh came into force in 1989, before the recent studies.

Some have already started to phase out the use of the most potentially hazardous glycol ethers.

Pal plus to lead widescreen introduction?

Could the first widescreen TV services be in Palplus – even on satellite? The question arose at the first live public demonstration of the system, hosted by Nokia as part of its London Trade Show in April.

"Unlike digital systems," said Richard Ellis, head of the UK Palplus development group, who introduced the showing, "Palplus is here today, specified and capable of being transmitted over terrestrial or satellite systems.

"With the apparent collapse of the Mac directive, it may be that a new European initiative for wide-screen will accelerate the introduction of Palplus."

The development programme is on schedule, and prototype receivers, working from over-air transmissions will be shown at the Berlin Funkaustellung this autumn, with a possible service launch in 1995.

The Palplus initiative came originally from a consortium of European terrestrial broadcasters, including the BBC and ITV, concerned that they would be squeezed out if satellite channels started widescreen Mac services.

It was a "defensive strategy" to protect the viewing base. There are still no plans "at

this stage" for terrestrial broadcasters to lead the field in widescreen, said Ellis. However, he emphasised its suitability for satellite.

"Palplus would go through existing transmitters with very little modification, and will travel satisfactorily over satellite channels," he explained. The cost of a decoder would be "something like a mid-market satellite receiver," he added.

Palplus relies on processing a 576 line source down to 432 active lines – a letterbox picture – and using a helper signal, similar to the colour subcarrier and buried in the black bands, to restore picture quality.

Palplus also reduces those old Pal bugbears, cross-colour and cross-luminance artifacts. The result, as seen in the demonstration, is a picture of impressive quality, though the chosen material – slow-pans and zooms around a garden, with close-ups of flowers – was criticised by some in the audience as too easy.

The only real drawback to satellite introduction at present is encryption.

"The Palplus algorithm is pretty complicated, and if you get anything out of place it won't work," Ellis explained. Encryption, he said, "is being looked at".

Peter Willis

Meteor scatter for voice comms

Mobile communications systems which bounce radio signals off meteor trails in the earth's upper atmosphere are becoming increasingly important to the military says the latest Jane's Military Communications. But they are also to enjoy a civil boost.

The Department of Trade and Industry has already awarded a UK licence to one operator planning to use the meteor burst technology for commercial use.

Last year US radio specialist Meteor Communications Corp (MCC), working with Alascom, demonstrated the first two-way voice link using meteor burst technology over a distance of 940km.

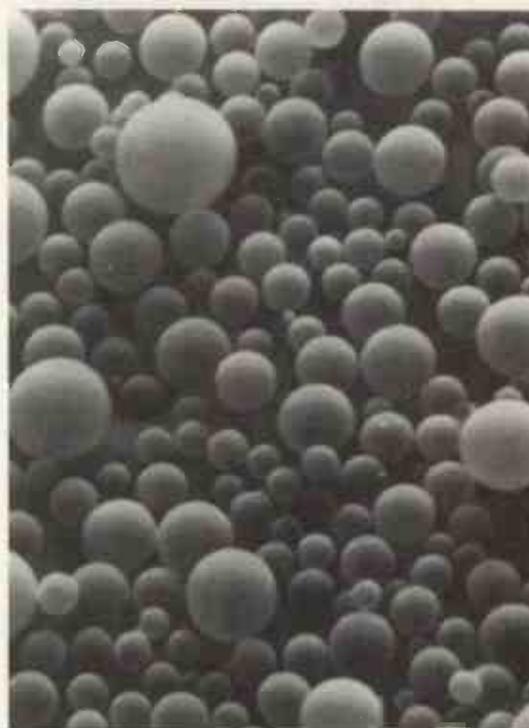
The systems bounce radio signals off the ionised vapour trails of meteors entering the

earth's atmosphere 100km above the earth. These trails are short-lived, typically less than 2s, but the sheer number of micro-meteors entering the atmosphere, billions a day, create a reliable radio wave reflector.

The meteor systems are long range, up to 2000km, and highly directional making them harder to jam or tap.

Meteor Communications' European subsidiary has been licensed by the DTI to use meteor burst technology for telemetry services to mobile or fixed terminals in the UK and Western Europe.

The company's meteor burst communications station operates in the 40 to 50MHz band with a 500W transmit power. The system can support 4 or 8kbit/s data rates over 2000km



Hollow spheres from just five microns across have been introduced by Emerson & Cuming to provide reduced dielectric constants for electronics materials. The Eccosphere SDT Series of microspheres have been designed for use in thin film laminates, pastes and tapes.

Texas talk of MPEG audio

Texas Instruments has entered the lucrative MPEG decompression chip market with a single-chip MPEG audio IC, with plans for video compression chips to follow later this year.

TI's first offering is a top-of-the-range audio chip, costing around \$15, with built-in datastream parsing and timestamp extraction for synchronisation with video. Gerard Benbassat, TI's multimedia technical marketing boss, said simplified versions selling at less than \$10 will appear soon.

TI has a technology exchange deal for MPEG video decompression technology with C-Cube Microsystems, and Benbassat said a chip will be announced later this year. He added that a real-time MPEG video compression chip, using homegrown technology, will also be launched this year.

TI is the latest of a clutch of major chip makers to announce MPEG ICs, including Motorola, SGS Thomson, Philips, and LSI Logic.

"We believe compression is going to be the major semiconductor market of the late 1990s, replacing the computer market of the 1980s," Benbassat forecast.

He sees digital radio, due to hit the market in 1995, as one of the largest potential markets for audio decompression: "There are around 700 million FM radio tuners in Europe," he pointed out. ■

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 HP Sampling Voltmeter (Broadband) type 3406A - £200.
 HP Vector Voltmeter type 8405A - £400 to £600 - old or new colour.
 HP Synthesizer/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.
 HP Down Converter type 11710B - .01-11Mc/s - £450.
 HP Pulse Modulator type 11720A - 2-18GHz - £1000.
 HP Modulator type 8403A - £100-£200.
 HP Pln Modulators for above-many different frequencies - £150.
 HP Counter type 5342A - 18GHz - LED readout - £1500.
 HP Signal Generator type 8640B - Opt001 + 003 - .5-512Mc/s AM/FM - £1000.
 HP Amplifier type 8447A - 1-400Mc/s £200 - HP8447F .1-1300Mc/s £400.
 HP Frequency Counter type 5340A - 18GHz £1000 - rear output £800.
 HP 8410 - A - B - C Network Analyser 110Mc/s to 12GHz or 18GHz - plus most other units and displays used in this set-up - 8411A - 8412 - 8413 - 8414 - 8418 - 8740 - 8741 - 8742 - 8743 - 8746 - 8650. From £1000.
 HP Signal Generator type 8660C - .1-2600Mc/s. AM/FM - £3000. 1300Mc/s £2000.
 HP Signal Generator type 8656A - 0.1-990Mc/s. AM/FM - £2000.
 HP 8699B Sweep Pl - 0.1-4GHz £750 - HP8690B Mainframe £250.
 Racal/Dana 9301A-9302 RF Millivoltmeter - 1.5-2GHz - £250-£400.
 Racal/Dana Counters 9915M - 9916 - 9917 - 9921 - £150 to £450. Fitted FX standards.
 Racal/Dana Modulation Meter type 9009 - 8Mc/s - 1.5GHz - £250.
 Racal - SG Brown Comprehensive Headset Tester (with artificial head) Z1A200/1 - £350.
 Marconi AF Power Meter type 893B - £200.
 Marconi RCL Bridge type TF2700 - £150.
 Marconi/Saunders Signal Sources type - 6058B - 6070A - 6055B - 6059A - 6057B - 6056 - £250-£350. 400Mc/s to 18GHz.
 Marconi TF1245 Circuit magnification meter + 1246 & 1247 Oscillators - £100-£300.
 Marconi microwave 6600A sweep osc., mainframe with 6650 PI - 18-26.5GHz or 6651 PI - 26.5-40GHz - £1000 or PI only £600.
 Marconi distortion meter type TF2331 - £150, TF2331A - £200.
 Microwave Systems MOS/3600 Microwave frequency stabilizer - 1GHz to 40GHz £1k.
 Tektronix Plug-ins 7A13 - 7A14 - 7A18 - 7A24 - 7A26 - 7A11 - 7M11 - 7S11 - 7D10 - 7S12 - S1 - S2 - S6 - S52 - PG506 - SC504 - SG502 - SG503 - SG504 - DC503 - DC508 - DD501 - WR501 - DM501A - FG501A - TG501 - PG502 - DC505A - FG504 - P.O.R.
 Alltech Stoddart receiver type 17/27A - .01-32Mc/s - £2500.
 Alltech Stoddart receiver type 37/57 - 30-1000Mc/s - £2500.
 Alltech Stoddart receiver type NM65T - 1 to 10GHz - £1500.
 Gould J3B Test oscillator + manual - £200.
 Infra-red Binoculars in fibre-glass carrying case - tested - £100. Infra-red AVF sights £100.
 ACL Field Intensity meter receiver type SR - 209 - 6. Plugs-ins from 5Mc/s to 4GHz - P.O.R.
 Tektronix 491 spectrum analyser - 1.5GHz-40GHz - as new - £1000 or 10Mc/s-40GHz.
 Tektronix Mainframes - 7603 - 7623A - 7633 - 7704A - 7844 - 7904 - TM501 - TM503 - TM506 - 7904 - 7834 - 7104.
 Knott Polyskanner WM1001 + WM5001 + WM3002 + WM4001 - £500.
 Alltech 136 Precision test RX + 13505 head 2 - 4GHz - £350.
 SE Lab Eight Four - FM 4 Channel recorder - £200.
 Alltech 757 Spectrum Analyser - 001 22GHz - Digital Storage + Readout - £3000.
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 Precision Aneroid barometers - 900-1050Mb - mechanical digit readout with electronic indicator - battery powered. Housed in polished wood carrying box - tested - £100-£200-£250, 1, 2 or 3.
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 HP141T+8552A or B IF-8553B RF - 1kHz-110Mc/s-A IF - £1300 or B IF - £1400.
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 HP141T+8552A or B IF-8556A RF - 20Hz-300kHz-A IF-A IF - £1200 or B IF - £1300.
 HP8443A tracking generator/counter - 100kHz-110Mc/s - £500.
 HP8445B tracking pre-selector DC-18GHz - £750.
HP ANZ UNITS AVAILABLE SEPARATELY - NEW COLOURS - TESTED.
 HP141T mainframe - £550 - 8552A IF - £450 - 8552B IF - £550 - 8553B RF - 1kHz-110Mc/s - £550 - 8554B RF - 100kHz-1250Mc/s - £650 - 8555A RF - 10Mc/s-18GHz - £1550.
 HP 3580A LF-spectrum analyser - 5kHz to 50kHz - LED readout - digital storage - £1600 with instruction manual - internal rechargeable battery.
 Tektronix 7D20 plug-in 2-channel programmable digitizer - 70 Mc/s - for 7000 mainframes - £500 - manual - £50.
 Datron 1065 Auto Cal digital multimeter with instruction manual - £500.
 Racal MA 259 FX standard. Output 100kc/s-1Mc/s-5Mc/s - internal NiCad battery - £150.
 Aerial array on metal plate 9"x9" containing 4 aerials plus Narda detector - .100-11GHz. Using N type and SMA plugs & sockets - except - £100.
 EIP 451 microwave pulse counter 18GHz - £1000.
 Marconi RF Power Amplifier TF2175 - 1.5Mc/s to 520Mc/s with book - £100.
 Marconi 6155A Signal Source - 1 to 2 GHz - LED readout - £600.
 Schlumberger 2741 Programmable Microwave Counter - 10Hz to 7.1GHz - £750.
 Schlumberger 2720 Programmable Universal Counter 0 to 1250Mc/s - £600.
 HP 2225CR Thinkjet Printer - £100.
 TEK 576 Calibration Fixture - 067-0597-99 - £250.
 HP 8006A Word Generator - £150.
 HP 1645A Data Error Analyser - £150.
 Texscan Rotary Attenuators - BNC/SMA 0-10-60-100DBS - £50-£150.
 HP 809C Slotted Line Carriages - various frequencies to 18GHz - £100 to £300.
 HP 532-536-537 Frequency Meters - various frequencies - £150-£250.
 Barr & Stroud variable filter EF3 0.1Hz-100kc/s + high pass + low pass - £150.
 S.E. Lab SM215 Mk11 transfer standard voltmeter - 1000 volts.
 Alltech Stoddart P7 programmer - £200.
 H.P. 6941B multiprogrammer extender. £100.
 Fluke Y2000 RTD selector + Fluke 1120A IEEE-488-translator + Fluke 2180 RTD digital thermometer + 9 probes. £350 all three items.
 H.P. 6181 DC current source. £150.
 H.P. 59501A - HP-IB isolated D/A/power supply programmer.
 H.P. 3438A digital multimeter.
 H.P. 6177C DC current source. £150.
 H.P. 6207B DC power supply.
 H.P. 7418 AC/DC differential voltmeter standard (old colour) £100.
 H.P. 6209B DC power unit.
 Fluke 80 high voltage divider.
 Fluke 431C high voltage DC supply.
 Tektronix M2 gated delay calibration fixture. 067-0712-00.
 Tektronix precision DC divider calibration fixture. 067-0503-00.
 Tektronix overdrive recovery calibration fixture. 067-0608-00.
 Avo VCM163 valve tester+book £300.
 H.P. 5011T logic trouble shooting kit. £150.
 Marconi TF2163S attenuator - 1GHz. £200.
 PPM 8000 programmable scanner.
 Fluke 730A DC transfer standard.
 B&K 4815 calibrator head.

B&K 4812 calibrator head.
 Farnell power unit H60/50 - £400 tested.
 H.P. FX doubler 938A or 940A - £300.
 Racal/Dana 9300 RMS voltmeter - £250.
 H.P. sweeper plug-ins - 86240A - 2-8.4GHz - 86260A - 12.4-18GHz - 86260AH03 - 10-15GHz - 86290B - 2-18.6GHz. 86245A 5.9-12.4GHz.
 Telequipment CT71 curve tracer - £200.
 H.P. 481A amplifier - 1kc-150Mc/s - old colour - £100.
 H.P. 8750A storage normalizer.
 Tektronix oscilloscopes type 2215A - 60Mc/s - c/w book & probe - £400.
 Tektronix monitor type 604 - £100.
 Marconi TF2330 or TF2330A wave analysers - £100-£150.
 HP5006A Signature Analyser £250 + book.
 HP10783A numeric display. £150.
 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.
 HP 5240A counter - 10Hz to 12.4GHz - £400.
 HP 3763A error detector.
 HP 8016A word generator.
 HP 489A micro-wave amp - 1-2GHz.
 HP 8655A spectrum analyser .01-22GHz - £4k.
 HP 5065A rubidium vapour FX standard - £5k.
 Fluke 893A differential meters - £100 ea.
 Systron Donner counter type 6054B - 20Mc/s-24GHz - LED readout - £1k.
 Takeda Riken 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.
 Tektronix plug-ins - AM503 - PG501 - PG508 - PS503A.
 Tektronix TM515 mainframe + TM5006 mainframe.
 Cole power line monitor T1085 - £250.
 Claude Lyons LCM1P line condition monitor - £250.
 Rhodes & Schwarz power signal generator SLRD-280 - 2750Mc/s. £250-£600.
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 Trend Data tester type 100 - £150.
 Farnell electronic load type RB1030-35.
 Fairchild interference analyser model EMC-25 - 14kc/s-1GHz.
 Fluke 1720A instrument controller+keyboard.
 Marconi 2442 - microwave counter - 26.5GHz - £1500.
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 B&K 2425 voltmeter - £150.
 B&K 4921+4149 outdoor microphone.
 Wiltron sweeper mainframe 610D - £500.
 HP3200B VHF oscillator - 10-500Mc/s - £200.
 HP3747A selective level measuring set.
 HP3586A selective level meter.
 HP5345A electronic counter.
 HP4815A RF vector Impedance meter c/w probe. £500-£600.
 Marconi TF2092 noise receiver. A, B or C plus filters.
 Marconi TF2091 noise generator. A, B or C plus filters.
 Tektronix oscilloscope 485 - 350Mc/s - £500.
 HP180TR, HP182T mainframes £300-£500.
 Bell & Howell CSM2000B recorders.
 HP5345A automatic frequency converter - .015-4GHz.
 Fluke 8506A thermal RMS digital multimeter.
 HP3581A wave analyser.
 Phillips panoramic receiver type PM7800 - 1 to 20GHz.
 Marconi 6700A sweep oscillator + 6730A - 1 to 2GHz.
 Wiltron scaler network analyser 560+3 heads. £1k.
 R&S signal generator SMS - 0.4-1040Mc/s - £1500.
 HP8558B spectrum ANZ PI - 1-1500Mc/s - dc - £1000. N/C - £1500 - To fit HP180 series mainframe available - £100 to £500.
 HP8505A network ANZ + 8503A S parameter test set + 8501A normalizer - £4k.
 HP8505A network ANZ + 8502A test set - £3k.
 Racal/Dana 9087 signal generator - 1300Mc/s - £2k.
 Racal/Dana VLF frequency standard equipment. Tracor receiver type 900A + difference meter type 527E + rubidium standard type 9475 - £2750.
 Marconi 6960-6960A power meters with 6910 heads - 10Mc/s - 20GHz or 6912 - 30kHz - 4.2GHz - £800-£1000.
 HP8444A-HP8444A opt 59 tracking generator £1k-£2k.
 B&K dual recorder type 2308.
 HP8755A scaler ANZ with heads £1k.
 Tektronix 475 - 200Mc/s oscilloscopes - £350 less attachments to £500 c/w manual, probes etc.
 HP signal generators type 626 - 628 - frequency 10GHz-21GHz.
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 HP3730B down converter - £200.
 Bradley oscilloscope calibrator type 192 - £600.
 Spectrascope SD330A LF realtime ANZ - 20Hz-50kHz - LED readout - tested - £500.
 HP8620A or 8620C sweep generators - £250 to £1k with IEEE.
 Barr & Stroud variable filter EF3 0.1Hz-100kc/s+high pass+low pass - £150.
 Tektronix 7L12 analyser - .1Mc/s-1.8GHz - £1500 - 7L14 ANZ - £2k.
 Marconi TF2370 spectrum ANZ - 110Mc/s - £1200-£2k.
 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.
 HP8614A signal gen 800Mc/s-2.4GHz old colour £200, new colour £400.
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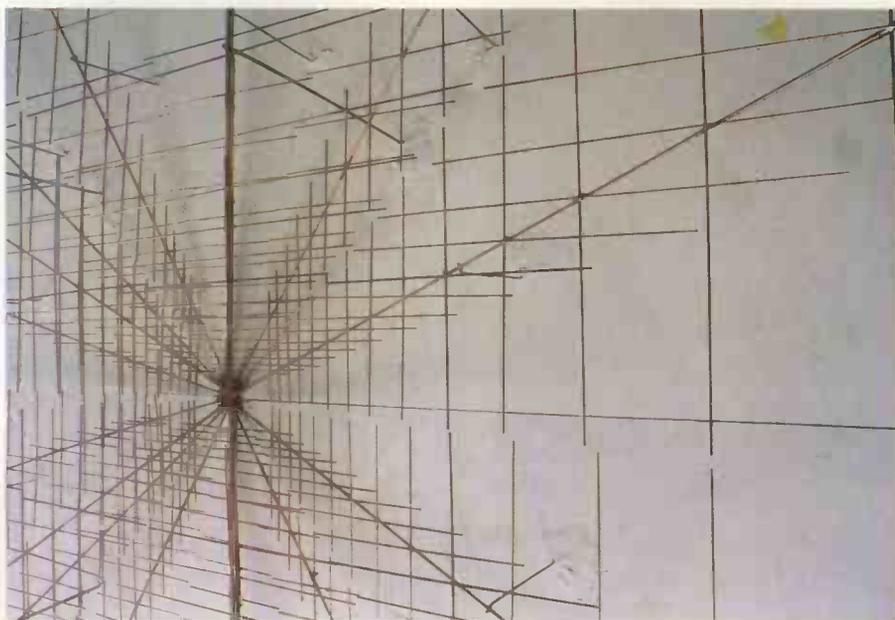
RESEARCH NOTES

Limits pushed back for phased array antennas

A Penn State University researcher has developed a novel nesting technique which could eliminate many of the drawbacks of that vital ingredient of communications and weather forecasting systems – the phased array antenna. The technique is the brainchild of James Breakall, who has designed what he calls a three dimensional frequency-independent phased-array antenna, or 3D-fipa.

In theory the system has no upper frequency limit, requires no band switching, no frequency selective filters and no other frequency dependent hardware. Large bandwidth also offers rapid beam steering and frequency agility.

Conventional phased array antennas are



Scale model of the 3D-fipa antenna constructed at the Applied Research Lab, Penn State University.

made up of numerous single antennas, individually limited by their size to a particular narrow frequency range. The range of such arrays is also limited by their configuration, especially the spacing between the centres of the antennas. This is physically fixed, but altered electrically to accommodate small frequency changes and to steer the beam.

With such conventional designs, an increase in the operating frequency eventually leads to development of extraneous lobes which compete with the main lobe, interfering with and skewing the beam. The effect can only be avoided by using complex multiple arrays in which the electrical spacing between the individual antennas changes only minimally with frequency.

Breakall's design sidesteps many of these limitations by using an entirely different geometry. His design relies on an original

version of the log periodic antenna, where every part of any section of a structure is a scaled version of another part of that structure. In effect the antenna is a 3D fractal.

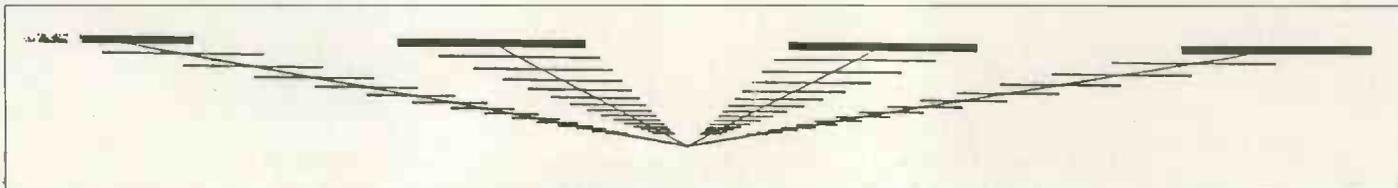
The 3D-fipa has its elements arranged to form a multi-layer or nested structure in which the spacings and heights are constant over a specified frequency range. The arrangement allows wideband working and is also claimed to reduce radically the development of unwanted sidelobes or main beam distortion.

First full-scale implementation of a 3D-fipa antenna will most likely be in Sweden where it will form part of a 10-13GW ionospheric radar covering the frequency range of 5-70MHz

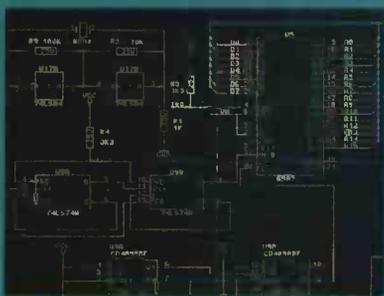


Professor James Breakall has been tackling drawbacks of the phased array antenna.

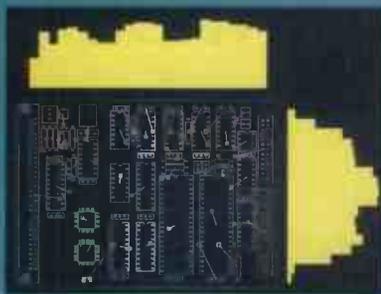
Single plane phased array log-periodic system



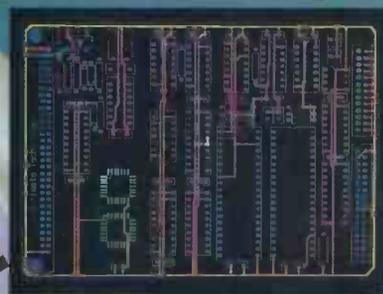
FROM CONCEPT TO ARTWORK IN 1 DAY



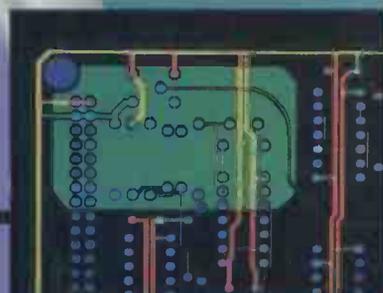
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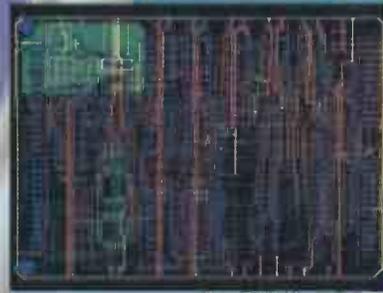
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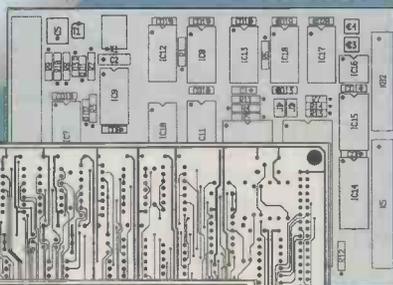
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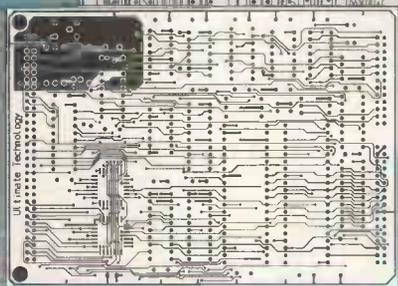


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Longer days are not just for summer

Readers in the UK and much of the northern hemisphere are now enjoying longer hours of daylight thanks to the arrival of summer. But how many realise that the day as a whole has been getting ever so slightly longer for thousands of years? Now, work deciphering ancient clay tablets at the British Museum has enabled researchers to put an accurate figure on this change.

Astronomers have long known that day length has been on the increase – Royal Greenwich Observatory investigators have been studying variations in the day length for many decades – though it has been very difficult to make any real measure of how much. But recently, Dr Leslie Morrison, an astronomer at the RGO, and Dr Richard Stephenson of Durham University explained to a meeting of the Royal Astronomical Society how they had used the tablets to quantify the effect.

The tablets, originally made in Babylon 2700 years ago, contain records of lunar and solar eclipses in cuneiform script. They have only fairly recently been translated and contain information about when and where eclipses were seen in ancient Babylon, all carefully and accurately recorded.

It seems the ancient Babylonians noted the time at which the Sun went over the horizon and then measured the interval between sunset and the onset of lunar eclipses. The time was measured, not with hour glasses, but by measuring the angular movement of the stars across the heavens. One degree of movement corresponds to four minutes, so ancient eclipses were recorded to an accuracy of four minutes.

Using the inevitable computer, Morrison and Stephenson worked out when these various eclipses *should* have happened, on the assumption that the time from then until now has consisted of days that are the same

length as today's. They found that in every case the discrepancy between the time the Babylonians saw an eclipse and the time calculated by the computer was something like five hours. This implies that over a period of 2700 years the Earth's rotational clock has in fact lost five hours.

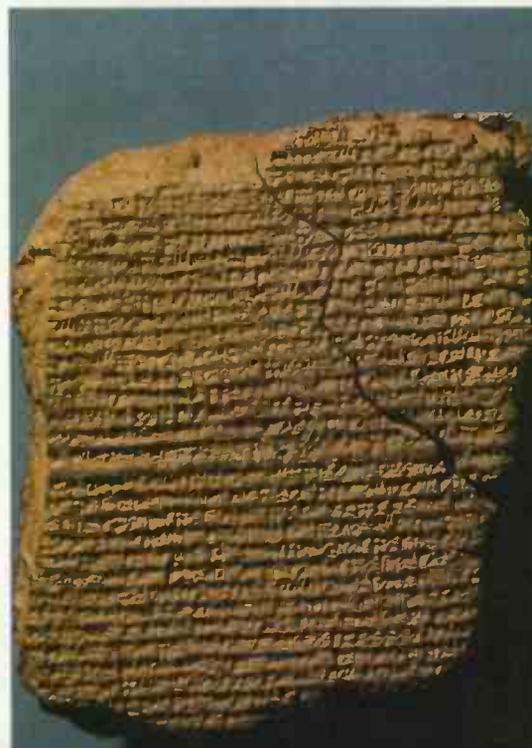
Five hours over nearly 3000 years may not seem like very much – and indeed it isn't. Translating the figures into everyday terms means that the day has lengthened by an average of 1-7ms every century. It is a tiny amount, even over the whole of human history. But as Dr Morrison points out, it is a great deal since the Earth was originally formed.

What has been happening over geological time scales is a transfer of angular momentum between the spin of the Earth and the orbit of the moon. The earth is gradually spinning more slowly while the moon moves away from the Earth at about 0.03m a year. Eventually (in millions of years time) the earth will end up with one face toward the moon all the time.

Then the length of the day and the length of the month will both become about 47 of our present days. Just think of the administrative complications that will cause!

Top, Babylonian tablets are still revealing secrets for modern astronomers.

The actual time of lunar eclipses has been different from the calculated time, due to lengthening of the day.



Sound method for keeping cool

Instruments lowered deep into boreholes to listen to underground murmurs can survive only a short time before the natural heat from the Earth overwhelms the package and causes the electronic components to fail. The limited duration considerably hampers attempts, for example, to record and analyse seismic signals. But an "acoustic" refrigerator no bigger than two beer cans stacked end to end could offer a new way to protect electronics from the high temperatures encountered at the bottoms of geological exploration wells.

The miniature fridge, designed and patented by Gloria Bennett of the Nuclear Technology and Engineering Division of the Los Alamos National Laboratory in

New Mexico, would allow instruments for geological analysis to work for many days at a time at maximum depth, where the most useful and sensitive information is obtained.

Design relies on a sound wave to compress a gas and initiate a cooling cycle not unlike that of a domestic fridge. But there are no moving parts and no fluids to leak.

The unit starts with a heat-driven acoustic engine, converting heat into sound. Emitted into a volume of the right shape, the sound can be maintained as a standing acoustic wave capable of compressing a gas that fills the volume. The standing wave powers a cycle of compression and expansion that makes

one end of the refrigerator hot and the other end cool, providing a way to expel heat from the system.

Miniature pipes conduct heat from an electronics package to the cool end of the refrigerator while larger pipes convey the heat from the hot end to the well-bore. In practice this can cool the electronics package from a damaging 300°C to a manageable 150°C

The patented design was developed specifically for the needs of the Los Alamos Laboratory's dry rock geothermal energy programme. But it could find an application in anything from deep-bore oil exploration to basic geological or seismic research.

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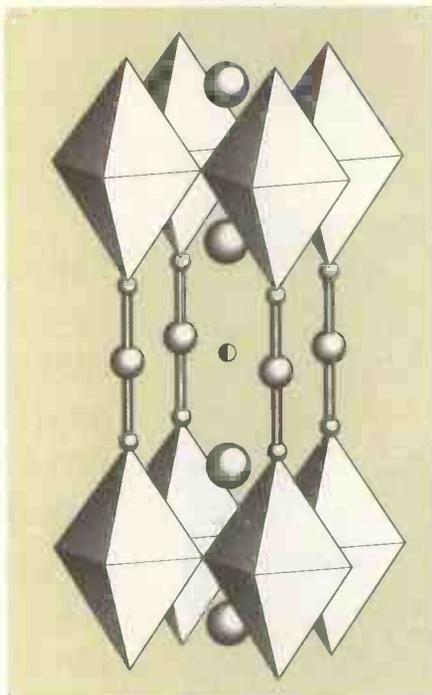
Progress in high temperature superconductors – materials that lose all their resistance at temperatures well above absolute zero – looks to be revitalised by a development of a composite material containing mercury.

Most ceramic materials are either difficult to manufacture or difficult to operate under practical conditions. They also often regain their electrical resistance in the presence of strong magnetic fields or when carrying large currents. Since one of the main applications for superconductors would be in the creation of powerful fields, the drawback has proved a major obstacle to commercialisation.

Structure has tended to be increasingly complex variations on a theme of barium, bismuth, thallium, copper and oxygen. There are dozens of them and more than one commentator has observed cynically that for anyone who is not a specialist in this field, it is difficult to keep up with the vast volume of published literature.

But in *Nature* (Vol 362, No 6417), Sergei Putilin at Moscow State University, together with French and American colleagues, describes a mercury-based superconductor that could well lift the flagging spirits of superconductor research. There are suggestions, too, that it might provide a way of raising the present superconducting temperature record.

Composite superconductors have been developed that are significantly more workable and more resistant to the effects of magnetic fields. The superconducting



Structure of $\text{HgBa}_2\text{CuO}_4$, with the large circles representing Ba, medium representing Hg and small, O atoms. Cu atoms are inside the octahedra.

temperature, or T_c , has also slowly crept up to the -146°C record currently held by a superconductor containing two layers of thallium oxide sandwiched between three layers of copper oxide.

But the new material just developed by

Putilin *et al* contains a single layer of mercury oxide (HgO) sandwiched between copper oxide layers. As with so many discoveries of this type, it was arrived at partly by calculation and partly by trial and error. Over a rather crackly phone line from Moscow, Putilin said: "I tried to predict the approximate structure of the compound and then to synthesise it. I varied the conditions and one of the synthesised compounds gave a positive result."

In this instance, positive means that the new material is much easier to make than existing thallium-based superconductors and has superior electrical characteristics. Early indications also suggest that because of its compact crystal structure, the new superconductor should be capable of carrying more current and operating in higher magnetic fields.

Other researchers in the field are now suggesting that if the new mercury-based superconductor were to be synthesised as a double or triple sandwich like the thallium ones, it might easily set new records in terms of operating temperature.

The new material is clearly still a long way away from becoming the ideal superconductor – cheap, easily worked and operating at room-temperature. But it has certainly opened up a new avenue in what was becoming a highly congested and slow-moving area of research.

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

Lithium power under the bonnet

A practical electric car seems to have been always "five years away", with the reasoning being that some new battery development "just needs perfecting". Perhaps at last the lithium battery really could make that timescale a reality.

Lead acid cells are getting better, but they are still extremely heavy; nickel-based cells have problems of cost and toxicity, while sodium/sulphur batteries need temperatures of around 350°C to work efficiently.

The lithium battery offers hope, though the technical problems of developing a high energy density rechargeable version have been immense. Lithium is a highly reactive metal, placing considerable restraints on the electrolyte material, and substances which are chemically compatible usually have far too much electrical resistance to be satisfactory in a high power battery.

Other potential electrolyte materials

undergo physical changes – such as volume alterations – which would soon destroy the integrity of a cell. The ideal electrolyte for a rechargeable lithium battery would have high ionic conductivity, be chemically compatible with lithium and at the same time be flexible enough to survive the physical rigours of numerous charge/discharge cycles.

Unfortunately, of the various theoretical options for an electrolyte, lithium salts dissolved in organic liquids carry the risk of leakage and fire, while glasses and ceramics containing lithium ions are inherently brittle. Elastomeric polymers would be ideal, though up to now have not possessed sufficient conductivity at room temperature.

But progress looks to have been made (*Nature*, Vol 362, No 6416) by a group based in the Department of Chemistry at Arizona State University. C A Angell and

his colleagues report the development of an electrolyte that combines the ionic conductivity of ceramic electrolytes with the flexibility of polymeric ones. What they have done is to dissolve polymers – polyethylene oxide and polypropylene oxide – in a mixture of low melting point lithium salts (acetate, perchlorate and chloride etc). The result is a new class of rubbery materials that are good electrical conductors, even at room temperature.

This "salt-in-polymer" system is still a long way from being the lightweight practical vehicle battery for which everyone has been waiting – there are still important questions relating to safety and durability. But the ability to create a flexible ionically conductive electrolyte that works at room temperature in a lithium system is an important step forward.

Perhaps the all-electric car really is only five years away.



An integrated audio amplifier

Towards the end of the 1970's, following a period when I had been involved in the creation of increasingly complex transistor audio amplifiers, I was asked to consider a minimalist approach to design. The resulting amplifier was to be about 30W output, which would be fairly easy to build and free from problems.

To avoid the additional complications incurred by the need for speaker protection, I opted for a single supply-line system with a capacitor coupled output. This layout also allowed for a slow ramp-up on output DC voltage lessening the normal 'switch-on' plop.

The entire system, with the exception of a transformer/rectifier plus reservoir capacitor PSU, would be held on two PCBs, one for the preamp. and one for the power amp.

The resulting design was eventually published in *Hi-Fi News* in January 1980, since I felt that there might be others who would welcome a pleasant sounding DIY circuit which was not too complicated to build. Predictably, this led to it becoming a lot more widely known and used.

The problem now was that the design criteria for the original system as specified did not

aim at providing the ultimate performance. In the event, the design which I suggested was substantially better than anything which could be bought from the average high street shop.

In the hands of the golden ears reviewers, it was immediately compared with the best units which were around, at five to ten times the price. Not surprisingly, some of these were thought to be a bit better. This comparison was, perhaps, a bit unfair, since it was not a comparison between equals. But I knew that the original design could be persuaded to deliver more performance by altering the output stage. This was based on monolithic Darlington power transistors in the original design. Replacing these with mosfets would certainly improve this section.

Mosfets in the output

Because power mosfets have a very much higher transition frequency than normal power transistors, their use allows the negative feedback compensation characteristics of the circuit to be designed around a unity gain frequency which is ten times higher than the equivalent value for bipolar power transistors.

Moreover, one can do this without having to

forego the desirable single pole (-6dB/octave)* type of compensation system, with feedback applied over the whole gain block, which makes such a contribution to the quality of the transient response, and makes amplifiers which use this type of HF compensation so restful to listen to.

In addition, mosfets are free from a variety of problems which affect bipolar devices such as hole storage, a shortcoming which can impair the audio performance of conventional transistors when they are used at high current, as is the case in output stages.

I therefore evolved a modified design, at the end of 1980, which used power mosfets in the output stages. This mod gave a noticeable increase in the transparency of the sound, particularly in the middle to upper frequency range; this puts the sound quality some way ahead of its bipolar transistor rivals. It also ends to output transistor failures.

The only remaining disadvantage of the original simple amplifier was rather less weight and solidity in the bass through the use of loudspeaker coupling capacitors. This deficiency would only be noticed by comparison with the best there was, and then only when

The audio amplifier presented here by John Linsley Hood has evolved directly from his much respected original design which made its first appearance in the late Seventies. Using a combination of mosfets and other modern components, it preserves the minimalist approach which contributed much to the excellent performance of its ancestor.

the LS units were of very high quality, and the size of the listening room was large enough to accommodate the LF pressure waves. The answer was straightforward – to use a direct coupled design, and a stabilised power supply, to give really rigid bipolar supply lines, but this, of course, would make it rather more expensive to build.

Re-organising the power amplifier layout to operate as a direct coupled system between a bipolar pair of power supply lines did not, in itself, present any great difficulty, and the availability of a $\pm 15V$ pair of lines for the preamp would greatly assist design in this part of the circuit.

Component choice

My feeling about the use of ICs in low power audio applications is that, in unity gain stages, there is no audible difference whatever between an IC op-amp and a carefully thought out discrete component equivalent circuit – provided that one doesn't ask it to give too much gain, and one uses a decent audio quality device. In this context, there are some very good ones now available in the convenient dual device packages, such as the *LF353*, the *TL052*, the *NE5532*, the *SSM2139*, or the recently introduced *LM833*, which is possibly the best of them all. All of these are dual amplifiers; one need not worry about cross talk in using the two halves in parallel parts of a stereo chain.

I feel that high stage gain amplifiers still require discrete components to deliver the ultimate in sound quality. One pays for this small benefit with greater circuit complexity, and added chance of constructional error.

A practical compromise is therefore to use ICs only in unity or low gain level positions, or to organise the circuitry, as, for example, in an RIAA equalising stage, so that the gain requirement is divided as evenly as possible between two halves of an op-amp package, and both parts operate at a gain level which is low enough to avoid any detectable differences. These basic design concepts were adopted for the new design.

The basic circuit layout of the new amplifier is shown in Fig. 1, and owes a lot in its thinking to an earlier 80W design in that I have arranged the layout to allow the tone controls or rumble filter to be switched out of circuit when they are not needed, and generally to introduce as few elements as possible

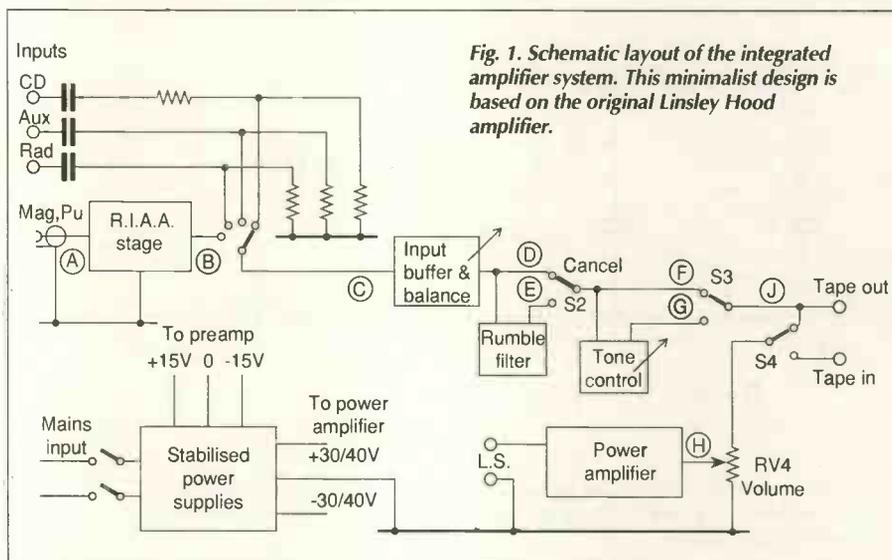


Fig. 1. Schematic layout of the integrated amplifier system. This minimalist design is based on the original Linsley Hood amplifier.

into the signal path from the input sockets to the input to the power amplifier.

Indeed, the only constant part of the preamp circuit is the input buffer stage which operates at a normal gain of 2. It provides channel balance facilities, and matches the output impedance of some of the auxiliary units to the low source impedance needed at the inputs of the tone controls, rumble filter and power amplifier.

The preamplifier

RIAA stage. I originally intended to use an IC version of an earlier design, but experience with a system based on an input impedance conversion stage, followed by a shunt feedback RIAA equalisation stage has convinced me that this topology is preferable. The final circuit is shown in Fig. 2.

In order to cater for the use of both moving magnet or low output moving coil cartridges, I have used a low noise IC type (*IC₁*) as a flat frequency response input gain stage, followed by a fet input IC, (*IC₂*) as a shunt feedback RIAA stage, with the frequency response correction obtained by *R₇/R₈* and *C₅/C₆/C₇*.

Used in this way, the noise background associated with either series or shunt feedback systems is very low, and this removes the major reason for the use of series feedback networks.

Preference for the shunt feedback type of equalisation circuit, where practicable, is due

to the difference between these two systems in the gain and phase characteristics of the upper (1kHz - 20kHz) part of the RIAA curve. The almost universally employed series feedback system cannot, reproduce the upper part of the required frequency response curve completely accurately without the use of additional gain correction networks: the system ultimately reverts to a unity gain stage as the frequency increases, whereas the RIAA specification calls for network gain to decrease to zero.

The performance difference between shunt and series feedback RIAA gain block can be seen quite clearly on an oscilloscope, and there are, I believe, audible sound differences between designs in which the 1-20kHz part of the curve is correctly equalised, and those in which it is not.

In addition, by virtue of its nature, a virtual earth shunt equalisation layout offers a degree of protection from possible input overload due to common mode input voltage excursions of the kind which can happen very easily with series feedback circuits on vinyl disc replay.

(*Footnote. It is never possible to achieve a pure single pole type of feedback loop stabilisation, which would offer a true 90° margin of stability at the unity gain point, because both stray capacitances, and the semiconductors themselves – especially the output devices, will introduce unwanted phase shift effects, which will lessen the phase margin. For this reason, it is desirable to tailor the HF roll-off so that the loop gain is adequately low before these effects occur).

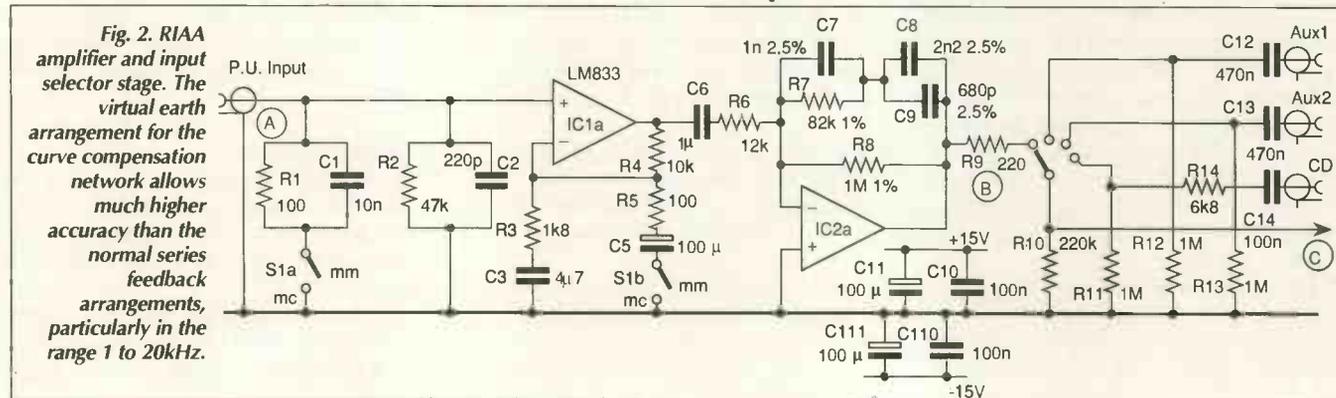


Fig. 2. RIAA amplifier and input selector stage. The virtual earth arrangement for the curve compensation network allows much higher accuracy than the normal series feedback arrangements, particularly in the range 1 to 20kHz.

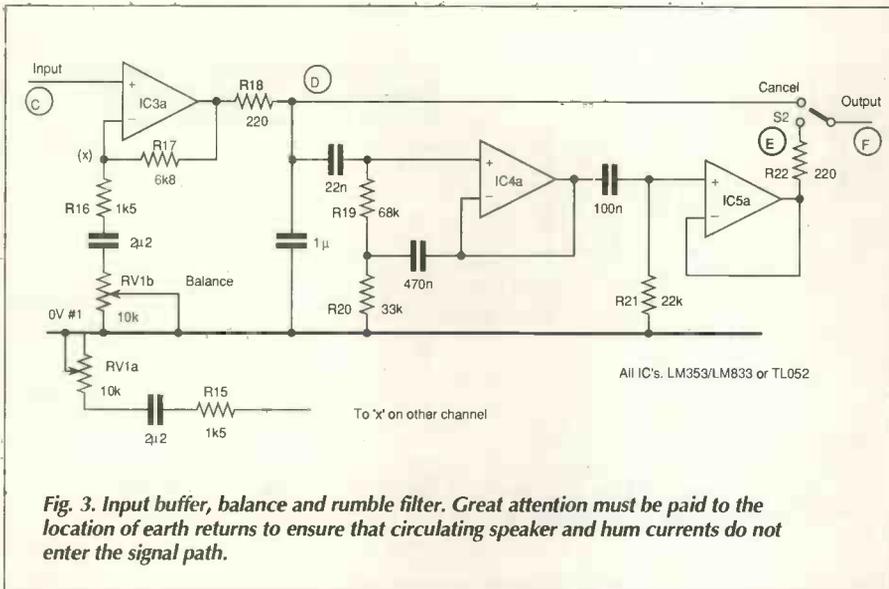


Fig. 3. Input buffer, balance and rumble filter. Great attention must be paid to the location of earth returns to ensure that circulating speaker and hum currents do not enter the signal path.

Physical design considerations

Since the design of the preamplifier allows the rumble filter and tone controls to be switched out of circuit, it follows that, if not required, they may be omitted altogether. Similarly, if the user only wishes to use CD, tape or radio tuner inputs, the RIAA stage can also be omitted leaving only the input switching, buffer, gain and balance controls between the input sockets and the power amplifier input.

As with the original prototype, it is suggested that all the preamp circuitry, with controls and switching components, and all the power amp circuitry, with heatsinks, should be collected on two separate PCBs for ease of testing. It would be prudent to set aside some space on the preamp PCB to accommodate any add on boards for optional stages not originally included in case they are required later. A further PCB will, of course, be needed to accommodate the components used in the stabilised power supply system.

A toroidal mains transformer should be used since the RIAA stage, in its moving coil input sensitivity position, will be sensitive to hum pick up from stray fields; the lower external mains field from a toroidal transformer will help to keep the overall hum background to a low level.

As with all audio amps, great care should be taken with the layout of the earthing system. Remember that the peak currents which flow between the mains transformer, the rectifiers and reservoir capacitors may be in the hundreds of amperes region. Audibly significant voltages will be generated if any part of them is allowed to flow between points in the signal earth line.

Adopt the normal practice of choosing some single, central earthing point to which all these heavy current wires – from transformer centre taps, reservoir capacitor 0V tags, LS return terminals, and the output line of the power amplifier – can be taken individually.

In the case of a single box power amp/preamp combination, it is preferable that the metal case should be earthed to the preamp circuit at the earthy side of the pickup input terminals, and nowhere else, so the central earthing point should be isolated from the chassis. The preamp 0V line should be connected to the signal input sockets at its input end, and to the common earthing post at its output.

The power amplifier input 0V line, 0V₁ (see Fig.5), should be connected for preference to the output 0V line of the preamp and the bottom ends of the gain control pots. The constructor should be prepared to do a little experimenting (switching off power between times) to see which of the preamp 0V connection layouts gives the lowest background hum level.

Care should also be taken that all the input signal leads are screened, and that their paths should keep them well away from the mains transformer and power amp outputs. It should also be a matter of common sense to ensure that the amplifier input sockets and the speaker output and mains input connectors are kept well apart.

I hope that Hart Electronic Kits, Ltd., of Penylan Mill, Oswestry, Shropshire phone 0691-652894 will provide a kit for this design, which will simplify construction, and avoid possible problems in circuit layout.

These voltage peaks can arise as a result of surface scratches, or dust particles in the groove, producing instantaneous relatively high output voltages from the p.u. cartridge. If an input buffer stage is used, care should be taken to ensure that its response is fast enough and its gain low enough, to avoid merely referring the problem one stage further back.

Input switching. The required input drive voltage for the power amplifier stage is in the range 570 - 40mV rms depending on the chosen output power rating and, since the input buffer stage has a gain of 2 at the centre setting of the balance control, this means that the aux input signal levels required for maximum output power will be in the range 280 - 370mV. I have provided a separate, lower sensitivity, input for CD players to reduce their normal 2V rms output to a level which is more convenient in use.

Input buffer stage. This is as shown in Fig. 3. With the balance control set in the middle position, the gain of this stage is approximately two, as noted above, and the lower -3dB point in the frequency response curve is 11Hz. The balance control allows a gain adjustment of ±10dB. The maximum input signal which can be accepted at the mid position of the balance control is 5V rms.

Rumble filter. This is arranged to have a turn-over frequency of 33Hz and an attenuation slope below this frequency of -22dB/octave. Above the cut-off frequency, the frequency response is flat, and the stage has unity gain. As in the other stages, I have arranged the two halves of the op. amp. so that they handle the equivalent portions of the signal in each channel, because this makes for a slightly tidier PCB layout.

Tone control module. This is shown in Fig.4, and is of a normal Baxandall type layout, and allows about ±18dB lift/cut at 40Hz and 20kHz w.r.t. 1kHz, at maximum or minimum settings of the controls. With the Treble and Bass control potentiometers set in their half-way positions, the frequency response of this module is flat, and it has unity gain. As with the other stages, it can be switched out of circuit when not required.

The output to the power amplifier is taken directly from either the input buffer stage via the gain control, RV₃, or the output from a tape recorder or from the outputs of the rumble filter or tone control module, all of which have a typical output impedance of about 250Ω.

Power amplifier

The circuit of this is shown in Fig. 5, and is of a fairly conventional layout, apart from one or two innovations, borrowed from my earlier class A/AB 35-40W design, (EW+WW March 1989).

An input long-tailed pair stage, TR₁ and TR₄, is fed from a 550μA constant current source, (TR₂/TR₃), and has an output DC offset adjust-

ing potentiometer, RV_1 , in its emitter circuit.

The input signal is taken from the volume control via a DC blocking capacitor, C_1 , and an input HF roll-off network, (R_2/C_3), which attenuates the input signal beyond 100kHz, and helps ensure freedom from slew-rate limiting effects. The gain and symmetry of the input stage is improved by the use of a current mirror circuit, (Tr_2/Tr_5), as its collector load, and the output from this is used to drive an mos gain stage, Tr_7 , which again has a 15mA constant current source as its drain load.

This configuration is very linear and is used to drive a push-pull pair of Hitachi mosfets, Tr_{10}/Tr_{11} . HF stabilisation of the whole feedback system is provided by R_9 and C_8 , which is in a position which does not lead to slew-rate limiting.

The output mosfets are sufficiently robust that adequate overload protection is given by the input diode/zener chain, $D_1/D_2/ZD_2/ZD_3$, and the current limit in the stabilised power supplies. The quiescent current for the output stages should be set to about 100mA (it is not at all critical) by means of RV_2 .

Performance

The general behaviour of the amplifier is fully up to the best of contemporary expectations, with a good transient response, free from overshoot or ripple, apart from a small amount with pure capacitive loads, due to L_1 , which is included to avoid any possible unwanted effects if the amp. is coupled to LS cables with a high self capacitance: power mosfets tend to oscillate at HF when used as source followers to drive capacitive loads.

Over the range 100Hz - 10kHz, the harmonic distortion is well below 0.01% at all power levels below the clipping point, and the whole design seems free from slew-rate lim-

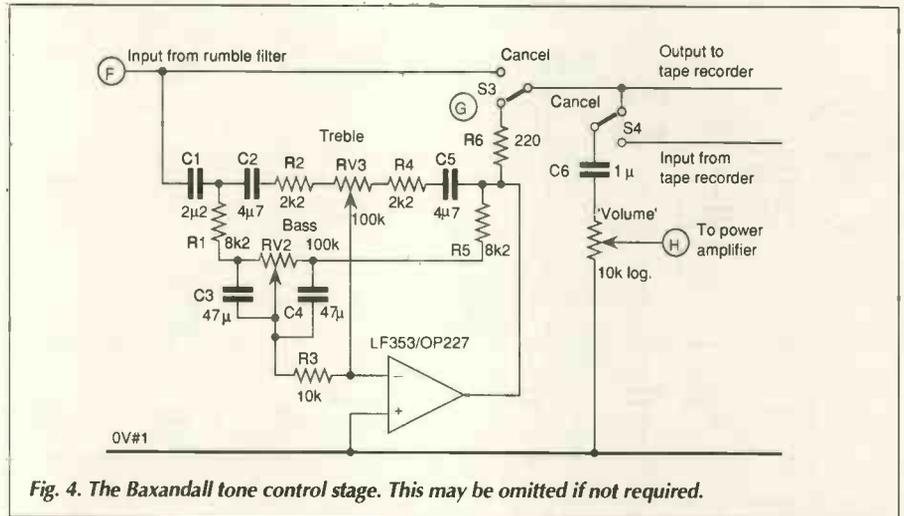


Fig. 4. The Baxandall tone control stage. This may be omitted if not required.

iting, or transient intermodulation distortion. The absence of TID makes an important contribution to the comfort of listening to any amplifier for a lengthy period of time.

As an aside, I think it is a sad commentary on reviewing practice that hi-fi writers often give ecstatic reports on new equipment which presents an immediately exciting sound, when experience should have taught that it often conceals, or is an aspect of, performance deficiencies which eventually grate on the listener's ears.

Power supplies.

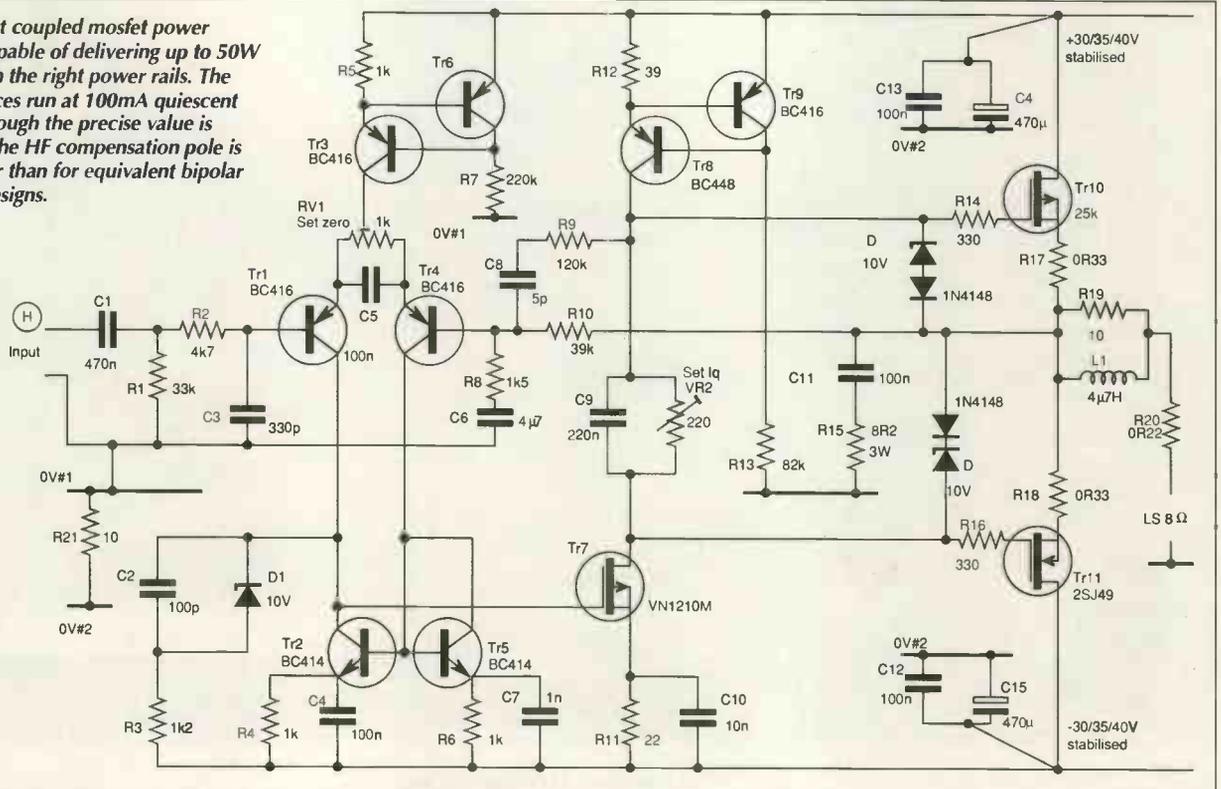
The rigidity, and freedom from ripple and other rubbish, in the power supply lines is an important factor in the sound quality of any audio amplifier at the bottom end of the audio frequency band. Because most of the young golden eared fraternity, whose views dominate the pages of hi-fi magazines form their sonic

judgments on rock music, the assessed sound quality of hi-fi amplifiers or any other audio gear is heavily loaded towards vocal and string bass reproduction. This quality depends mainly on the low impedance of the power supply lines, and a stabilised supply is a more cost-effective way of achieving this than the use of massive supply line reservoir capacitors.

The circuit I have used for the power supplies is shown in Fig. 6. IC_1 , as the basic control element. R_2 forms an output load for this, and causes it to draw current, (up to 10mA), from the pass transistor, Tr_2 , via a buffer transistor, Tr_3 , which is used to isolate the input of IC_1 from the output of the power supply, which could exceed IC_1 's maximum input voltage rating.

If the current drawn from the pass transistor causes the output voltage to fall, more current is drawn from Tr_1 base than if the system

Fig. 5. Direct coupled mosfet power amplifier capable of delivering up to 50W into 8Ω with the right power rails. The output devices run at 100mA quiescent current although the precise value is uncritical. The HF compensation pole is much higher than for equivalent bipolar transistor designs.



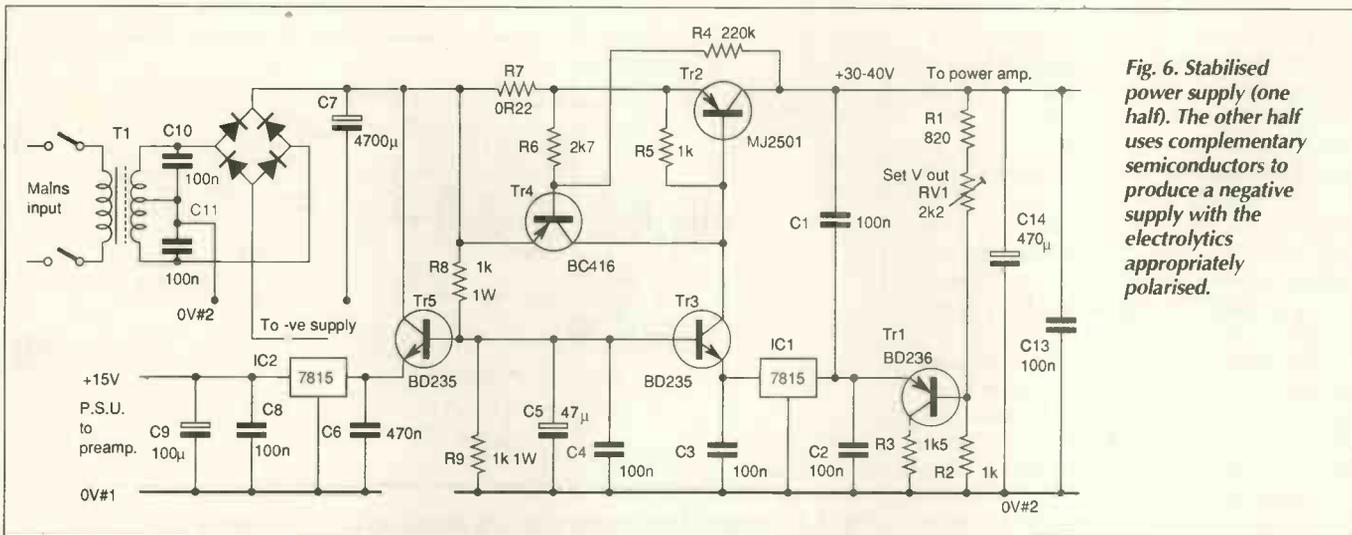


Fig. 6. Stabilised power supply (one half). The other half uses complementary semiconductors to produce a negative supply with the electrolytics appropriately polarised.

operates with the $Tr_1/IC_1/Tr_3/Tr_2$ chain in equilibrium, and Tr_2 is turned on harder to supply this extra current.

The current limiting function is performed by Tr_4 and R_7 , in that if the current flow through R_7 causes the voltage developed across Tr_4 base to exceed some 0.55V, then Tr_4 will be turned on and will steal the base current from Tr_2 . This provides an absolute current limit of about 3.8 amperes from the supply, when there is no significant voltage drop across Tr_2 .

However, if the output voltage falls (increasing the voltage drop across Tr_2), then an increasing proportion of that voltage drop will appear on Tr_4 base, via R_4 and R_6 , which reduces the total output current, to keep within Tr_2 's safe operating area. This re-entrant power supply characteristic will offer some measure of speaker protection without the need for output fuses.

I have only shown one half of the bipolar supply system since the other half is identical, with the exception that IC_1 and IC_2 are 7915

and the transistors, Tr_1-Tr_5 , are NPN types rather than PNP and vice-versa.

The specification of the transformer, and the voltage ratings of the reservoir capacitors, will depend on the required output power and I have shown some suitable options in a small table at the top of Fig. 6. The transistors specified will serve for all of these power options, though one should make sure that the heat-sinking on Tr_2 remains adequate at higher power levels.

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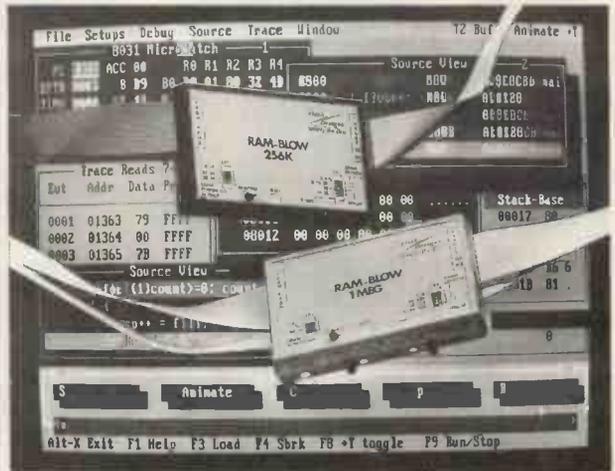
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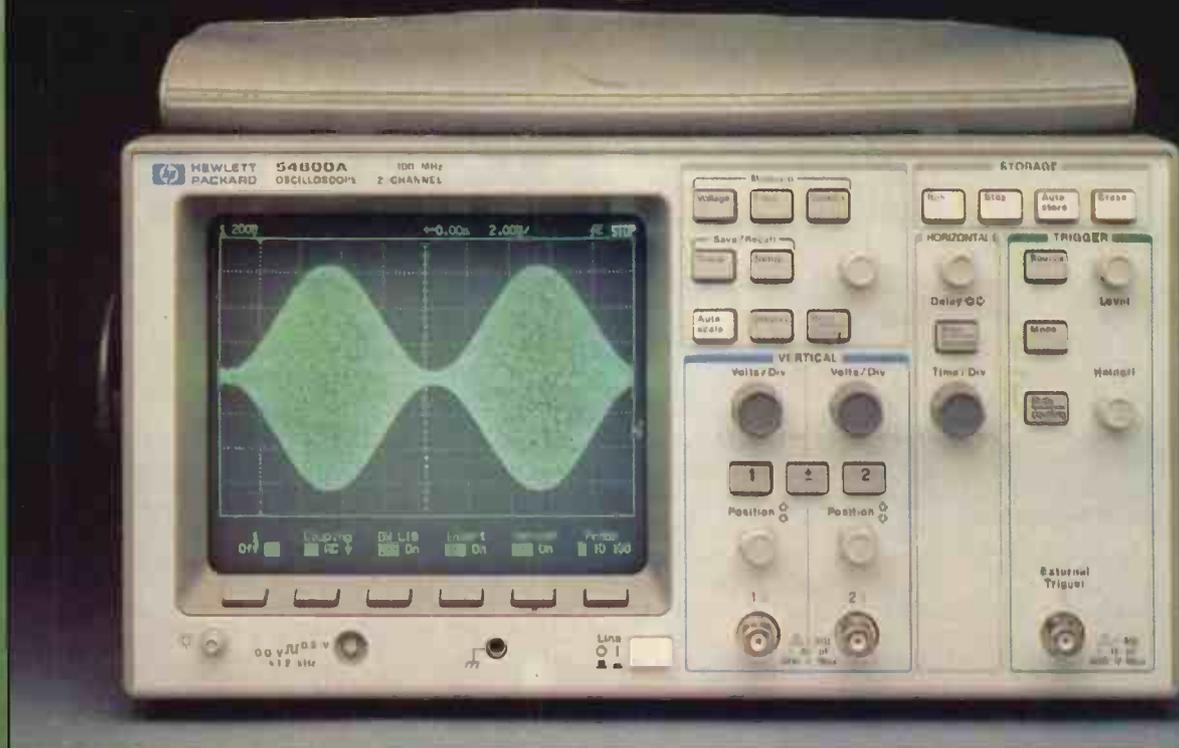


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461

8051 DEVELOPMENT EASED BY C

System requirements

PC compatible (AT or above recommended)
640K
Spare half size PC card slot
5.25/3.5in floppy
Hard disc recommended
Mono display (CGA/EGA/VGA/SVGA recommended)
Mouse recommended

Flash Designs 8051 emulator and C development system.

Development systems range from the basic and cheap blowing of an eeprom and testing it – suitable only for the simplest systems – to advanced and expensive hardware ice (in-circuit emulation). At this expensive end, the microprocessor chip is temporarily replaced by hardware running exactly like the real processor, but allowing full access to registers and bus lines for the development process.

Between these two extremes is a third option, accessing the system through the memory chips rather than the processor. This is the solution offered by Flash Designs (and also by Trace Technology, see December *EW + WW*). The memory system contains not just the program under development, but an extra one used solely during the development phase. Once finished the extra code is not needed, and the eeprom can be blown with just the finished program. In the Flash Designs product the debug routine is very small – around 250 bytes – and the solution provides most of the benefits of a complete hardware ice, but at a much lower cost.

C/8051 combination

Combining the C language and the popular 8051 single chip controller is a popular option. But a knowledge of the 8051

Flash Designs universal emulator package provides a software development system for a range of processors – including the 8051 – at an entry level price of £400, says Len Freeman.

assembly language is still desirable when it comes to debugging, as one C instruction may convert to a dozen assembly language instructions. To see what is happening requires a dip into the actual language of the processor. Beginners are well advised to start with the development system built around just the assembler – currently being included free of charge by Flash Designs.

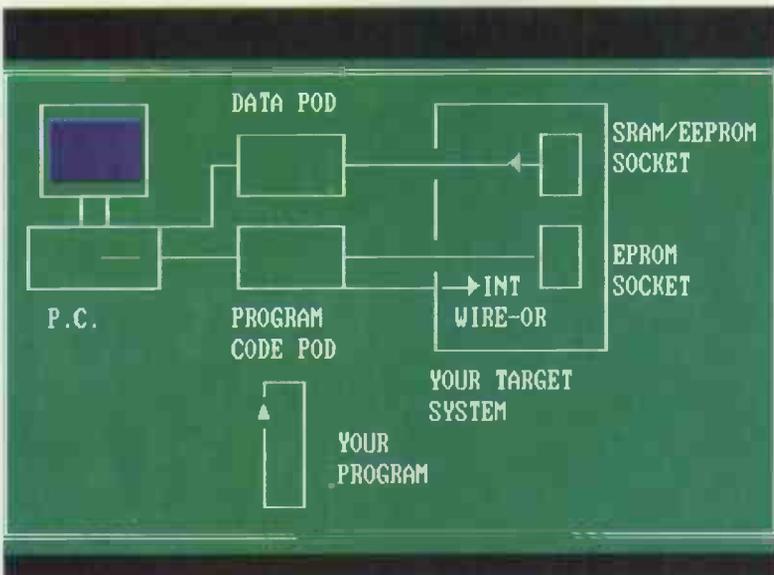
The C-IAR compiler, available at additional cost, supports all C data types, which can also be viewed and edited “on the fly”. Flash Designs also provides support for other compilers such as Keil.

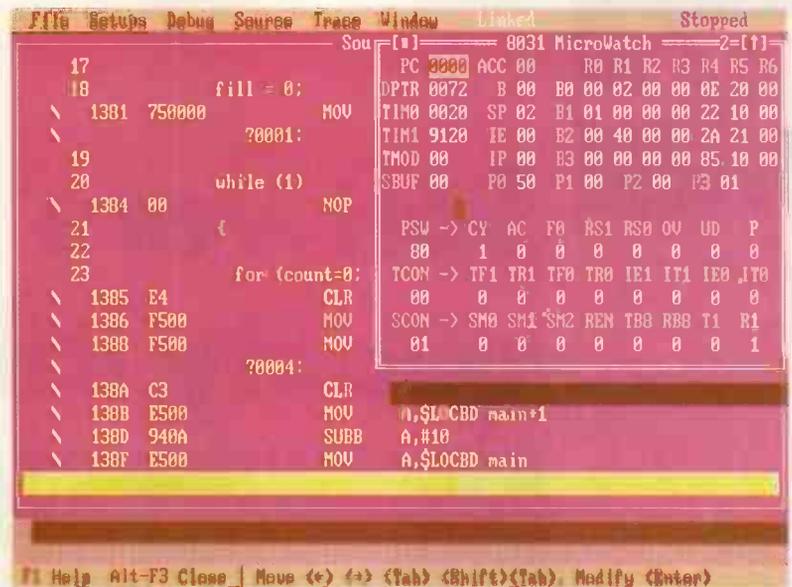
Hardware consists of a plug in PC controller card, with ribbon cable to the ram emulator pods. In turn, these are connected to the memory sockets of the target board. The pod (or pods if both code and data emulation is chosen), are stackable with a footprint of only 9cm by 14cm. The required 5V power supply is taken from the target board.

Programs and data can then be down-loaded/up-loaded from the PC into the ram pods using direct memory access, making this a very fast process. Most of the tools needed for software development are available, such as setting break points, single stepping and a display of all processor registers. The other main feature is that operation can be carried out while simultaneously viewing the source file – in assembly language, in mixed C and assembler, or in C, to provide source code debugging.

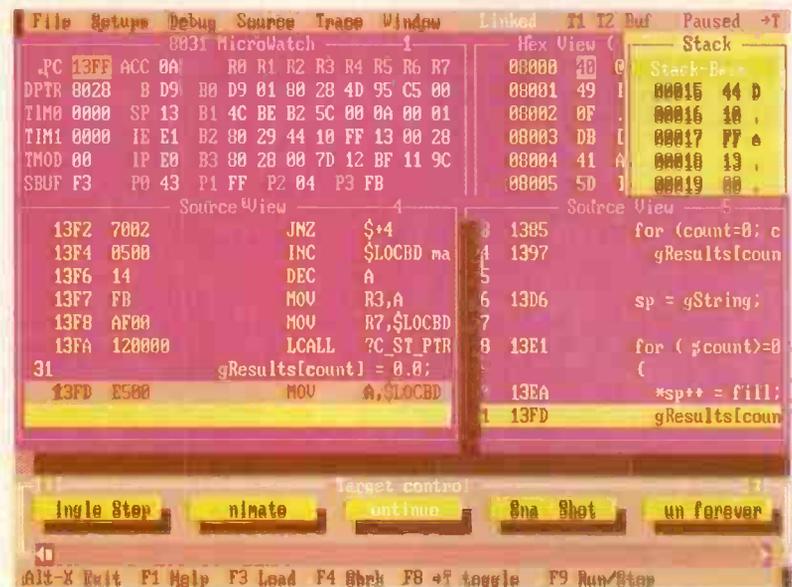
Turbo-Trace

The software is menu-driven, with a window type display,





Microwatch window at top right displays all registers. In the background a source file is shown, in this case a mixed C and assembler file.



Options displayed under the title "Target Control" when a break point is encountered. Also shown are sections of the two source files, one in the original C and the other in mixed C and assembler.

Supplier details

Beginners starter pack: ram-blow 256K £349 (1Mbyte £489 + VAT) includes ram-blow emulator, PC emulator controller card, 4Mbyte prommer, universal AMICE software. Microwatch requires Mwatch personality.

Starter pack plus: Starter pack plus TurboTrace £798 (£938 1Mbyte).

Professional : Starter pack plus TurboTrace and second ram-blow emulator pod, 2 x 256K £1028 (256K + 1Mbyte £1148, 2 x 1Mbyte £1268).

Flash Designs Ltd, PO Box 167, Crawley, West Sussex RH11 9YE. Tel/Fax: 0293 551229

and the mouse is used in the usual way, moving or resizing windows for example. A "Microwatch" window shows all registers, and probably the most useful source file to view in the background behind this window is a mixed C and assembler file.

The software will run under dos, or Windows 3 where multi-tasking is allowed, though the run-speed is then somewhat slower. Basic version of the system allows only for software break points. An optional "TurboTrace" allows hardware break points to be set as well, and these seem more versatile and easier to use. But they should not be set immediately after JUMP-type instructions. When a hardware break point is reached, there are several ways of continuing:

- Single stepping (Microwatch is updated);
- animate – a slow motion mode which allows registers to be observed while a "tracker bar" highlights the relevant instruction in the source file;
- continue until the next breakpoint, and
- snapshot, where the program runs at full speed until it hits the break point again, when it gives a system "snapshot" and continually repeats. Break points can be toggled on and off from the keyboard, provided program execution is stopped first, and options are automatically displayed at the bottom of the screen under the title "Target Control" when a break point is encountered. TurboTrace also has four logic probes, allowing the state of individual lines to be recorded before or after a breakpoint is encountered (as well as the data and address bus lines).

Processor type can be changed for only £25 – excluding the compiler of course – a great advantage for organisations, such as colleges who may work with several different processors

More pros than cons

The product is a very useful, cost effective package, representing a quantum jump forward from the simple "burn and test" system. Disadvantages centre mainly around the learning curve for the package, compounded by a manual, which while comprehensive is not very user friendly.

A more comprehensible manual would be a great help, particularly for first time users who might be keen to move on to a development system such as this. Experienced users will probably not find the manual a problem.

The help menu is good, though fairly basic, and is context sensitive.

Occasionally, difficulties can be experienced in using it, and the only solution seems to be to come out of the system and start again. The odd telephone call to Flash Designs might be necessary (it was for me) though the help needed was readily available.

Overall, the criticisms certainly do not outweigh the great benefits provided by such a versatile and cost effective development system. ■

Software with the quality label

Effective quality management relies on good analysis of process data. Don Bradbury finds Northwest Analytical's QA charting and analysis package has quality running all the way through it

System requirements

PC-XT, AT, PS/2 or compatible
CGA, Hercules, EGA or VGA
Dos 3.2 or later
512K ram
3.5in or 5.25in floppy
Hard disk
Compatible with OS/2 and Windows

QA's aim is to help identify process variation and determine its cause.

Adoption of total quality management (TQM) by manufacturing industry is irresistible and irreversible. Version 4.2 of Northwest Analytical's *Quality Analyst (QA)* sets out to crunch TQM data down into various process efficiency indices, backed up by a drawer full of charting and statistical techniques. *QA*'s aim is to help identify process variation and determine its cause – the very essence of quality control – a goal well achieved by an excellent performance on almost every level.

Those experienced in the field, with an established understanding of quality control and statistical concepts, will quickly grasp the operation essentials of this heavyweight program. Users without such knowledge might need a course in the basics. To help them, UK agent Adept Scientific runs periodic seminars on statistical process control using *QA* to demonstrate the principles.

Admittedly, at £550 plus £10 postage and 17.5% VAT, *QA* potential purchasers will need a firm justification for its acquisition. But even this price does not put it among the more expensive examples of software in its class. Unfortunately, the market is relatively small and niche, so giveaway prices or bulk-sale discounts are not very likely.

Menu or command operation

A main menu gives direct access to creation and editing of files, variable control charts, attribute (defect) control charts, graphs and reports, statistical analysis, file manipulation, run file execution, program set up and utilities, and the import and export of data from other files. Function keys, described along the bottom of the screen, can also be used to run certain activities.

In menu mode, general types of function are presented for selection from submenus. But more experienced users will tend to favour the command mode, even though routine names have to be remembered to use it, as it gives access to all routines including some not listed in menus. (There is also

Protracted installation

Installation is rather protracted, though for those familiar with PC configuration, the process need take only about ten minutes.

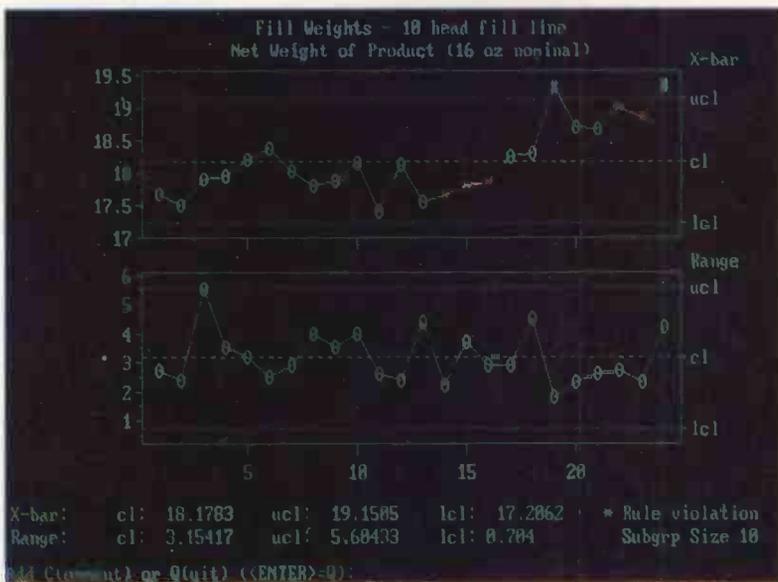
Modular design allows a wider range of problem solving than is possible with a rigidly structured system, and this module approach is reflected in the 105 files occupying about 4Mbyte of disk space that automatically unpack from archives during installation – DAT files also included contain excellent tutorial material.

A fair amount of hardware may need to be configured by selecting various option and this can be a little daunting. But a short cut is to accept the default options and then modify the configuration later if necessary.

If any of essentials are omitted, specifically the video and printer and/or plotter types, the program prompts for answers it needs before allowing the user to proceed. More detailed configuration involves setting dos environment variables, the required hard copy mode, and additional device definitions including colour modifications, chart size and fill patterns, font selection, multipliers.

It is also possible to have automatic wrapping of long output so that the width of a graph can be set to a value greater than the physical width of the paper.

Where continuous paper is used, output is then wrapped onto the next page.



Heavyweight manuals

QA is big in every sense. Even the smallest of the three manuals, *Program set up and configuration*, contains no less than fifty eight A4 pages. As all the manuals, it is very detailed, with many screen dumps and much repetition of instructions to keep new users on the right track.

Another exhaustive manual takes the first time user through a set of tutorials and exercises to master the options for quality control charting and analysis. They also show how production and modification of a process capability report is achieved, to create a histogram with a normal, folded, or skewed probability distribution, with mean, process capability, target, and specification limits and index options.

Tutorials also cover the control of printer output, setting up control limits and evaluating potential process capability.

Data transformation is prominent in QA, and along with such processes as new data creation from existing variables, methods for grouping individual measurements, creating a short-run control chart, and the use of median and median/individual charts, is well covered.

a batch mode, see "Creating run files".) Modes can be switched at will.

Prompt-and-response operation always gives a default option which is the most likely choice depending on previous operations. But once all options are understood a first letter mode of response speeds progress.

A point to note is that the row/column data entry routine must leave an "even" file structure because data files are essentially text files. Another is that restrictions on naming variables include a length of between two and eight unique alphanumeric characters which a few users might find a little limiting.

Two data editors are included. One allows data cells to be ordered in a spreadsheet-like array, while the other is line-oriented. Other data manipulating facilities include file merge, extracting row and/or column data, sorting data, using a conversion utility from single column to multiple column or subgroup form, and a report generator that takes a QA data file and produces a layout in a more formal format.

Data manipulation

Data subsets can be created based on user defined criteria and the package also offers a random number generator and subgroup convertor outputting average, range, and standard deviation.

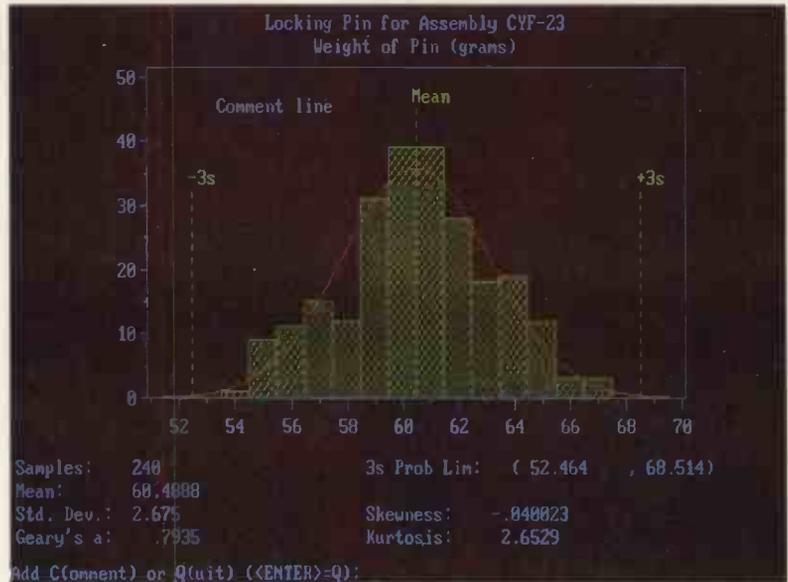
Choice of control charts is wide and will be more than enough for most people, while graphics facilities allow scatter plots of data sets to be added – even data from different files.

Defining a data window for charting enables QA to produce a rolling progression. So that, for example, a day's data can be added while simultaneously dropping off the first day's data in a file, producing a graph of a fixed number of points.

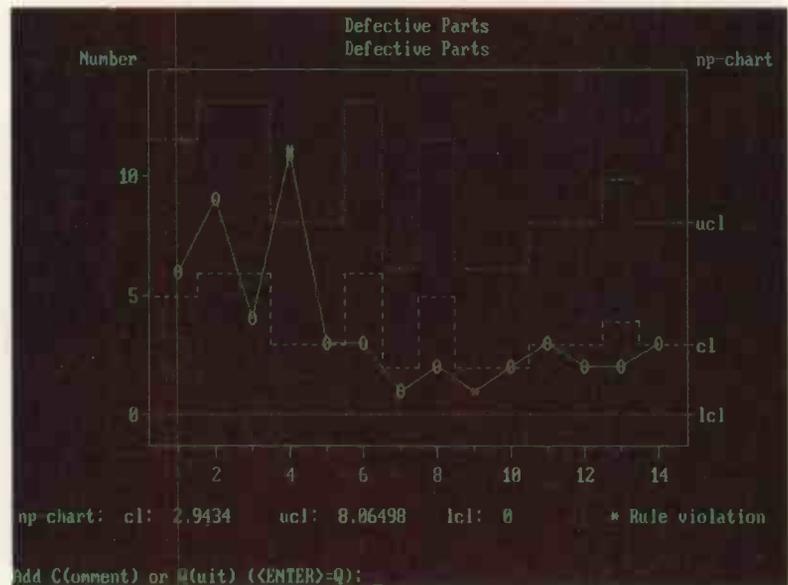
Presentation can be brightened up by using the graphics editor to place and modify comments, add titles and footers, or change a scale in a chart.

Sensibly, statistical analyses functions are targeted squarely at the quality analyst and will surely cover most analysis needs in this department.

The substantial data processing required, even with quite modest data sets, means QA does not run with conspicuous rapidity on a machine such as a 16MHz 286 PC (though my



The choice of charts should cover all the needs of the quality analyst.



np-chart showing number of units defective.

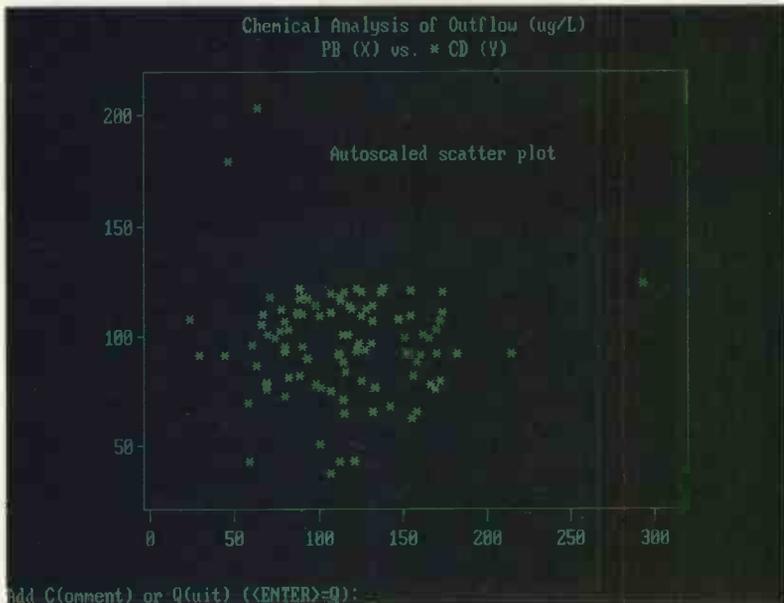
File conversion

Most users will not be interested in file conversion. Those that do need it, will probably want to interface QA with a spreadsheet and QA has no problem converting data for the common ones.

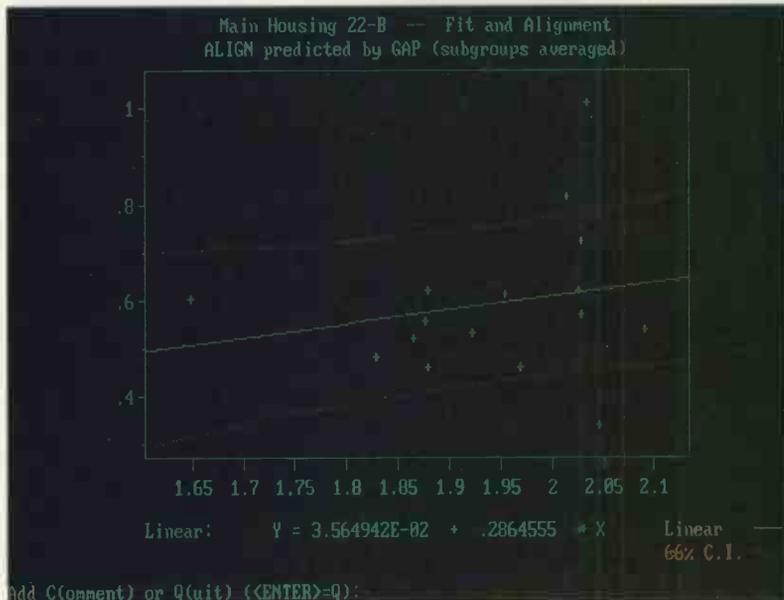
For example many existing analysts use the package with Lotus 1-2-3 data files, or tack QA output on to their current information management program. These practices are some of the reasons why the text data file format was chosen for QA.

But other external ascii files can be handled too. Word processor output, database managers, or mainframe down-loads can be cleaned up by one of QA's file conversion utilities. Other utilities convert between QA data files and DIF format or SYLK files, and Datatype data-only-report files and Mitutoyo DL-10 data logger information can be converted to QA format. Data files can also be rotated through 90°, turning rows into columns and visa versa.

There are also graphics file conversion utilities handling encapsulated postscript, HPGL (Hewlett Packard Graphics Language), and Lotus PIC files.



Autoscaled scatter plot.



Linear regression with user-specified confidence limits.

Invicta 25MHz 486SX proved itself more than adequate to the task). All modules run more efficiently with higher processing power, and a fair amount of hard disk activity will test the patience of users with limited hardware performance.

No competition?...

At first glance there seems to be a lot of competitors to QA. For example many have tried – and some still do – to make a spreadsheet or standard commercial graphing program such as the otherwise excellent Lotus 1-2-3 or Statgraphics do the job. But QA is much better aligned to the task, and the best of the competition generally comes from programs costing up to twice as much.

PQ Systems' SQCPack, Third Millennium's Inpro-SPC, and Lighthouse Systems' SPC-Lite come to mind. But where QA scores is in its friendly user interface and its facility for customisation – points which have rated highly in the list of favourable comments from users.

Northwest Analytical has plainly been in the SPC (statistical process control) business long enough to know the needs of its customers, and the six versions of QA produced

since the early eighties have steadily improved the reputation of the product.

Now it looks unlikely to disappoint, though special situations can bring particular needs, so check the alternatives if you are in the SPC software market. Look around and you might find a few companies who will supply demo disks, but this generally applies to programs with much wider appeal. ■

Creating run files

Batch mode operation allows creation of run files, which may include rem statements to explain what the programmer is intending, executed from a menu or from a command line prompt. Batching can considerably speed up repetitious work, so that such factors as the number of plots per page and how they are placed, for example, can be defined.

The list of supported output devices is extensive and fully documented in the set up and configuration guide.

Charting and analyses capability

Control charts include:

X-bar, range, and standard deviation; p (percentage defects), np (number defective), u (non-conformities per unit), and c (control chart for non-conformities); individual measurement, median, and range control charts; exponentially weighted moving average, range, and standard deviation; and cumsum, autoscaled run, and out-of-control summaries. Other graphics add scatter plots of data sets and can even include data from different files.

Statistical analyses include:

analysis of variance, linear correlation, multiple regression, polynomial regression, and single variable regression for linear, exponential, logarithmic, and power equations. Chitest, calculations for the Kruskal-Wallis statistic, Sumstat for computing descriptive statistics such as mean and standard deviation, several types of t-statistic for making comparisons, the Weibull probability plot and analysis of life test routine, as well as the Wald-Wolfowitz run statistic for evaluating the tendency for data to form runs. All these add substantially to the functionality of the program.

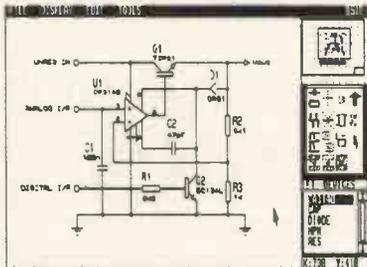
Supplier Details

Quality Analyst (QA) £550 + £10 postage + VAT is available from Adept Scientific Micro Systems Ltd, 6 Business Centre West, Avenue One, Letchworth, Herts, SG6 2HB. Tel: 0462 480055 Fax: 0462 480213.

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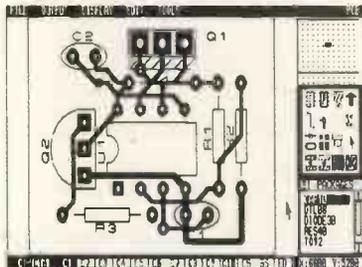
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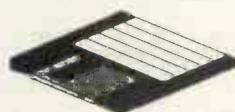
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A lost listener

Your news item about the coverage of scientific matters by the BBC (*EW + WW*, May 1993) couldn't be more apt at this time because this sad state of affairs was thrown into sharp relief quite recently for me, since as a music lover and an early refugee from our "new and improved Radio 3", I have found a comfortable haven in German satellite radio.

Last month, the AES held its convention in Berlin and one of the events was a seminar to commemorate the development of stereo tape recording by German radio 50 years ago. The two hour discussion with copious and impressive demonstrations from archive tapes (including a complete Beethoven Emperor concerto recorded in 1944) was broadcast.

I received it via a German receiver, from a digital transmission on one of the 16 regional and national network stations accessible from TV-Sat 2 in near-CD quality. Up to eight will be transmitting classical music at any one time, from excellent live sources from all over Europe and fine in-house digital tapes. And no analogue compression, the latest iniquity to degrade standards on BBC Radio 3.

Can one possibly imagine the BBC ever doing anything like this? I think not. There, too, we have a true reflection of the status of the engineer in this country, which is a national disgrace.

Reg Williamson
Whitehill Staffordshire

E=Zzzzzz²

Am I the only reader who is getting bored with the correspondence about the validity of Einstein's theories?

There are plenty of other exciting topics out there waiting to be tackled. How about the nature of the electron (or the photon); are they particles or waves? Do sidebands really exist? Can we trust quantum mechanics? What is the life expectancy of Schrodinger's cat, as judged by (a) Schrodinger and (b) the cat? Has anyone seen any N-rays, polywater or phlogiston recently? Is the earth flat?

Come on, give Einstein's theories a rest.

John Ball
Woodridge Suffolk

Quantum leaps over the top

The claims made for quantum cryptography (*EW + WW*, Research Notes, May 1993) are more than a little over the top.

While actual transmission would have the security to be expected of an optical fibre, the code itself is actually a variant of the well-known Enigma code, conveniently ready prepared in binary. It also has the same structure as the genetic code of which, as we all know, the human variant is in the process of being decrypted in its entirety.

Quantum spin geometrical dissection is a much more secure code in that it has very high redundancy combinatoric complexities. The codes for individual alphanumerics are also intermixed and the resulting partitioned encryptions can be further randomised. Resolution of the enormous combinatoric complexity of a message of more than minimal length would be beyond any computer since the transmitted version would, to all intents and purposes, be returned to a random binary state.

A simple version consists of first assigning specified 4 x 4 matrix codes to the alphanumeric and other symbols it is required to transmit. An 8 from 16 code would give 40,320 (factorial 8) permutations, of which only 256, say, would be used as symbol codes.

Messages are initially constructed by assembling symbol matrices edge to edge as a contiguous tiling in the plane, then subjected to geometrical dissection by partitioning into four 2 x 2 matrices. These corner matrices are then exchanged, under a rotational key system, with those of adjacent 4 x 4 matrices. This operation only requires rotations of 2 x 2 elements as complete blocks.

The partitioned elements can, after re-assembly, be further exchanged under different nested scalings such that the final transmission is completely randomised when transmitted in serial form. The key can be constructed from unused 4 x 4 symbol codes and hidden in the transmission or transmitted in another message. Dataflow methods can be used for decryption.

Brian Clement
Crickhowell Powys

War or money?

I was a little shocked by Frank Ogden's comment: "The nature of power" (*EW + WW*, March 1993) about the Balkan war. His opinion is that the rest of Europe should stay out of it if there is no money in it for us!

To me this is an extremely narrow-minded way of thinking. In the Balkans, more than 200,000 have been killed so far, and the Serbs are acting more and more aggressively. It is the worst situation in Europe since World War II regarding brutality and torture.

If human rights are supposed to have any meaning, the international community should react to such a situation, by military intervention if necessary.

If the conflict were to spread to Turkey, Greece, and Bulgaria, we would have a major conflict on the doorsteps of Europe. From this neither Frank Ogden, Europe, nor the rest of the world would benefit or get any money. There would just be more and more human suffering.

Stephan H Jensen
Helsingor
Denmark

Drawing the line

With regards to your comment "Right to eavesdrop?" (*EW + WW*, April 1993), you may think that "official surveillance systems are, on balance, a good thing in a democracy". You should know that if the technology exists officially then there is a high probability that it will be available unofficially.

When people use what they think are reasonable secure systems, the results of unofficial surveillance can be costly or somewhat embarrassing – remember Camillagate.

Your motoring analogy is also weak – number plates are very easy to change. The line between official and Big Brother is narrow and anything that can reduce the potential power of state over citizen, or citizen over citizen should not be dismissed.

It is probable that the rules on telephone tapping do not apply at the border interface. When I call the UK are my conversations monitored and my faxes printed? This faxed letter may have been read by someone we don't know, even before it is on your desk. Doesn't that make you uncomfortable? It does me.

Mark Plews
Hersching
Germany

Why don't pigs fly?

Arenon Goldberg's letter, "Did the Earth Move?", (*EW + WW*, April 1993) suggests that light is slowing down; starlight's red shift is not due to the Doppler effect; the universe is not expanding; it is only a few thousand years old; and the earth is at the centre of the universe.

This reminds me of another outbreak of vernal equinoctial correspondence of a few years ago when it was proposed that the aeronautical way to the stars was to be by an air breathing rocket launcher.

On studying this latter proposal in detail it turned out to be totally impracticable. To explain why this

is so, one can make an analogy and transpose air to water. It is like asking a naval architect to design a submarine to travel underwater at 460knots so that it can leap like a salmon out into the air to an altitude of 1.6km!

Euan Orr
Portchester
Hants

Sounds like a great idea to me - Ed.

Catt's challenge

A new initiative is needed to break the stalemate in electromagnetic theory. I am probably the best known name in this field, but nobody with accreditation in the

subject will admit to having heard or read my theories, or comment favourably or unfavourably on my theories or competence.

In particular, nobody with accreditation in electromagnetic theory will admit to having heard of, or comment on, the Catt Anomaly (*EW + WW*, September 1987), on which I rest my case.

As a result, the question of whether text books and college courses should be modified cannot be addressed.

This is why I am setting up the Classical Electromagnetic Theory Defence Fund as an attempt to break the log-jam which has persisted for more than a decade. I am donating £100 to the fund.

The idea is to pay three accredited experts in electromagnetic theory if and when they attend a meeting where they defend the classical theory against the Catt Anomaly. The meeting will be tape recorded and each defender given a copy of the tape. They will be paid whether or not they succeed in their defence. Critics of classical electromagnetism will not be paid.

Before the meeting, suitable institutions such as the IEE and Institute of Physics will be asked to authorise the three experts as appropriate defenders of the classical theory. Donors to the fund will also have a say in who the experts are.

If, as I suspect, nobody with proper accreditation can be found to defend against the Catt Anomaly and the fund gradually builds up over the years, then classical electrodynamics will take its proper place alongside poltergeists and other idiocies.

Ivor Catt
St Albans
Hertfordshire

Diabolical waves

On the subject of light, Martin W Berner's letter (*EW + WW*, January) repeated the often-asked question "Waves of what?". A substantially correct answer was given by James Clerk Maxwell in 1865: "Light is an electromagnetic [em] disturbance propagated through the em field according to em laws."

In the wake of more than a century's developments, I brought Maxwell's theory up-to-date in 1983 in "On Maxwell's Ether" published in *Lettere al Nuovo Cimento*. This paper suffers from the drawback that it was written in such a way as not to offend too much of the post-1905 sensitivities (otherwise it would have not been published). Nevertheless, it clarifies some of the fundamental issues.

Einstein maintained that relativity

did not overthrow the classical theories (of Newton and Maxwell), but instead generalised and improved on them. That this is incorrect is demonstrated thus: if an experimenter measures the speed of light and obtains the value c (about 300,000km/s) with respect to the source, one would be entitled to say that this is a particle behaviour. If an experimenter measures the speed of light and obtains the value c with respect to a medium of propagation, one would be entitled to say that this is a wave behaviour. But if one postulates that c may be referred to any and every coordinate system,

that would be a truly diabolical behaviour!

T Theocharis
London

A case of blurred vision

Martin W Berner asks in "Read shift" (*EW + WW*, January 1993), for an explanation of the cosmical red shift Doppler effect in terms of the invariant, relativistic velocity of light basic concept.

The sad message is that no explanation seems to exist on this basis. And, from a lack of reasonable alternatives, cosmologists have introduced imaginary expanding effects of space and the universe.

However, from a clear physical point of view, there may exist two possible reasons for a Doppler effect to appear: a true Doppler effect if a light ether exists; or a false Doppler effect if light is pure particles.

In the first case the true Doppler effect is associated with an active light ether medium.

If $w\lambda$ = wavelength as produced on

Entering the SpiceAge

Your reviewer John Anderson did an excellent job on our product *SpiceAge for Windows* (*EW + WW*, April 1993). The few criticisms that he made have nearly all been made independently by our users and have now been dealt with.

He is quite correct in pointing out the continuous nature of the program's development and indeed the current version of *SpiceAge for Windows* is 2.059 with the manual now on its third release. Because all our customers have been updated free of charge "from birth" right through to version 2.027 (the penultimate release), it has been necessary to ensure that the manual paper costs are kept reasonably low.

To answer a specific criticism, the component ribbon at the foot of the edit window has its prime purpose in accessing the hypertext help system for that component and it is not intended as a means for entering a component. To use a component's key

letter for help implies that very knowledge and rather defeats the purpose.

Also, the edit window in v2.059 can now be placed on top of the analysis windows.

The scaleable fonts of *Windows 3.1* can result in very small text when windows are contracted but are always clear on full screen windows. The alternative, which Graham Baxter removed in response to user's requests from earlier versions, is to disable the text when it becomes smaller than a threshold size.

A point that Baxter and I would like emphasised is that *SpiceAge* is not *Spice*. Baxter built it from the ground up on entirely original code. The compatibility with *Spice* is provided for convenience but *SpiceAge* is entirely free of the constraints of *Spice* and will develop accordingly.

CJT Clarke
Those Engineers
London

Share...

The product review, "Ranger 2: the shareware worth the layout?", (*EW + WW*, April 1993) misapplies the term "shareware". True shareware comprises the complete program and documentation, which is not the case with this version of *Ranger 2*.

Shareware is one of the most interesting phenomena in all of personal computing. Seetrix is not the first marketer to attempt to change the shareware concept to suit its own marketing plans. I doubt it will be the last.

Therefore, I am very careful about which program gets called "shareware" in *EDN*. MicroSim has always had the good grace to call the reduced capacity version of *pSpice* their "educational" version. Another acceptable term is "demo" version.

Charles H Small
EDN Magazine
Newton
Massachusetts
USA

Editor's note: The whole point of shareware is to give people a chance to try out a working version of a program before deciding whether or not to buy the full version. If shareware always contained the complete program and documentation, there would be no incentive for users to pay up and register. That is why most shareware has key features, such as the ability to save, disabled, or alternatively has annoying "please register" messages popping up at inconvenient times.

...where?

First of all, thank you for your generous review of my software *CC4 Calculus Calculator* (*EW + WW*, March 1993). Your readers should know, however, that this software is not supported as shareware.

Three years ago I sold my work to Prentice Hall, but I forgot to tell my beta testers not to distribute their copies. Some of them released the product to their local bulletin boards thinking that it was still shareware. From there it has spread as far as the UK..

I don't mind if it circulates this way, since I gain more in publicity than I lose in sales, but the free version is not the final one.

The final version is sold as *X(Plore)*, and it comes with a 250 page manual from Prentice Hall. The cost in the US is around \$30.

Your readers should also know that Prentice Hall sells a version of the manual without the disk, and they can avoid purchasing that version by specifying the ISBN number (0-13-014226-3) or demanding the book with disk.

A Macintosh version of *X(Plore)* will appear in January.

David Meredith
Department of Mathematics
San Francisco State University
USA

the transmitter side with the receiver at rest, $w_r =$ wavelength as measured on the receiver side with the transmitter and receiver moving, $w_0 =$ the wavelength as measured when the transmitter and receiver are at rest, $f_t =$ frequency of the transmitter $= 1/dt$, $c =$ propagation velocity of wave in ether relative to the ether, $v_t =$ velocity of the transmitter related to the ether, $v_r =$ velocity of the receiver related to the ether, and $dt =$ time interval between two transmitted pulses:

At transmitter side

$$w_t = dt \cdot c - dt \cdot v_t = w_0 / c \cdot (c - v_t)$$

At receiver side

$$w_r = w_t$$

giving

$$w_r = w_0 \cdot (1 - v_r/c)$$

Frequency at the receiver side

$$1/f_r = w_r / (c + v_r)$$

$$f_r = (1 + v_r/c) / w_r$$

$$f_r = (1/w_0) \cdot (1 + v_r/c) / (1 - v_r/c)$$

$$f_r = f_0 \cdot (1 + v_r/c) / (1 - v_r/c)$$

Hence for the ether wave:

$$w_r = w_0 \cdot (1 - v_r/c)$$

$$f_r = f_0 \cdot (1 + v_r/c) / (1 - v_r/c)$$

In the second case of light as a particle or matter wave, light is transported by pure particles. Velocities of those particles are determined by the output velocity at the source and the velocity as measured by an observer by adding the source velocity in relation to this observer.

For the transmitter side

$$w_t = d_t \cdot (c + v_t) - d_t \cdot v_t = d_t \cdot c = c/f_0 = \text{constant}$$

On the receiver side, the time for an interval to pass the receiver moving with velocity v_r in relation to a common reference (the same as for the transmitter) is:

$$t_r = w_t / (c + v_r - v_r) = 1/f_r$$

$$f_r = f_0 \cdot (1 + (v_r - v_t)/c)$$

For light as matter wave

$$w_r = w_t = c/f_0$$

$$f_r = f_0 \cdot (1 + v_r - v_t)/c$$

Hence, if light is a stream of particles, the Doppler shift is a false effect, giving only a shift in frequency.

The Doppler effect in the particle case is received because of an invariant in transmitting velocity in relation to the receiving system, hence making Einstein's concept useless. And because even the ether concept is inhibited in Einstein's theories, it is impossible to see any reasonable cause or reason to the cosmical red shift described in relative terms. Only the pure old Newtonian method seems to remain.

Ove Tedenstg

Märsta
Sweden

Clocking on

In answer to RMW Quaille's letter (EW + WW, February) on the French long wave station *France Inter*, the

station uses a very sophisticated microprocessor-driven clock which was described in the French version of *Elektor Electronic*. A kit is also available from Selectronic (Lille, Tel: 20 52 98 52 Fax: 20 52 12 04).

Signature unreadable
Carrieres
France

UK's stifled innovation

Phillip Darrington's article "Pursuing a lost course" on Captain Heinz Lipschutz (EW + WW, May 1992) and the ensuing correspondence highlights a problem faced in this country by many inventors and theoreticians. It is a problem which needs to be seriously addressed by manufacturers and, I'm sorry to say, the paper committees of some professional journals.

I have written a number of papers in an attempt to transfer technology from the air transport to the shipping industry, where it was, and still is, very badly needed. Most of these appeared in the proceedings of the *Ship Control Systems Symposium* held in this country and abroad, although some appeared in the journals of the Institute of Navigation and the IERE.

But publication only came about after manufacturers in the UK, particularly in the shipping industry, rejected them for illogical and obscure reasons. In one case I am confident that the rejection was to protect a monopoly.

In every case my proposals were ultimately taken up, mainly by foreign competitors, with zero financial rewards to me.

I have found that patenting, unless you are prepared to manufacture yourself, is a waste of money. Patents can easily be broken. And some firms hire a small team to see how patents can be legally breached.

Paper committees on professional journals are little better. I served on one and saw excellent articles rejected because of a tendency to scorn work in an attempt to display one-upmanship. Some reviewers love to make cutting remarks scrawled in red ink across the submission.

This stifles innovation and has a knock on effect on industry. Obviously, some articles should be rejected, but the importance of unbiased reviews cannot be overstressed.

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Frequency synthesisers produce output signals whose frequency and stability is proportional to that of a reference oscillator. It may generate one frequency or many at exact intervals. Tuning is in discrete steps in contrast to that of LC or RC oscillators with analogue tuning elements. Step size may be as small as a few hertz or some fractions of hertz.

There are several independent ways of constructing a synthesiser, all of which have their advantages and disadvantages, but the PLL version is the cheapest and the most common method. Figure 1 shows a simplified, basic PLL frequency synthesiser which is the same as that analysed in the first part of this series. All the theoretical conclusions made earlier hold up for the examples presented here.

In the steady state $F_{out} = F_{cf} \times N/M$, where F_{cf} is the crystal reference frequency, N is the programmable division ratio between VCO and phase comparator, while M is the fixed division ratio between reference oscillator and phase comparator.

Prescalers

The system shown in Fig. 1 is, in principle, universal and valid for any frequency band, but there are two practical considerations; the speed of the programmable divider and the bandwidth of VCO retuning.

Nowadays, programmable divider speed may extend to 10GHz with the latest GaAs ICs, but there is a price in high output noise. ECL chips allow programmable divider design up to 500MHz, but price and power consumption are both high. Inexpensive 3GHz silicon fixed prescalers are available and exhibit good noise performance, reasonable prices and low power consumption. When working at UHF/SHF, the design shown in Fig. 2 is used, in which a prescaler with a fixed divider ratio K precedes the programmable divider proper. This makes the VCO frequency K times lower and makes for simpler design.

With a prescaler in the basic circuit of Fig. 1, the formula for output frequency will change to $F_{out} = F_{cf} \times N \times K/M$. Frequency spacing is K times wider and to regain the smaller spacing, the comparison frequency must be changed. One also should decrease the cutoff frequency of the loop filter and redesign the band-stop filter which rejects the pulses and the comparison frequency (if such a filter is used) although a smaller loop filter bandwidth inevitably increases switching time.

All the effects mentioned above are particularly severe when K is more than 10-20. It leads to this sort of situation. A programmable divider might have a maximum operating frequency of 10MHz while the VCO output frequency could be 1.5GHz. The prescaler has $K=150$ and if frequency spacing is, say, 10kHz, the comparison frequency turns out to be $10\text{kHz}/150=66.6\text{Hz}$ which is unacceptable for all sorts of reasons. The prime among them is that the loop filter would have to possess a very long time constant to deal with the low reference frequency ruling out the possibility of reducing sideband output noise through loop servo action.

Dual-modulus dividers

A simple circuit allows the use of a prescaler with the divider ratio K without changing the PLL comparison frequency. This is achieved with a dual-modulus divider, a counter with two possible division ratios, K or $K+1$ and a control input to determine which division ratio is being applied.

Figure 3 shows an arrangement for a DMD made up from two programmable dividers, N_1 and N_2 . In general, this circuit replaces the whole programmable divider shown in Fig. 2. N_1 and N_2 are presettable counters producing a borrow output when the state of the counter is zero and the RS trigger controls the divider

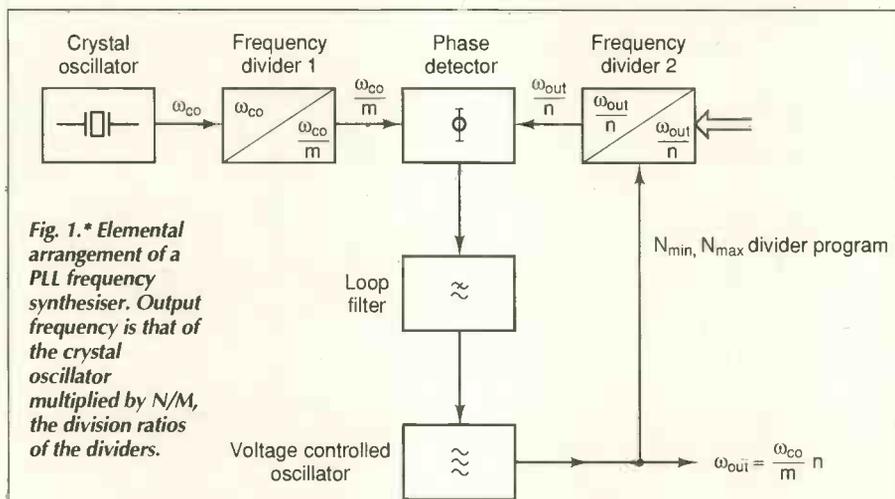


Fig. 1. * Elemental arrangement of a PLL frequency synthesiser. Output frequency is that of the crystal oscillator multiplied by N/M , the division ratios of the dividers.

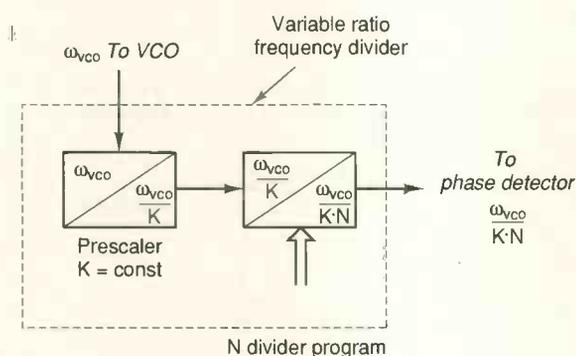


Fig. 2. * At high frequencies, prescalers before the voltage-controlled oscillator offer better noise performance than UHF GaAs dividers.

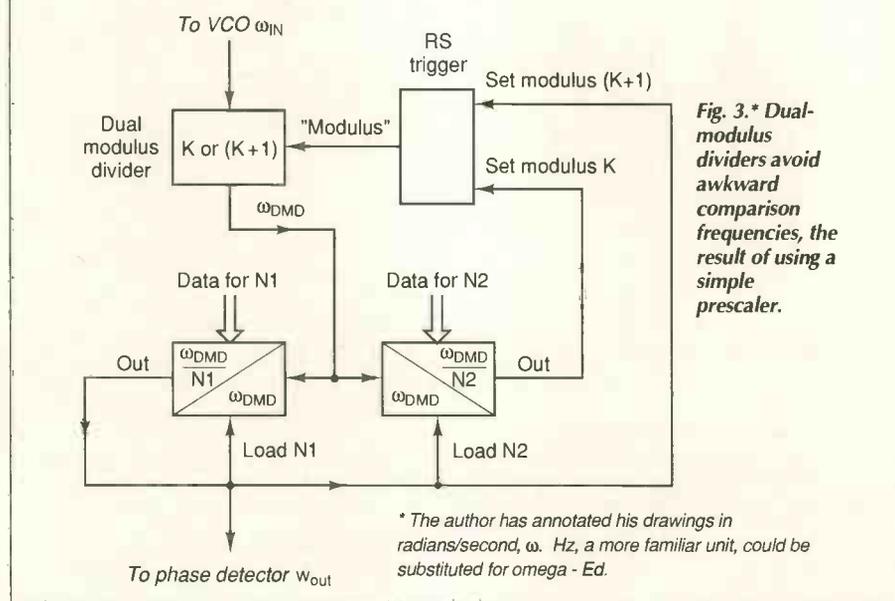


Fig. 3. * Dual-modulus dividers avoid awkward comparison frequencies, the result of using a simple prescaler.

ratio (modulus) of the DMD. Borrow output of the counter N_1 is the circuit output and sets the trigger so that the modulus is equal to $K+1$; counter N_2 sets the modulus equal to K . Assume the counters N_1 and N_2 be loaded with the coefficients N_1 and N_2 , N_1 being more than N_2 ; the trigger is set for modulus $K+1$ and the input is present. DMD divides the input frequency by $K+1$ and the output is $F_{in}/(K+1)$. When N_2 has counted down to zero, the pulse from its borrow output sets the

trigger to give modulus K ; the state of the counter N_1 having already decreased by the value N_2 . Counting continues, but the frequency at the DMD output is already F_{in}/K . When counter N_1 output becomes zero, signal from its borrow output sets up modulus $K+1$ and also loads the counters N_1 and N_2 . While N_2 is at zero, $N_2(K+1)$ input pulses are received, after which $N_1=N_1-N_2$. While N_1 is at zero, the input sees $K(N_1-N_2)$ pulses. The total number of input pulses for one output

Cardinal characteristics

Stability. Unstabilised crystal oscillators provide a basic stability of 50ppm which can be reduced to around 10ppm with thermal stabilisation. Selected, oven-controlled crystals can provide 1ppm of stability or better.

Frequency spacing. This depends on the application. For example, the channel selector of a television receiver usually has a frequency spacing of 62.5kHz, but taking into account frequency prescalers, the comparison frequency in such a PLL is $62.5/8=7.8125$ kHz. The output frequency spacing can be higher than the comparison frequency but the inverse is quite possible and it will be considered later.

Spectrum purity. This is extremely important, especially in synthesisers for communications sets and precision measuring circuits. Output noise may be divided into amplitude and phase or frequency modulation. The drive for spectrum purity requires extra circuitry and a considerable amount of calculation at the design stage.

One should not aim for a purer spectrum than is necessary in a specific case, or the design will be more expensive and circuit complexity will escalate.

For example, in a channel selector for a television receiver using a PLL as the local oscillator, the measured signal-to-noise ratio of the oscillator signal might be 54dB and the frequency stability no better than 10ppm. No synthesiser noise is visible on the screen or can be heard over the loudspeaker; quality is quite sufficient for this case. But the SNR and frequency stability allowable for a TV receiver would not be good enough for a TV transmitter.

Switching time of the output frequency – the retuning rate of the synthesiser – again depends on the application. In a communication receiver, for example, it must be less than 10ms to allow rapid frequency scanning when searching for and fine tuning on the signal. Switching time in a stationary broadcast transmitter could be a few seconds. It is important to remember that if requirements are too high, the design will always be expensive and complex.

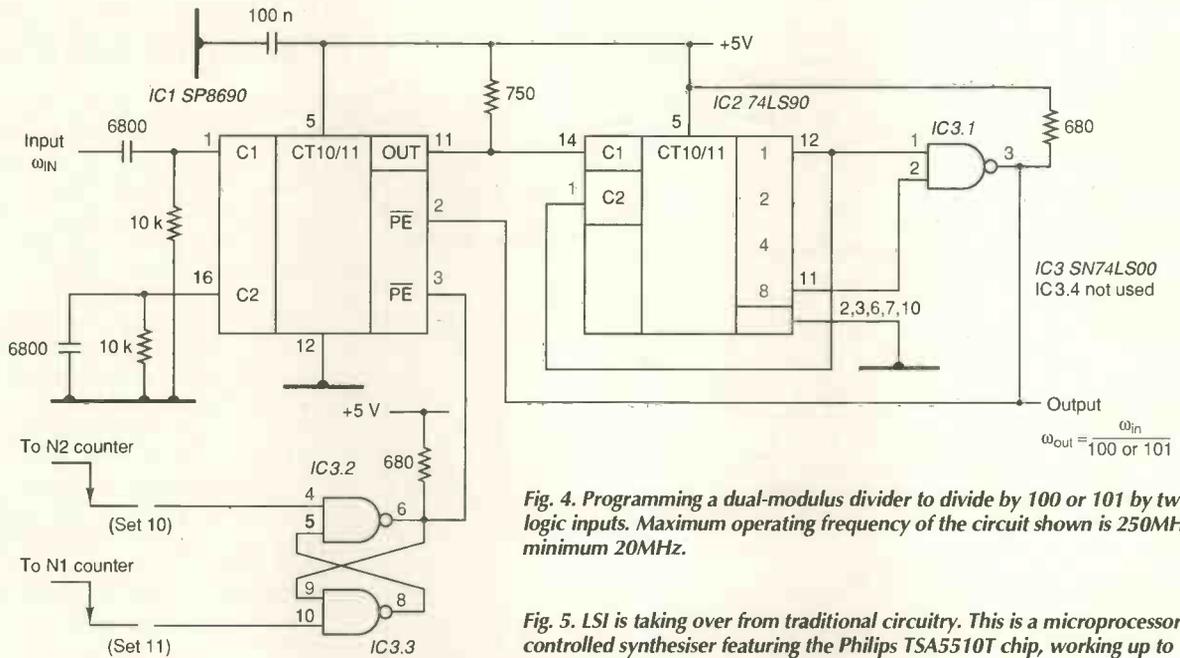
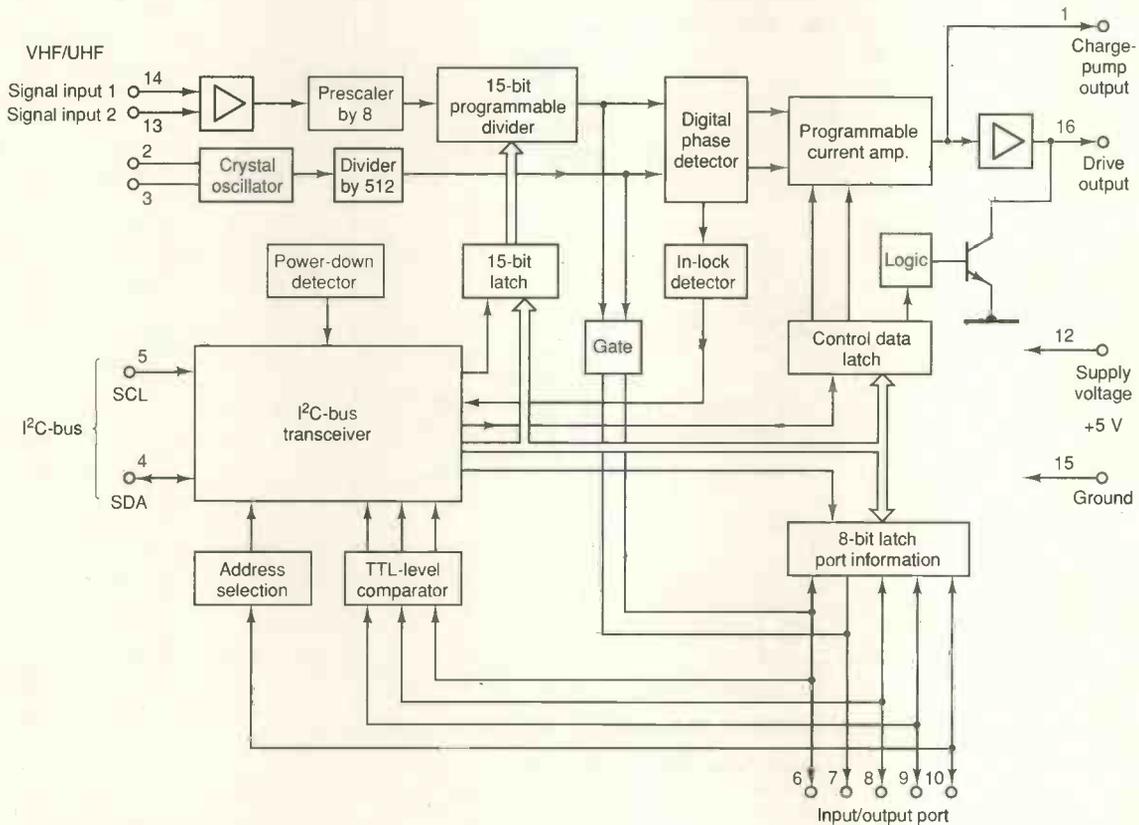


Fig. 4. Programming a dual-modulus divider to divide by 100 or 101 by two logic inputs. Maximum operating frequency of the circuit shown is 250MHz; minimum 20MHz.

Fig. 5. LSI is taking over from traditional circuitry. This is a microprocessor-controlled synthesiser featuring the Philips TSA5510T chip, working up to 1.3GHz or more with an external prescaler.



pulse is $N_2(K+1) + K(N_1-N_2) = KN_1+N_2$, the divider ratio in the circuit shown in Fig. 3; output frequency is $F_{out} = F_{in}/(KN_2+N_1)$. The result is that counters N_1 and N_2 operate at maximum input frequency, F_{in}/K , but the counter needs no change in comparison frequency in the PLL, as it did when using a prescaler only (Fig. 2).

The circuit, of course, becomes complicated, but justifiably so since it simplifies the circuit as a whole. But the technique has a drawback:

the minimum value of N_1 must be more than N_2 or the counter will be in trouble. For example, when $K=3/4$, $N_{2min}=0$, $N_{1min}=1$ and the total minimum divider ratio is equal $0+3 \times 1=3$; and when $K=100/101$ the minimum divider ratio grows to 100.

The maximum operating frequency of a dual modulus divider system without any other form of prescaling is currently about 1000MHz. Fixed dividers ahead of the DMD can currently extend reliable PLL synthesis to

around 8GHz at narrow channel spacing.

Note that UHF/SHF prescalers not only have an upper operating frequency, but also a lower one for satisfactory operation that may be more than 30MHz. Fast counters chips may malfunction at low input frequencies since their inputs, which often exhibit negative resistance characteristics at VHF, may oscillate with stray input inductance during long transition times. This causes a miscount in the output.

Figure 4 shows how to change the divider ratio of a DMD by adding a divide-by-10 counter and a few logic ICs. The SP8690 is a 10/11 DMD, working up to 500MHz; two inputs, PE₁ and PE₂, are used to program the divider ratio. Until the 74LS90 counter reaches state 9 (1001 in binary), there is logical 1 at pin 1 of IC₃ and the DMD is set to divide by 10. A 9 at the output of IC₂ means that 9x10=90 pulses have been received at the DMD input. After this, there is logical 0 at the input PE₁ and the 9 at the output of IC₃ changes to 0 in 11 input pulses if PE₂=0, or in 10 if PE₂=1. Total divider ratios are therefore 90+11=101 or 90+10=100. The trigger formed by IC_{3,2} and IC_{3,3} is controlled by the output of counters N₁ and N₂.

Chip systems

Frequency synthesisers using discrete components and the type of DMD shown in Fig. 4 have become a thing of the past: LSI ICs having virtually taken over. They contain practically all the components for a synthesiser except the VCO. In two distinct groups, there are low-frequency types with the two counters N₁ and N₂ that work up to 10MHz and need an additional DMD for the work on higher frequencies; the UHF/SHF variety, having a prescaler, DMD and N₁ and N₂ counters to work up to 1GHz and more.

Synthesiser chips without prescalers working up to 150MHz, are mainly intended for use in consumer electronics. Such ICs consume considerably less power and are therefore well suited to battery powered equipment. All currently available synthesiser ICs are designed to work with microcontrollers, the

CHOICE OF COMPARISON FREQUENCY

Any pulse residue from the phase detector breaking through the loop filter causes parasitic frequency modulation (of the VCO. Assume the amplitude of these pulses to be very small and the deviation of the VCO frequency F_{VCO} caused by them to be much less than the comparison frequency F_{cf} , or $F_{VCO}/F_{cf} \ll 1$, where F_{VCO}/F_{cf} is the frequency-modulation index (K_{fm}). The VCO output spectrum with narrow-band FM caused by the breakthrough pulses can be represented as

$$A_{vco} = A \cos \omega t + (A \times K_{fm} / 2) \cos(\omega t + F_{cf}) - (A \times K_{fm} / 2) \cos(\omega t - F_{cf})$$

where A is the amplitude of the carrier frequency, and $AK_{fm}/2$ is the amplitude of the alias components, the two appearing in the ratio

$$R_{ac} = (AK_{fm} / 2) / A = K_{fm} / 2 = \delta F_{vco} / 2F_{cf}$$

For a VCO with the slope S_{VCO} and a pulse breakthrough amplitude at the loop filter output V_p , the frequency deviation caused by the pulses is

$$\delta F_{vco} = V_p S_{VCO} R_{ac} = S_{VCO} V_p / 2F_{cf}$$

Taken together these indicate that when the PLL comparison frequency increases, other things vary as follows: The same filtering coefficients in the loop filter being equal, the amplitude of the alias components in the spectrum of the VCO output signal decreases; decreasing the VCO slope leads to a decrease in the frequency modulation index of the spurious pulses and results in a reduction in parasitic components in the synthesiser output signal.

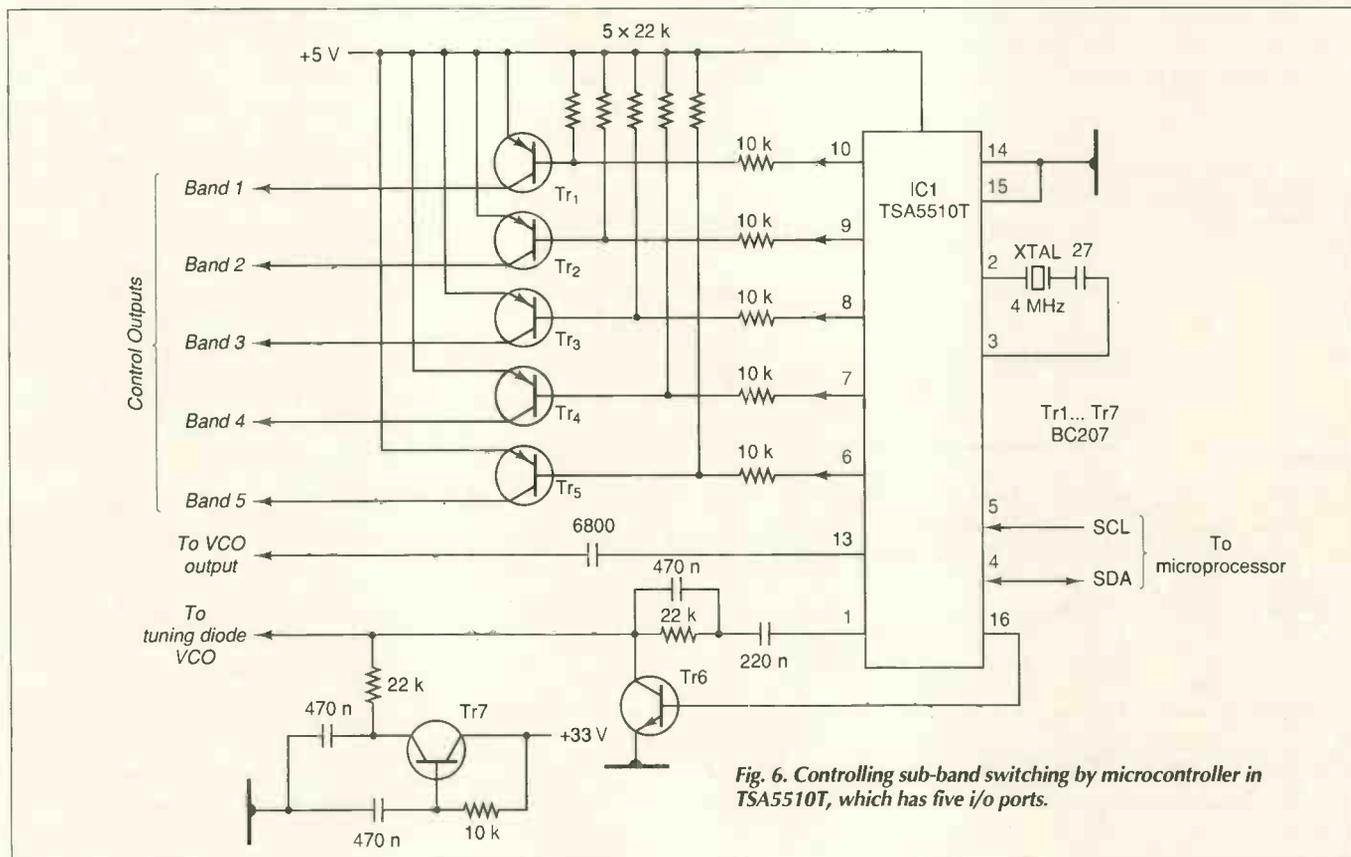


Fig. 6. Controlling sub-band switching by microcontroller in TSA5510T, which has five i/o ports.

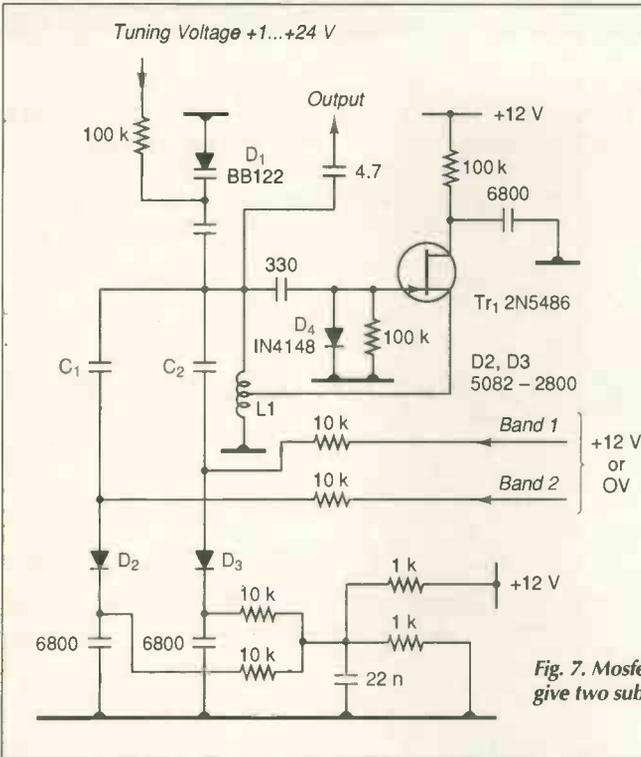


Fig. 7. Mosfet voltage-controlled oscillator with capacitor-switching diodes to give two sub-bands. Varactor diode D1 tunes within the bands.

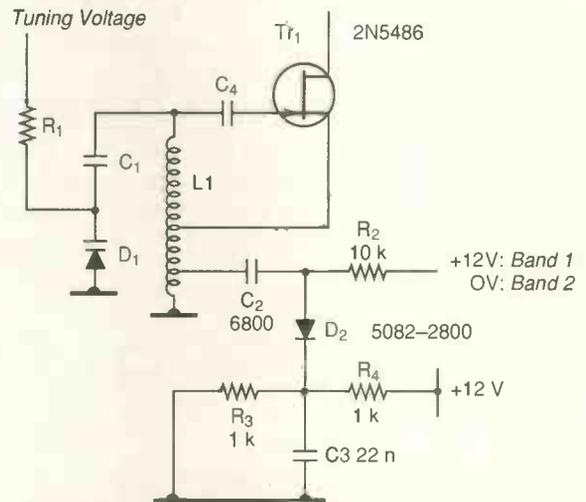


Fig. 8. Alternative scheme for sub-band switching, in which sections of tuning inductor are grounded via the switching diode.

counters being programmed by a three-wire bus or I²C bus, although there are some chips, for example the MC145146 by Motorola, with a parallel interface.

Figure 5 is a simplified block diagram of an I²C-controlled type by Philips, the TSA5510T, which contains all but the VCO. Its frequency band of 64-1300MHz is determined by the fixed prescaler with a divider ratio of 8, which is succeeded by a programmed divider with ratios of 256-32767; the upper limit of 1300MHz can be expanded considerably by using an additional prescaler, but 64MHz is

the guaranteed minimum operating frequency.

Access by the microcontroller to five I/O ports is a noteworthy feature of the chip.

Figure 6 shows a synthesiser in which the VCO has five sub-bands switched by the band-select signals. External transistor Tr₆ allows the tuning voltage range to increase to several tens of volts and Tr₇ works as an active noise filter to prevent power-supply disturbances getting to the VCO. With a 4MHz crystal, comparison frequency is 7.8125kHz so that, taking into account the prescaler by 8, the minimum frequency spacing is 62.5kHz.

This, as we have seen before, is acceptable for television receivers.

UHF synthesiser chips have their own problems, the main one being noise between internal IC circuits. To make an synthesiser with -110dB parasitic noise at 10kHz away from the carrier (as for a half decent signal generator) the PLL must be designed in discrete form, with extreme anti-noise precautions.

It is always necessary to use either several VCOs or one VCO with several sub-bands for broad-band synthesisers. The reason is simple - noise on the VCO control input causes parasitic frequency modulation at the output. Clearly, the higher VCO sensitivity in Hz/V, the higher the frequency deviation caused by the noise is and the worse the signal to noise ratio.

Figure 7 shows the mosfet VCO described in the first part of the article, modified to produce several sub-bands. Additional capacitors C₁ and C₂ tune the sub-bands, the varactor diode D₁, controlled by the phase-detector, tuning within each sub band. For reliable switching, D₂ and D₃ are given reverse offset by the resistive divider R₁R₂; the band-select inputs correspond to the control signals in Fig. 6.

It is sometimes convenient to arrange two sub-bands by grounding part of the tuning coil, as in Fig. 8. In UHF/SHF VCOs, a similar technique reduces the length of the microstrip line by switching diodes to get several sub-bands. This circuit, shown in Fig. 9, will work at frequencies up to 1GHz.

A common problem is the time taken to change frequency, particularly severe in true frequency modulators. The time constant of the loop filter must be such that the PLL does not respond to the lowest-frequency baseband products. In broadcasting, the lowest modula-

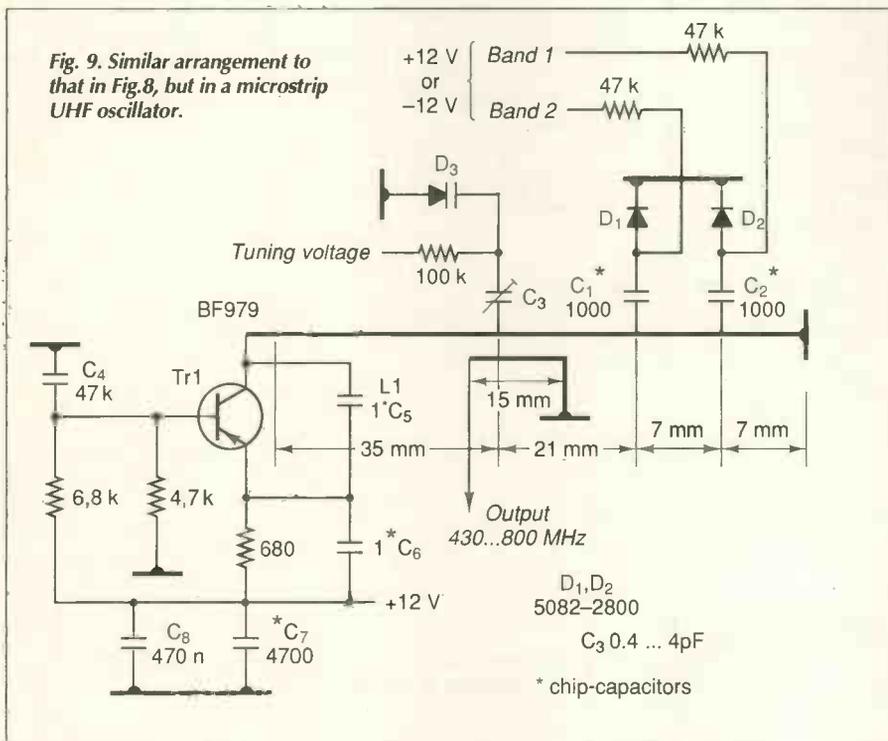


Fig. 9. Similar arrangement to that in Fig.8, but in a microstrip UHF oscillator.

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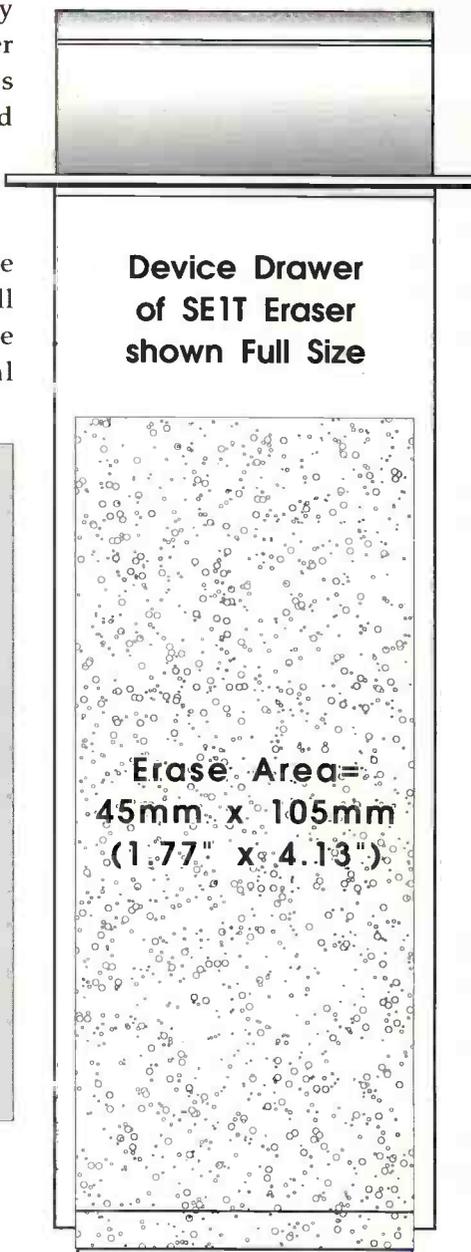


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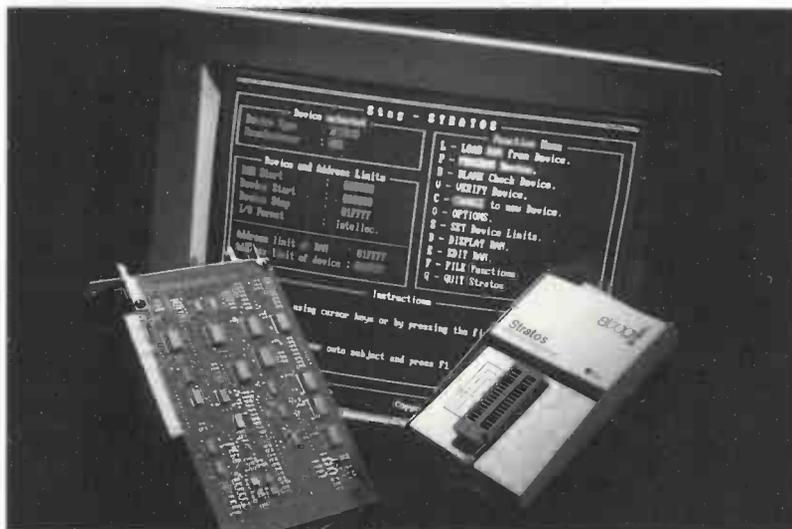
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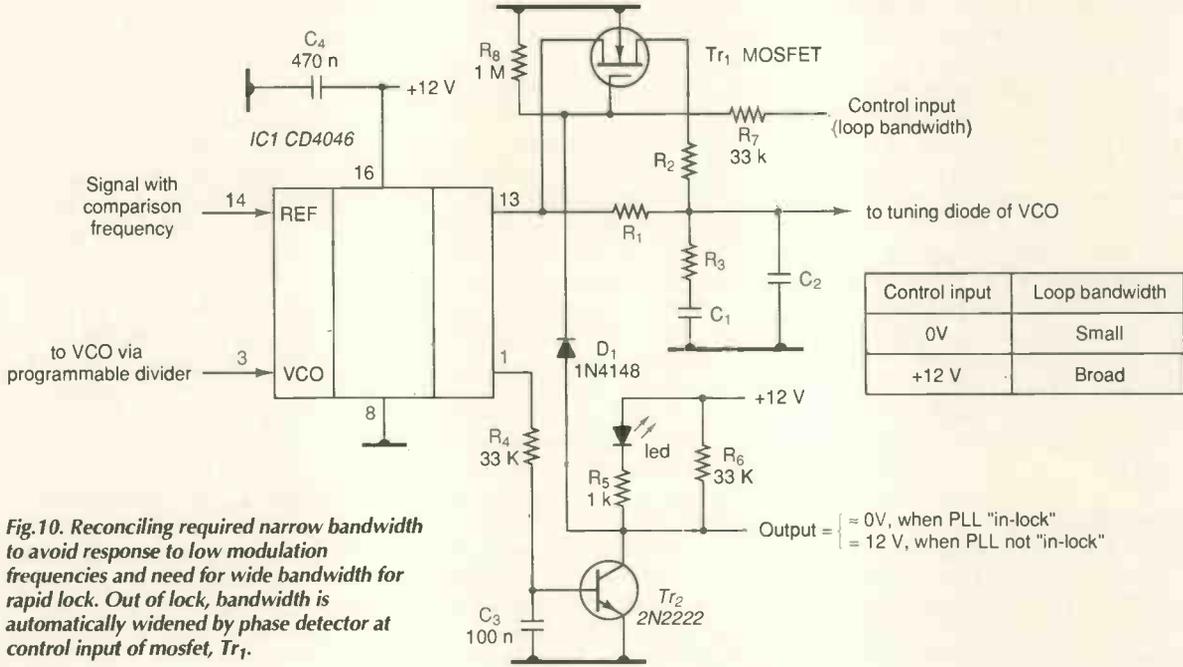


Fig.10. Reconciling required narrow bandwidth to avoid response to low modulation frequencies and need for wide bandwidth for rapid lock. Out of lock, bandwidth is automatically widened by phase detector at control input of mosfet, Tr_1 .

tion frequency is 30Hz, the time constant of the loop filter is 1-2s or more and switching time to within 0.1% takes 10-20s. This is often too long and special circuits must be used to reduce the duration of the transient processes.

An almost universal way of dealing with this is to widen the loop bandwidth when there is a mismatch in the loop and to narrow it when the PLL is in lock. A frequency-phase detector with the 4046 PLL chip makes this easier,

having an output indicating the state of lock in the synthesiser, the loop filter being formed by $R_{1,3}, C_{1,2}$ and Tr_1/R_2 controlling the loop bandwidth from pin 1 via Tr_2 .

In the in-lock condition, very short negative pulses appear at pin 1 and when not in lock, there are short positive pulses at this pin. The loop filter is the usual lag-lead filter with an additional capacitor C_2 for better pulse filter-

ing. Pulses from pin 1 are smoothed by R_4C_3 and the DC taken Tr_2 base so that, in the in-lock condition, there is a high level at the base of Tr_2 , Tr_1 is open-circuit and there is about 0.2V at the collector of Tr_2 . Lock is indicated by the led. Tr_1 is off and exerts no influence upon the loop filter time constant.

If loop is not in lock, Tr_2 is off, its collector sits at 12V, the lock indicator goes out and Tr_1 conducts placing R_1 in parallel with Tr_1 and

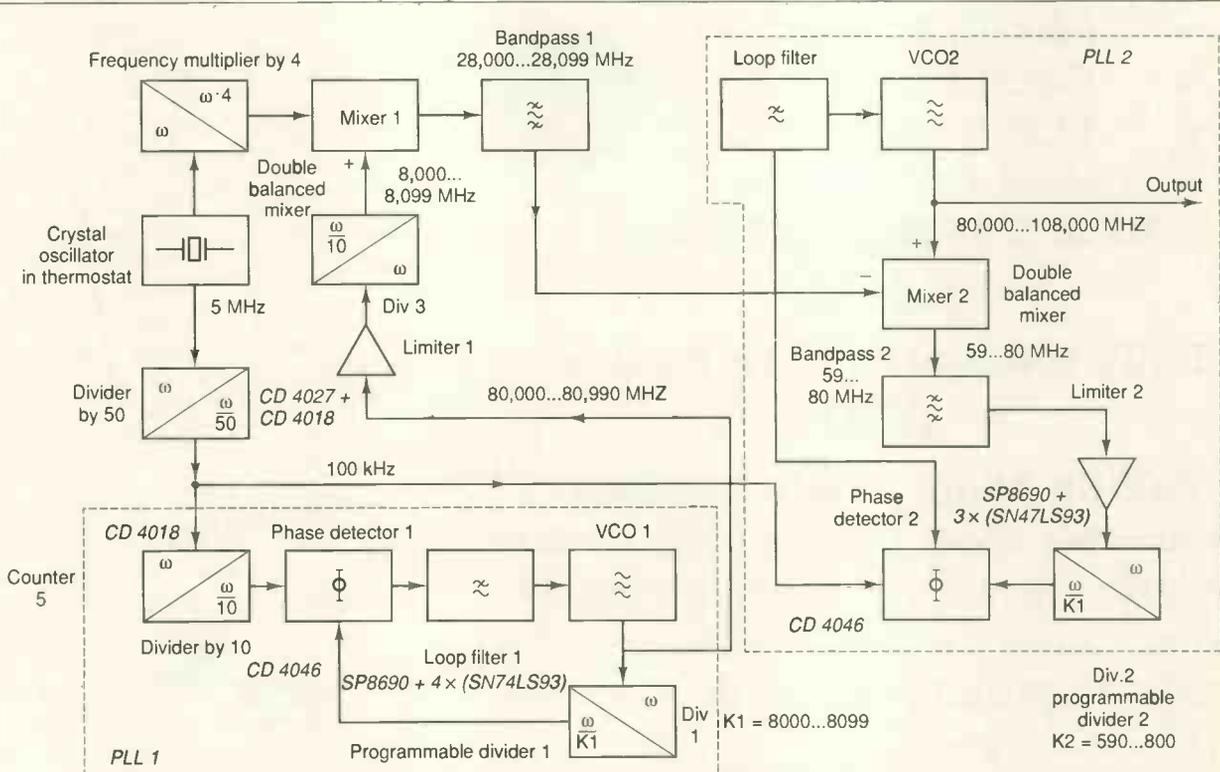


Fig. 11. Two-loop phase-locked loop producing 88-108MHz output with 1kHz spacing, designed to give the small spacing at high speed.

R_2 – about $700\Omega + R_2$. If this value is about $0.12R_1$, the loop bandwidth becomes 10 times wider and the transient processes correspondingly shorter. After having set the in-lock condition, T_{r1} is off and the initial loop bandwidth is restored. Due to the smoothing effect of R_4C_3 , everything proceeds smoothly.

A microcontroller can be used to switch loop filter bandwidth by way of the loop bandwidth input. Before switching frequency, the microcontroller widens the loop bandwidth by applying 12V at this input; it then switches the frequency, removes the 12V and passes control to the circuit described above. Diode D_1 decouples the control input. Signal at T_{r2} collector can also be used to eliminate odd noises that may arrive at the receiver output during tuning by either reducing volume or disabling the speaker output.

More than one loop

Discussion so far has centred on PLL synthesisers using one loop. In the case where high-speed frequency switching and small frequency spacing are required, types with two or more loops are usually used. Naturally, these devices are more complex but the end justifies the means and the use of UHF LSI ICs makes design considerable easier.

Figure 11 shows a two-loop PLL which

produces frequencies in the 87.5MHz to 108MHz range, with 1kHz spacing. It was designed as FM broadcast test equipment where there were no very high demands on spectrum purity. Signal to noise of the synthesiser output signal was to be no less than 60dB. On the other hand, however, the switching time was to be less than 10ms to speed up the test programme. Experience showed that a one-loop synthesiser with a 1kHz comparison frequency could be made with either the required spectral purity or with the switching time less than 10 ms, but not both, so a two-loop approach was adopted. In general outline this synthesiser consists of two almost identical PLLs and a few additional units.

A 5MHz, thermostatically controlled crystal oscillator with 0.05ppm stability was chosen as reference oscillator. Two counters divide its output by 50 and by 100 to give 10kHz for PLL_1 and 100kHz for PLL_2 . PLL_1 is an ordinary PLL with a phase-frequency detector; VCO_1 output frequency is in the range 80.00-80.99MHz, divided by 10 in counter DIV_3 to give 8.000-8.099MHz. The crystal oscillator signal also goes to a 4x frequency multiplier which consists of two double-balanced mixers with a mosfet buffer amplifier. Signal frequency from the output of DIV_3 and the 20MHz multiplier output are summed in

mixer 1, at the output of which a band-pass filter suppresses all conversion by-products of more than 60dB.

The IF of 28.000-28.099MHz goes to the input of *mixer 2*, which is part of PLL_2 . This is almost analogous to PLL_1 , but *mixer 2* is part of it and the signal at 2 output is the difference between the VCO_2 frequency and the IF. The second band-pass filter has a bandwidth of 59-80MHz, all by-products being suppressed by at least 60dB. DIV_2 has a division ratio 590-800.

Let the division ratio of DIV_1 be 8075, so that the output frequency of VCO_1 is 80.75MHz; the frequency at the output of divider 3 is 8.075MHz and the signal frequency at band-pass filter 1 output is 28.075MHz. If DIV_2 is set to divide by 655, then the signal frequency at VCO_2 output is $(655 \times 0.1) + 28.075 = 93.575$ MHz. It results that two PLLs with comparison frequencies 10kHz and 100kHz produce an output signal with 1kHz frequency spacing. The switching time in this synthesiser is determined, in general, by the parameters of the most passive unit in the system, the band-pass filter in PLL_1 . This is a lag-lead filter with a cutoff frequency equal of 4.5kHz. Total switching time of the synthesiser is 10ms or less.

Continued next month.

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CIRCLE NO. 120 ON REPLY CARD

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We are pleased to announce our twelve standard ranges of toroidal transformers approved to EN 60742.

EN 60742

The European Standard EN 60742 (based on the international standard IEC 742), covers the requirements for Isolating Transformers and Safety Isolating Transformers. It was ratified in 1988 by CENELEC, giving the standard the status of a national standard without any alteration for CENELEC members.

The standard is recognised as a national standard by each of the eighteen member countries of CENELEC and is known as BS 3535 in the UK and VDE 0551 in Germany.

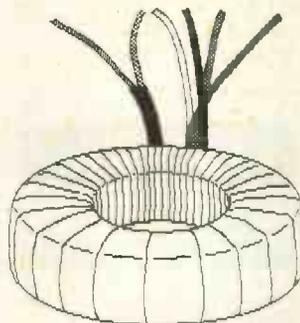
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CIRCLE NO. 122 ON REPLY CARD

Measuring L and C at frequency – and on a budget

Need a cheap-to-build instrument to measure capacitors and inductors at the operating frequency? Ian Hickman describes a constant level RF source coupled up to a versatile LC analyser.

Fig. 1a). Two transistor oscillator looks at first sight like an emitter follower driving a grounded base stage. But the earth point is an arbitrary convention. b) If the decoupling capacitors in a) are shown as short circuits at RF, the circuit is seen to be a balanced push-pull oscillator.

In the development labs of large companies, measurement of inductance or capacitance could not be easier. Simply connect the component to be measured to a network analyser and make an s_{11} measurement. The marker function can be used to obtain a screen read-out of the capacitance (or inductance) and the associated loss resistance – or the real and imaginary part of the impedance – at the frequency of interest. Change of apparent value with frequency can also be displayed as a Smith chart.

Unfortunately, the electronics engineer working outside the big lab will find the £15,000 cost of a network analyser to be a real obstacle.

Measuring capacitors presents no great problem since digital capacitance meters are cheap and readily available at less than £50 – many constructions have also appeared over the years. Most capacitors are near-ideal components and the frequency at which they are measured is largely immaterial. But engineers wanting to measure loss resistance at a given frequency are going to need a much more advanced measuring instrument. Similarly, inductors set much more severe measurement problems, since an inductor is really only usable over about two decades of frequency – at least for air-cored types. At higher frequencies the inductor resonates with its own self capacitance, while at about a hundredth of that frequency, Q drops to the point where it is of little use in a practical circuit.

FREQUENCY CHOICES

The traditional method of measuring inductance and capacitance is the Q meter. Models were available from manufacturers such as Hewlett Packard, Advance and Boonton. A well known early Marconi model came in a box almost a foot deep, with all controls, meters etc on the "front" panel, actually the top surface which was about two feet square. Highest operating frequency for this model was 25MHz and, perhaps for this reason, decade multiples of 250kHz are common frequencies for Q measurements.

The other common frequencies are decade multiples of 790kHz, perhaps because that is roughly root ten times 250kHz, giving two (geometrically) equally spaced spot test frequencies per decade and, it is also half the frequency corresponding to 10^7 radians per second.

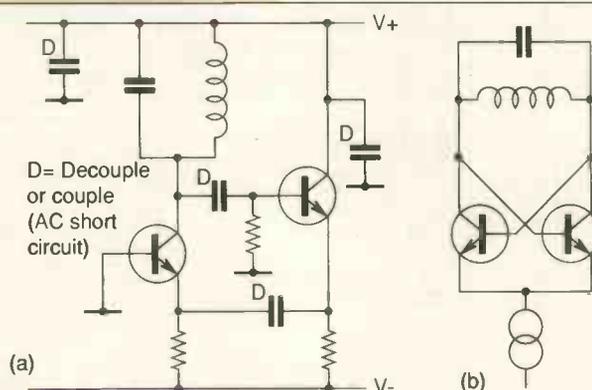
Since the Q of commercially available inductors, such as those used in this design, is commonly quoted at certain frequencies (see box), they were selected for the internal test frequency generator.

Circuit details

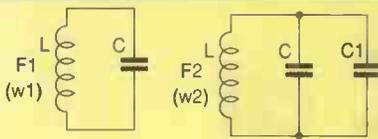
The two-transistor test generator basic circuit (Fig. 1a) looks at first sight like an emitter follower driving a grounded base stage, and it can indeed be analysed as such. But it is functionally equivalent to the push-pull oscillator of Fig. 1b. In any half cycle of the voltage appearing across the tank circuit, one transistor is cut off while current through both of the tail resistors flows through the other transistor. The tank circuit receives the total tail current, chopped up into a (near) square wave.

The transistors act largely as switches and amplitude of the tank voltage is given by its dynamic resistance R_d multiplied by the fundamental component of the current square-wave.

Full circuit of the test set (Fig. 2, excluding power supplies) shows that tank circuits giving seven fixed spot test frequencies are avail-



Quantifying internal capacitance



$$\omega_1 = \frac{1}{(LC)^{0.5}}, \quad \omega_2 = \frac{1}{(L(C+C_1))^{0.5}}$$

$$\frac{\omega_1^2}{\omega_2^2} = \frac{L(C+C_1)}{LC}$$

If C_1 is known then C is determined.

Let

$$\left(\frac{\omega_1}{\omega_2}\right)^2 = 1 + \Delta$$

then

$$1 + \Delta = \frac{C+C_1}{C} = 1 + \frac{C_1}{C}$$

So

$$\Delta = \frac{C_1}{C}, \quad C = \frac{C_1}{\Delta}$$

For example, if $C = (25\rho + C_{stray})$

and $C_1 = 8.33 (= 33.3\rho - 25\rho)$ then

$$\text{then } C_{stray} = (C - 25\rho) = \left(\frac{C_1}{\Delta} - 25\right)\rho$$

$$= \left(\frac{8.33}{\Delta} - 25\right)\rho$$

NB. Assumes F_1 is well below the self-resonant frequency of L , so that L is effectively the same at F_1 and F_2 .

Fig. 2. The switched frequency RF source (Tr_1 to Tr_3) provides a constant level drive source to the reactance under test, and detector/measurement circuit D_1 , IC_1 .

able, together with a facility for feeding in an external test signal of any desired frequency. The same LC ratio is employed for all the tank circuits, so that they all have the same R_d (about $4k\Omega$); or would do if the Q_s were all equal, as is roughly the case. The figure is reduced to about 700R by the shunting effect of R_2 and R_6 , giving a loop gain from Tr_2 base to Tr_1 collector of roughly 700 divided by R_4 , about times 14 or well in excess of unity, ensuring reliable oscillation. Given a total tail current through R_3 and R_5 of around 10mA, this provides a large enough swing across the tank circuit to chop the tail current into a respectable square wave, ensuring the amplitude of oscillation varies little from range to range.

On the other hand, in the EXT OSC IN position of S_1 , the collector load of Tr_1 is reduced to $47R$, giving a loop gain of less than unity, preventing oscillation. The RF tank voltage (or ext osc input) is buffered by Tr_3 , the output of which drives a test current (determined by the setting of R_9) into a cascode circuit composed of Tr_4 and Tr_5 .

The output admittance of a cascode stage is very low – especially when the first transistor is driven in grounded base – so that the test circuit is driven by a near ideal constant current generator. Of course, at the higher frequencies, the cascode's output impedance will fall, but so will the R_d of any practical circuit to be measured. So the design is adequate, and much easier to implement than the traditional Q meter scheme, where an RF current (measured by a thermocouple meter) is passed through a very low resistance placed in series with the LC circuit under test.

The voltage across the inductance under test, resonated with C_{15} , is detected by D_1 , which places very little loading on the circuit

owing to the high value of the following DC load, R_{16} . IC_1 acts as a buffer to drive the meter M_1 . T_{gain} of the buffer stage is adjustable over the range unity to $\times 12$ by means of R_{17} . The tuning capacitor C_{15} is a 500pF twin gang type, where one half has had all the moving plates but one removed. The operation is not nearly as tricky as it sounds, but is perhaps not for the ham-fisted. The modification reduces the maximum capacitance of C_{15a} to around 45pF, including the stray capacitances added by S_2 , S_3 , D_1 and Tr_5 .

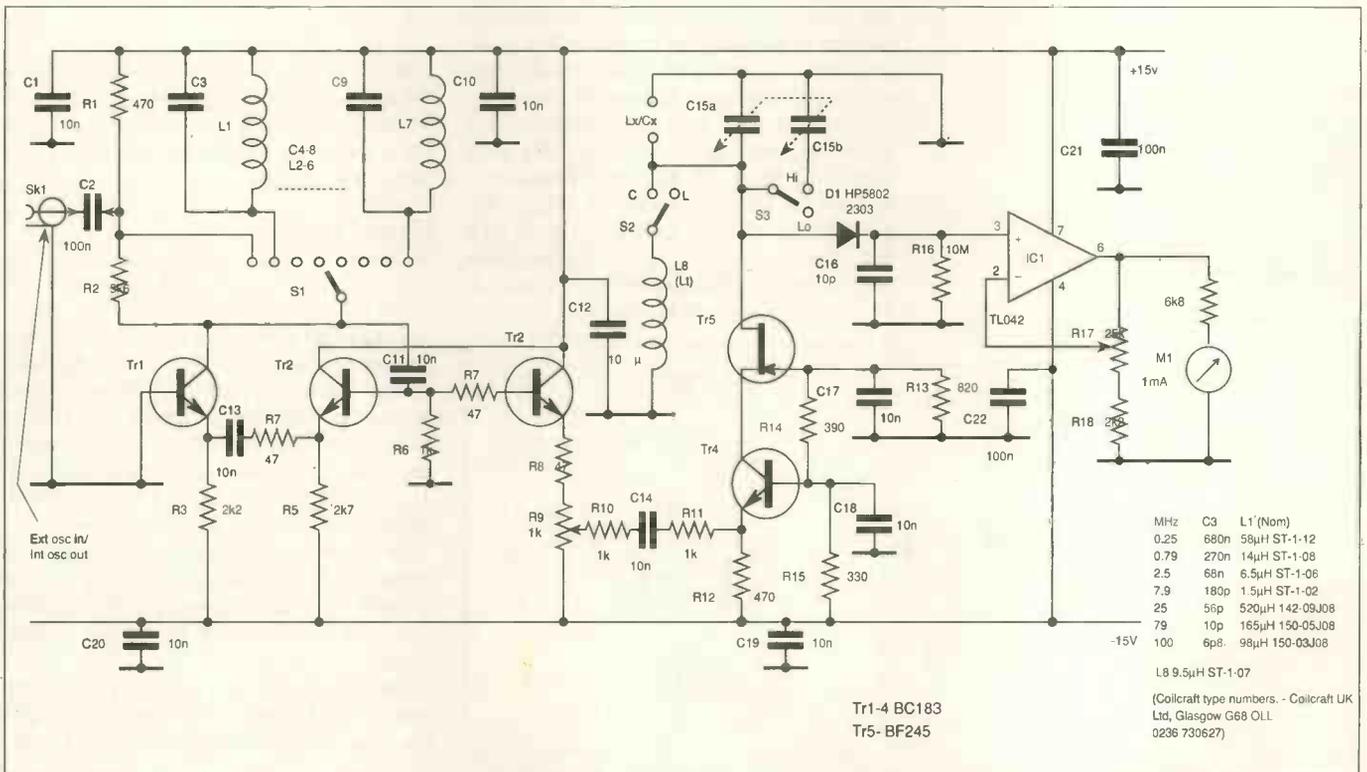
At lower frequencies, S_3 switches the 500pF section in parallel, enabling a wide range of inductors to resonate over the range 250kHz to 79MHz – or even 100MHz (using tank circuit C_9 L_7).

For measuring capacitors, the test inductor L_T (L_8) is switched into circuit, and resonated with C_{15} near maximum capacitance. The unknown capacitor is then connected to the test terminals – an Oxley pin projecting through the panel and an earth tag – and resonance restored by reducing C_{15} . The change in C_{15} capacitance gives the effective capacitance of the unknown capacitor at the test frequency used.

Constructional tips

This simple test instrument has proved very useful. But for best results some care is needed both in construction and in use.

The prototype was constructed in a diecast box, guaranteeing the absence of direct coupling between an inductor under test and whichever tank circuit is in use. Compact construction, especially around the test terminals, is essential to minimise stray inductance and capacitance which could cause problems at 100MHz.



To achieve this, while keeping the mechanics simple, the capacitance scales have been placed upon the side of the box, with all other controls on the top (Fig. 3a). Miniature, or sub-miniature, components are recommended, especially for S_2 and S_3 , as is use of a ground plane. In the prototype, the ground plane was simply a sheet of single sided copper clad SRBP, clamped to the underside of the front panel by the mounting bushes of S_1 and R_9 and connected by a wide piece of copper tape to the frame of C_{15} .

Fresh air construction is used for all parts of the circuitry operating at RF – 10nF decoupling capacitors being soldered to the ground plane wherever needed. Their opposite ends are used as mounting points for the other components – a form of construction which is as crude and ugly as it is cheap and effective (Fig. 3b).

As there was no intention to put the unit into production, there is no point in going through iterations to optimise PC layout. IC_1 was mounted on a scrap of strip board soldered to the ground plane, with the supplies brought in from the power unit mounted in the base of the box via a plug and socket.

Calibration

Calibration presents some interesting problems, solvable with the aid of four or five 100pF 1% capacitors readily ordered by post. Using various series/parallel combinations, capacitances of 20, 25, 33, 50, 67, 100, 125, etc up to 500pF can be made up. But the tricky part is how to take into account the stray capacitance associated with the test circuit.

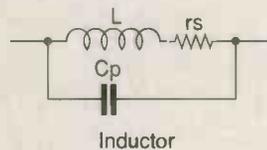
If the unit were only going to be used for measuring capacitors, the internal stray capacitance could be ignored, and the scale simply calibrated in terms of the capacitance added at the test terminals. But to measure inductors, knowing the frequency at which the circuit is resonated, requires also a knowledge of the true total circuit capacitance.

First step is to assemble the unit and fit the pointer knob of C_{15} . With the capacitor fully in mesh, make a reference mark on the blank scale so that the knob can always be refitted in exactly the same position if subsequently removed. Set S_2 to "C", S_3 to LO, the gang to minimum capacitance and connect a capacitance of 25pF to the test terminals. Feed in an external test signal, and note the frequency at which resonance is indicated. Now increase capacitance to 33pF and repeat the procedure.

From these results, the method shown (see box "Quantifying internal capacitance") will give a close approximation to the test circuit's true internal capacitance. Knowing this, the various combinations of the 100pF capacitors can be used to calibrate the HI and LO scales, making due allowance for the internal capacitance. The spot test frequencies of 250, 790kHz, 2.5, 7.9, 25, 79 and 100MHz should now be set up, by adjusting the cores of L_1 to L_9 , respectively. For this purpose, the frequency can be monitored at BNC coaxial socket Sk_1 .

L AND C EQUIVALENT CIRCUIT

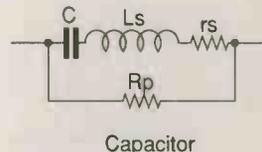
In addition to series loss component r_s and a series inductance L_s , a capacitor has a shunt loss component R_p . Except in the case of electrolytic capacitors, R_p is usually so high that it may be ignored. At the frequency F_r where C resonates with L_s , the capacitor looks resistive, and looks inductive above this frequency. For a tantalum electrolytic of a few microfarads, F_r is usually around 100kHz, with a very low Q. Dissipation due to the loss resistance r_s determines the maximum current that a capacitor, eg a mica type in an RF PA, can safely carry. Care should be taken when paralleling two decoupling capacitors, since for some types the Q at series resonance can be quite high. If having the same (nominal) value, one may resonate at a somewhat lower frequency than the other: at a slightly higher frequency its inductive reactance can resonate with the other capacitor – result, no decoupling at that frequency!



Inductor

Good practice is to make one capacitor at least ten times as large as the other.

Where a capacitor has a parasitic series inductance L_s , an inductor has a parasitic shunt capacitance C_p . This cannot accurately be considered lumped, being distributed between the various turns of the winding. Inductors are much more imperfect components than capacitors. Whereas the latter can be used over a frequency range of $10^7:1$ or more, the range between the self resonant frequency of an inductor and the frequency at which its Q has fallen to an embarrassingly low value is as little as 100:1 – at least for air-cored types, including those with a slug adjuster. High A_I inductor pot cores can provide a large inductance with very few turns, reducing C_p , especially if the turns are spaced, resulting in a wider useful operating frequency range.



Capacitor

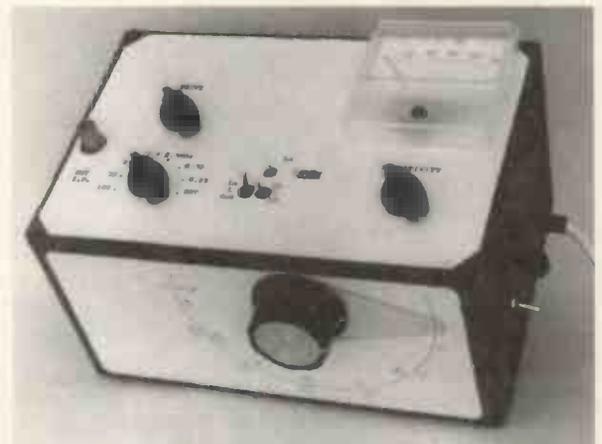
Operation

In use, R_{17} should normally be set anticlockwise at the minimum gain setting, with just enough drive applied to the test circuit by R_9 to give full scale deflection. Under these conditions, the RF signal into the detector is large enough to give a linear response.

So, by detuning either side of resonance to 71% of meter FSD and noting the two capacitance values, the Q of the inductor under test can be estimated. (If the average of the two values, divided by their difference, is 25, then the Q is 50, courtesy of an approximation based upon the binomial theorem.)

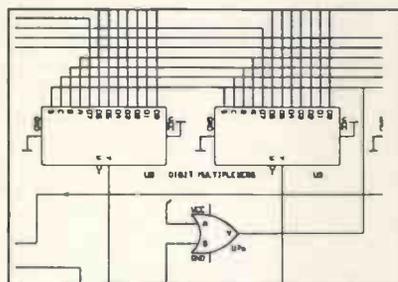
At higher frequencies, where the lower value of R_d of the test circuit is such that full scale deflection cannot be achieved even with R_9 at maximum, R_{17} should be advanced as necessary. As mentioned, capacitors are measured by switching the test inductor L_T into circuit and noting the reduction in value of C_{15} required to restore resonance when the unknown capacitor is connected to the test terminals. As the Q of the capacitor under test is likely to be greater than that of L_T , estimation of the capacitor's Q is usually not possible – a limitation the instrument shares with traditional Q meters.

Fig. 3a). Completed test set, showing controls on two faces of the box b) Internal view, showing the "fresh air" construction on a ground plane.



Electronic Designs Right First Time?

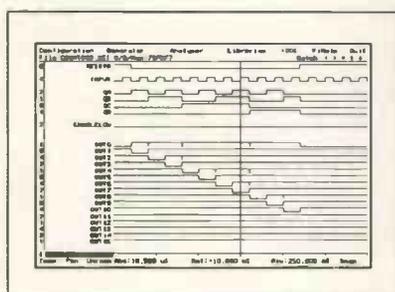
From Schematic Capture -



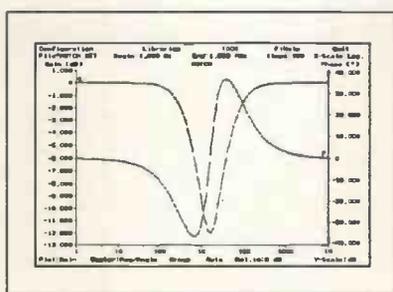
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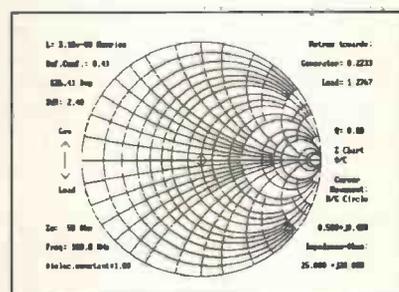
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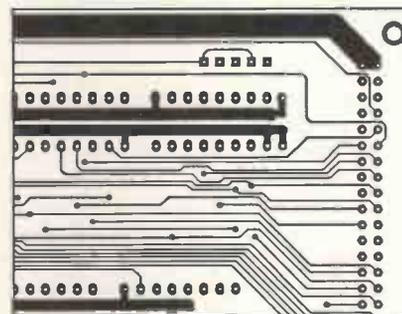
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Goertzel alternative to the Fourier transform

Sometimes the results of the FFT may not be in the best form for display on a particular screen. Allen Brown shows how the Goertzel algorithm is a useful alternative.

One problem with using the FFT to find the frequency content of a signal relates to the number of input data points and the number of spectral points. Normally if there are N data points then the FFT will give $N/2$ unique spectral points which may not always be desirable.

For example an engineer may require a spectral map to fill the display area of a screen which has 600 pixels (VGA for instance) but only has 100 data points to begin with. One way to provide a spectrum when the number of data points is different to the required number of spectral points is to use the Goertzel algorithm, developed as follows. The FT (or more precisely the discrete FT) is given by:

$$X(k) = \sum_{n=0}^{n=N-1} x(n)W_N^{nk},$$

where $x(n)$ is the input data (which can be complex) and

$$W_N^{nk} = \cos\left(\frac{2\pi kn}{N}\right) - j \sin\left(\frac{2\pi kn}{N}\right)$$

Bearing in mind that $W_N^{Nk}=1$ then we may express $X(k)$ as,

$$X(k) = \sum_{n=0}^{n=N-1} x(n)W_N^{-k(N-n)},$$

which can be thought of as a filter mechanism where the finite length filter $W_N^{-k(N-n)}$ convolves with the input data $x(n)$. New variable $y_k(n)$ can now be defined:

$$y_k(n) = \sum_{m=0}^{m=N-1} x(m)W_N^{-k(N-m)}.$$

Impulse response of this filter is given by:

$$h_f(n) = W_N^{-kn}u(n),$$

where $u(n)$ is a step input ($u(n) = 1$ for $n \geq 0$, and 0 otherwise). The Z-transform of $h_f(n)$ will lead to:

$$H_k(z) = \sum_{n=0}^{n=N-1} W_N^{-kn}u(n)z^{-n}.$$

Expanding the right hand side of the equation and recognising that it is a geometric progression, it becomes,

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Frank Ogden

Listing 1

Coding of the Goertzel Algorithm. The input data is fetched from two data files called test_r.dat and test_i which are in the root directory of drive C. The real and image spectral data is stored in files real.dat and imag.dat which are also in the root directory of C. Although the routine given here is for complex input data, it can be modified to cater for real data points only. */

```
#include <stdio.h>
#include <math.h>
#define N 100 /* Number of input data points */
#define K 400 /* Number of spectral points */

float GOERTZEL( float x[][2]);

static float x[N][2];
static n;
static char x_data[80];

main()
{
    FILE *fp; /* Open pointer to the read only file */

    /* Input data points from a disc file which
    is in C drive. The ASCII string data must be converted
    into floating point numbers. */

    if( (fp = fopen( "c:\test_r.dat", "r" )) != NULL )
    {
        for( n=0; n <= N-1; n++ )
        {
            fgets( x_data, 80, fp );
            x[n][0] = atof( x_data ); /* Real data points */
        }
    }
    else
    {
        printf( "Error reading test_r.dat \n " );
    }
    fclose( fp );

    if( (fp = fopen( "c:\test_i.dat", "r" )) != NULL )
    {
        for( n=0; n <= N-1; n++ )
        {
            fgets( x_data, 80, fp );
            x[n][1] = atof( x_data ); /* Imaginary data points */
        }
    }
    else
    {
        printf( "Error reading test_i.dat \n " );
    }
}
```

```

    }
    fclose ( fp );

    GOERTZEL( x );
}

float GOERTZEL( float x[ ][2] )
{
    float vr1, vr2, vi1, vi2, temp, yfr, yfi;
    float c, s, phi, wn;
    static float real[K], imag[K];
    int j, k ;

    FILE *rfp; /* Pointer for real spectral file */
    FILE *ifp; /* Pointer for imag spectral file */

    /* Begin the Goertzel Algorithm */

    wn = 3.14159265 / K ;

    for( k=0 ; k <= K-1 ; k++)
    {
        vr1 = vr2 = 0;
        vi1 = vi2 = 0;
        phi = wn * k;

        c = cos(phi);
        s = sin(phi);

        for( j=0; j <= N-1; j++)
        {
            temp = vr1;
            vr1 = 2 * c * vr1 - vr2 + x[j][0];
            vr2 = temp;
            temp = vi1;
            vi1 = 2 * c * vi1 - vi2 + x[j][1];
            vi2 = temp;
        }
        yfr = c * vr1 - vr2 - s * vi1;
        yfi = c * vi1 - vi2 + s * vr1;

        real[k] = yfr ;
        imag[k] = yfi ;
    }

    /* Open disc files real.dat and imag.dat to deposit the
    spectral data from the Goertzel Algorithm */

    rfp = fopen( "c:\\real.dat", "w" );
    ifp = fopen( "c:\\imag.dat", "w" );

    if( rfp != NULL || ifp != NULL )
    {
        for( k=0 ; k <= K-1; k++)
        {
            fprintf( rfp, "%f \n", real[k]);
            fprintf( ifp, "%f \n", imag[k]);
        }
    }
    else
    {
        printf( "Error with spectral file \n " );
    }

    fclose( rfp );
    fclose( ifp );
}

```

$$H_k(z) = \frac{1}{1 - W_N^k z^{-1}}$$

But to implement this filter would involve complex multiplications and these can be avoided by expressing this equation as:

$$H_k(z) = \frac{1}{1 - W_N^k z^{-1}} \cdot \frac{(1 - W_N^k z^{-1})}{(1 - W_N^k z^{-1})}$$

Leading to,

$$H_k(z) = \frac{1 - W_N^k z^{-1}}{1 - 2 \cos\left(\frac{2\pi k}{N}\right) z^{-1} + z^{-2}}$$

Now let

$$H_k(z) = \frac{Y_k(z)V_k(z)}{X_k(z)V_k(z)}$$

So

$$\frac{Y_k(z)}{V_k(z)} = 1 - W_N^k z^{-1}$$

and

$$\frac{V_k(z)}{X_k(z)} = \frac{1}{1 - 2 \cos\left(\frac{2\pi k}{N}\right) z^{-1} + z^{-2}}$$

By using the inverse Z-transform these last two equations will lead to two difference equations,

$$v_k(n) = 2 \cos\left(\frac{2\pi k}{N}\right) v_k(n-1) - v_k(n-2) + x(n)$$

and

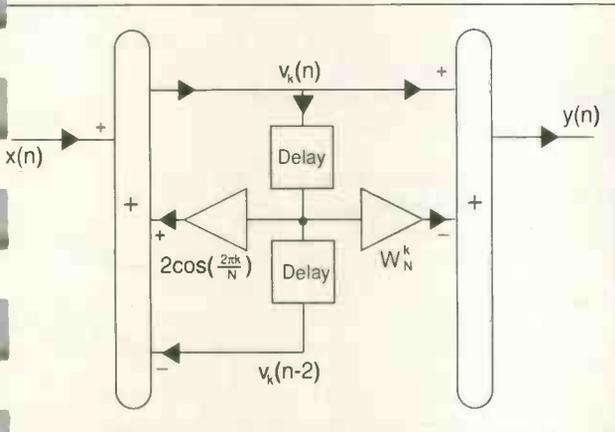
$$y_k(n) = v_k(n) - W_N^k v_k(n-1)$$

$y_k(n)$ represents the spectral data points generated from the input data points $x(n)$. These equations (shown diagrammatically in the figure) form the kernel of the Goertzel algorithm and can be implemented directly. The listing shows the coding of the two difference equations in C. As an example The second figure shows two traces of input data (upper traces) which are combined to form the complex data set $x[][2]$. These are stored as separate data files, test_r.dat and test_i.dat. Also shown in the figure, (lower traces) are the output spectra from the Goertzel algorithm (400 spectral points each).

Although the code was compiled using Microsoft C compiler (Version 7), it can be compiled for any processor (including a DSP device) which has provisions for handling floating point numbers – provided suitable changes are made to the i/o features.

The Goertzel Algorithm is not as fast as the FFT, but it does have a speed advantage over the discrete FT (see first equation).

Note also that the above routine does not include a windowing function which is sometimes required to minimise the "fringe" effects of finite data sample sets.



Schematic representation of the Goertzel algorithm.

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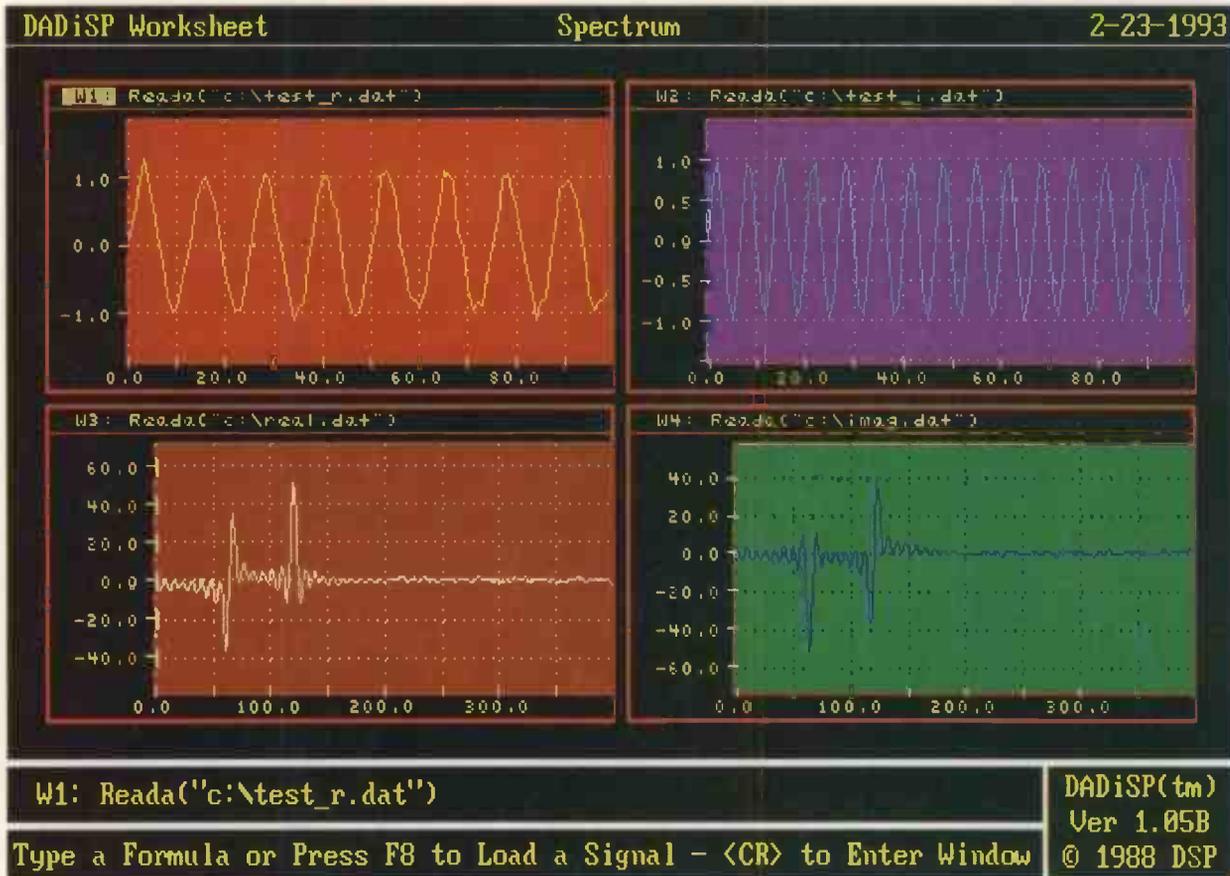
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Please send your ideas to Frank Ogden, EW + WW, Quadrant House, The Quadrant, Sutton, Surrey, SM2 5AS.

FFT

When presented with a signal, the engineer often requires knowledge of its frequency content. If the signal is subjected to the Fourier Transform (FT), the resulting complex data represents the frequency spectral from which the phase and magnitude information can be derived. The well known algorithm for deriving the FT is the Fast Fourier Transform (FFT), but it is not the only method of calculating the FT.

Input data and output spectra from the Goertzel algorithm



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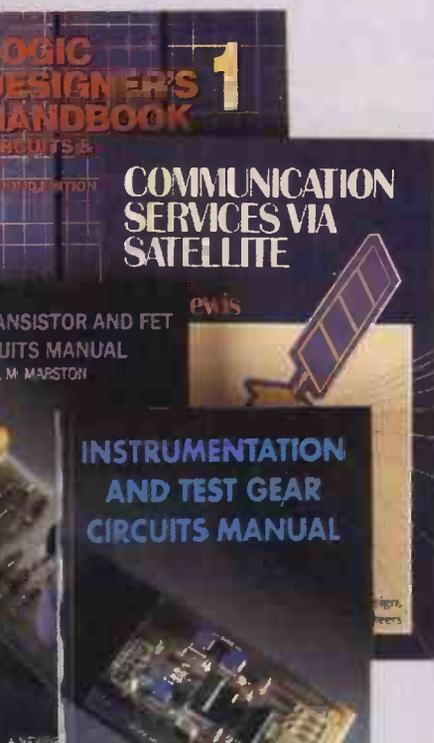
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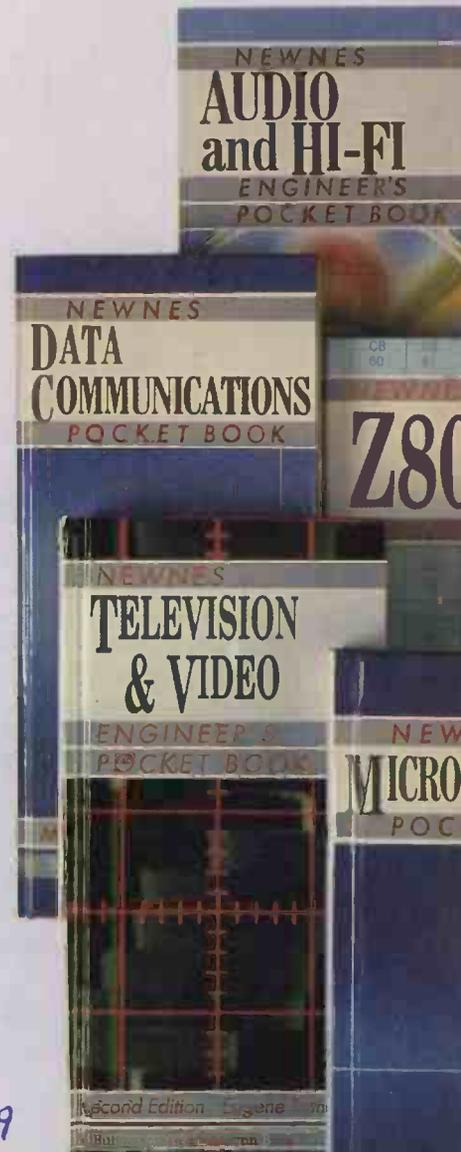
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An easier route to light measurement

Typical light sensing elements convert light to a signal current, voltage, or resistance. Most applications require amplification before use. Digital output will additionally require some form of ADC to process the linear signal. The light-to-frequency converter provides all this and many other useful benefits.

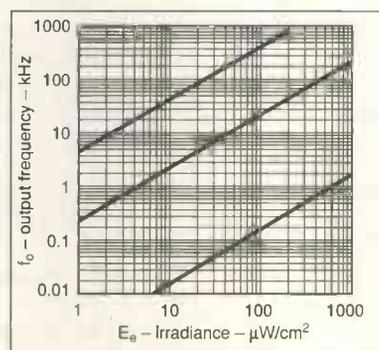
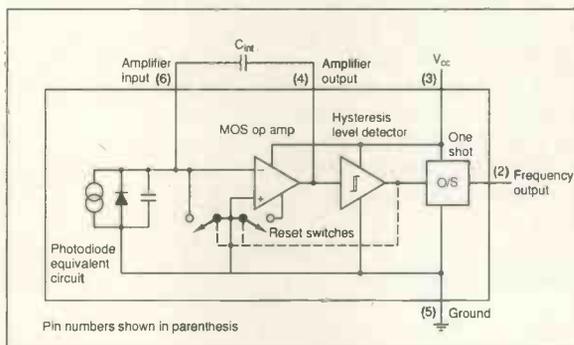
Since the signal is in the form of a frequency, dynamic range is not limited by supply voltage and noise – over 100dB may be easily obtained. The TSL220 can be set up to measure light intensity from dusk to full sunlight in a single range.

Once converted to a frequency, the signal is virtually noise immune and may be transmitted over cables from

The TSL220 light-to-frequency converter is designed to convert light intensity levels into a digital format. It provides a simple, low-cost method of light-to-digital conversion. The output may be coupled to isolated circuitry or other equipment through the use of a simple optocoupler.

By Frank Ogden.

TSL 220 internal diagram and pinouts. The supply voltage should be kept within the range 4 to 10V. It draws about 7mA.



Output frequency vs irradiance for a number of capacitor values.

remote sensors to other parts of the system. Being a serial data form, the signals of several sensors may be multiplexed to a single processor port or counter using digital logic. Integration of the frequency signal will eliminate low frequency (such as 50 or 100Hz) interference or allow measurement of long-term light exposure rates.

Circuit description

The TSL220 current-to-frequency converter circuit consists of an op-amp integrator, transistor reset switches, a level detector, and a one-shot pulse generator. The output voltage is a CMOS compatible pulse train of a frequency directly proportional to the light intensity falling on an integral photodiode. The output frequency may be

measured using pulse counting, period timing, or integration techniques; the output frequency for a given light intensity at the input depends on an external capacitor value.

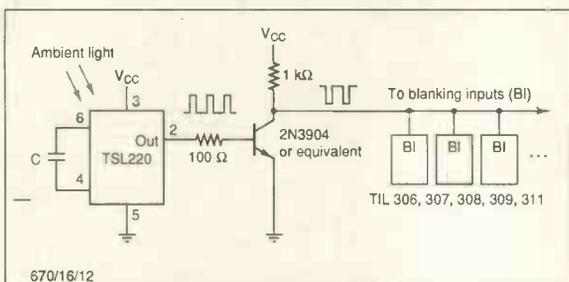
The integrator produces a ramp with a slope proportional to the photodiode current. The level detector resets the integrator when the ramp reaches an upper threshold voltage. The cycle is then repeated, thus producing the frequency signal. The output of the level detector triggers the monostable to produce the fixed-width output pulse.

During the integration phase of operation, the reset switches are open, allowing the photodiode current to charge the integrating capacitor C towards the upper threshold voltage of the level detector. The charging time is proportional to photodiode current and capacitor value. When the upper threshold voltage is reached, the level detector changes state, triggering the one-shot. The level detector also turns on the reset switches, shorting the photodiode node to ground. The capacitor is then rapidly discharged through an internal load resistor toward the lower threshold voltage. Since this discharge time is very short compared to the integration time, the frequency is determined by the integration period. When the lower threshold voltage is reached, the reset switches open, and the photocurrent again begins to charge the capacitor toward the upper threshold voltage.

The device is packaged in a clear plastic 8-pin dual-in-

READER SERVICES OFFER:

To obtain your free sample of the Texas Instruments TSL220 light-to-frequency converter, fill in and send off the special reply card located between pages 496 and 497 of this issue. This reader services offer is being handled directly by Texas Instruments. Our editorial office is unable to assist in any queries relating to it.



Automatic dimming circuit. The timing capacitor is set to maximum and minimum brightness levels

line package and has an active chip area of 4.13mm².

Design considerations

The capacitor value should be determined by the light level available, the required measurement time, and the measurement technique employed. The capacitor value also affects output pulse width and thus the duty cycle of the output signal. Precision measurement requires use of a low leakage component. Its leads should be kept as short as possible and the capacitor placed close to the device to minimise noise pickup and stray capacitance. Additionally, a grounded guard ring should be used around pins 6, 7, and 8 and the capacitor leads to isolate this input from potential leakage currents. Its value will depend on the maximum amount of illumination available to the sensor, and the desired output frequency range.

If a system is to function outdoors, it must be designed to operate under daylight conditions. Although the TSL220 will work at much higher light levels than those quoted in the spec sheet (1000µW/cm²), there is a limit beyond which the device no longer operates. This corresponds to approximately 0.4mA of photocurrent which may easily be exceeded in full sunlight.

If the device is to operate under high intensity light conditions, a neutral density or other type of filter must be used to ensure proper operation over the full light range. From Table 1, the maximum light condition (full sunlight) would generate a photocurrent of 0.225mA, which does not exceed the internal photocurrent limitation.

A typical application

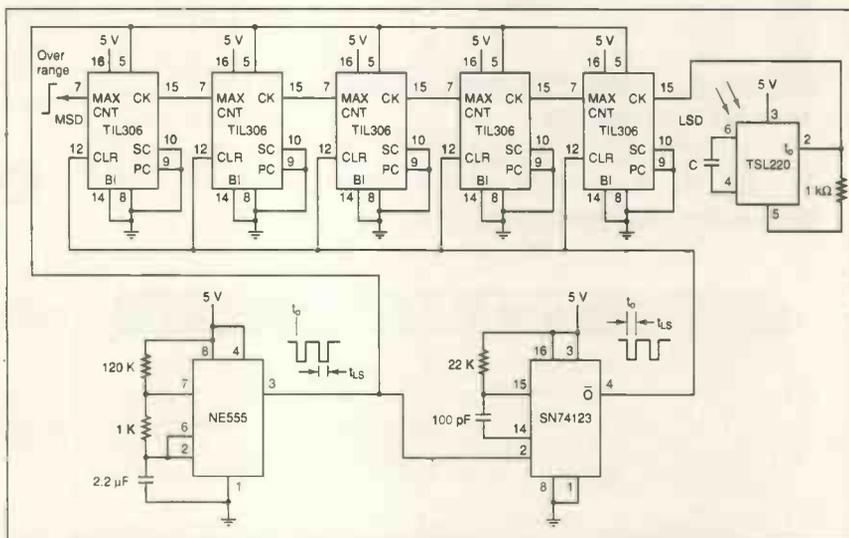
The operating range and simple interface of the TSL220 make it ideal for use in microprocessor-based systems to adjust display brightness to compensate for ambient lighting conditions. The system described here, built around a Motorola 6805 series microcontroller, could be used to provide display adjustment in systems such as vehicles or portable equipment under indoor and outdoor conditions. The device connects directly to an MCU input port. No other interface components are required for this; if it had been required to hook up with LS TTL, a 3.3kΩ pull-down resistor to ground would have been needed. To determine the optimum capacitor value, the limitations imposed by the microprocessor speed must be considered. The maximum clock speed of the processor (4MHz) and the necessary instruction cycle time to perform the counting function will limit the maximum input frequency from the TSL220 to about 41kHz, with a minimum pulse width of 6µs. A capacitor value of 0.05µF will satisfy both conditions with plenty of safety margin.

The microcontroller, which is assumed to be resident in the system and already performing other functions, is used to scale the input received from the TSL220 and adjust the display brightness based on empirically determined thresholds or lookup tables. Various methods could be used to adjust the display brightness. When decoded alphanumeric displays are used (such as the H-P

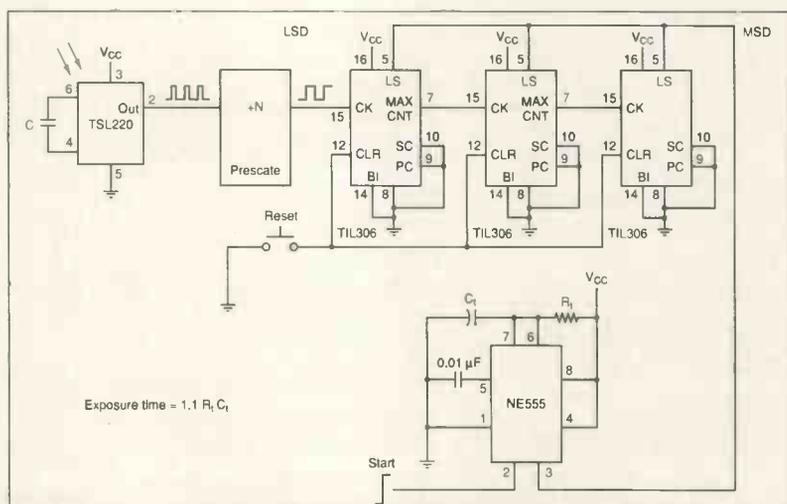
TSL220 DIODE PHOTOCURRENT WITH VARIOUS FILTER TYPES

Condition	No filter µA	IR filter µA	Light Lux	F _{out} C=10nF kHz
Sunlight (noon)	1100	225	140,000	130
Sunlight (afternoon)	980	200	125,000	120
Clear sky (zenith)	53	10	6250	7
Clear sky (horizon)	21	5	3125	3.5
Evening twilight	15	3	1875	2
Shade (sunny day)	6	0.7	435	0.485
Office lighting	2	0.8	550	0.61
Dusk	0.85	0.12	70	0.005
Nightfall	0.018	0.0025	1.5	0.002

HDSP21xx series), the brightness may be controlled in discrete steps by writing the appropriate control word to the display. Other displays such as the TIL311 may be dimmed using the blanking input to periodically turn off the display at a rate high enough to be unnoticeable to the eye. The blanking signal may be generated using periodic interrupts or a timer on the MCU. Similarly, the variable duty-cycle technique may be used to control the intensity of incandescent bulbs or discrete leds for display backlighting or dial illumination.



Digital light meter. The NE555 controls the timing period over which the internal display counters integrate the pulses from the TSL220. It also enables the display latches to provide a stable reading. The monostable provides a short reset pulse to the display counters clearing them down for the next count.



A slight variation on the light meter is the digital exposure meter. The integral display counters show a single count total for a preset period determined by the 555 time constant, the actual value of which is 1.1R₁C₁.

APPLICATIONS

Drivers for IGBTs include short circuit protection

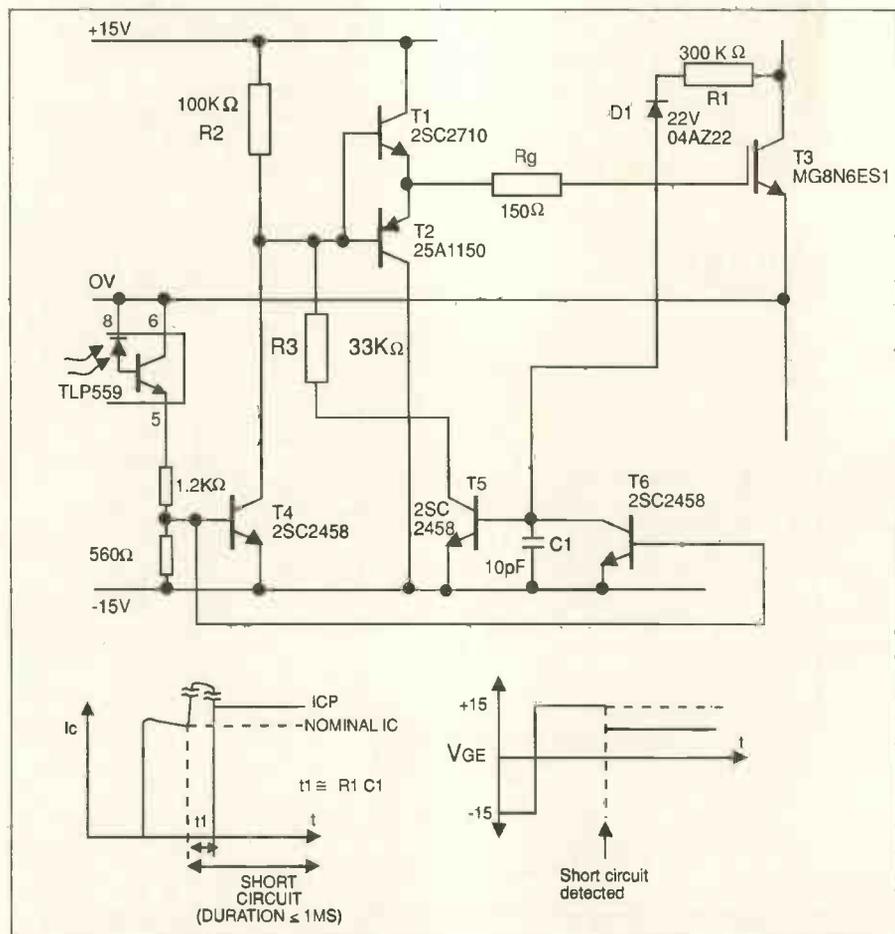
Requirements for insulated gate bipolar transistor drive circuits are similar to those for a power mosfet driver. A series gate resistor damps oscillations due to stray inductance and input capacitance, as well as limiting charge and discharge currents. But, as Toshiba's note *IGBT Drive Conditions* points out, the IGBT gate resistor not only determines switching performance, but also controls the device's dissipation under short-circuit conditions.

All five drivers mentioned within the note provide the necessary $\pm 15V$ swing and all but one, which receives its signal via a pulse transformer, are optically isolated. In the circuit shown, Tr_5 offers protection by limiting current to the IGBT's peak collector rating under fault conditions. When the collector of Tr_3 is pulled upwards

excessively by its load during saturation, Tr_5 turns on to reduce the IGBT gate voltage to a level determined potential divider $R_{2,3}$. The zener diode determines the IGBT collector voltage level at which Tr_5 turns on and Tr_6 is needed to discharge the capacitor.

One of the remaining circuits is identical to the one shown with the exception that it has no R_3 . As a result, the protection mechanism turns the IGBT off rather than limiting it. There is also a driver that provides an error signal, via logic gates, indicating a short circuit. Transient overloads, which the IGBT can handle, are filtered out by the logic.

Toshiba Electronics (UK) Ltd, Riverside Way, Camberley, Surrey GU15 3YA, Phone 0276 694600.



Non-linear filter improves wideband noise performance

Noise is commonly filtered to improve signal-to-noise ratio in data conversion systems. Such filtering has the drawback that it decreases bandwidth which in turn increases settling time. One solution is claimed to be a non-linear low-pass filter which provides a 4-to-1 improvement over a traditional single-pole RC filter, reducing settling time from 147 to 37 μ s.

With simple non-linear filters based on diodes strategically based to provide charge/discharge paths for the capacitor elements, benefits of the circuit reduce as the input step voltage reduces. This improved circuit reduces that drawback by reducing the threshold below the diode voltage drop. A full discussion is presented in Burr Brown's application bulletin AB022.

To understand how a non-linear filter can improve settling time, consider the simple diode clamped nonlinear filter shown in Fig. 1.

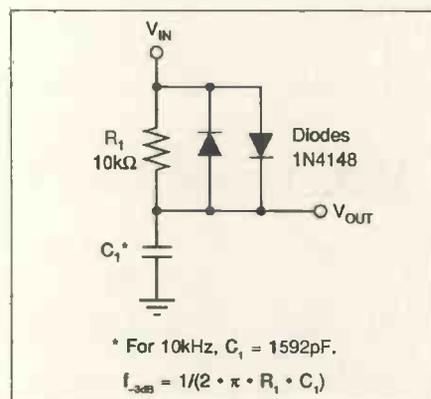


Fig. 1. Settling time is improved because the filter capacitor C_1 is charged faster through the low forward biased diode impedance during the initial portion of a large input step change.

Settling time is improved because the filter capacitor C_1 is charged faster through the low forward biased diode impedance during the initial portion of a large input step change. When the difference between the input and output voltage becomes less than the forward biased diode drop, the diode turns off and C_1 reacts with R_1 alone. At this point, the circuit behaves like a normal single pole RC filter.

Assuming diode resistance is negligible,

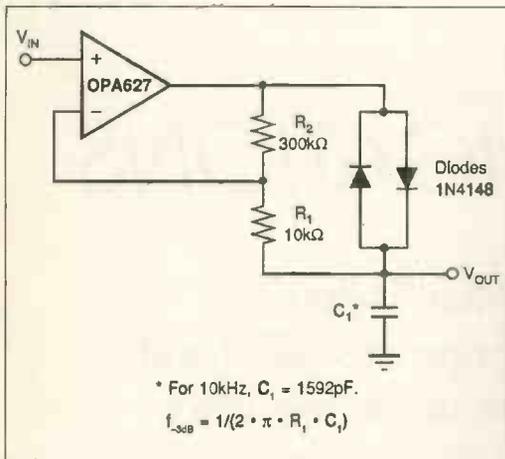


Fig. 2. Improved nonlinear filter can improve 0.01% settling time for a conventional filter by 4/1 for a 20V step.

the improvement in settling time depends on the ratio of the input step voltage to the forward biased diode voltage.

Improvement in settling time only occurs when the voltage step is much larger than the diode drop. Small steps show less improvement over normal circuit time constant dictates. As the step change approaches the forward biased diode voltage, the simple nonlinear filter offers no improvement.

By reducing the threshold to below one diode drop, the settling time can be improved for smaller inputs. The improved nonlinear filter shown in Fig. 2 allows adjustment of the threshold to a small arbitrary value by adjusting the ratio of R_1 and R_2 .

To see how the improved non-linear filter works, notice that the op-amp forces the voltage at its inverting input to be the same

as at the non-inverting input. For small differences between the output voltage and the input voltage, the difference is dropped across the 10kΩ resistor, R_1 and the filter behaves like a single-pole filter with an RC time constant. The voltage divider formed by R_1 and R_2 amplifies the voltage difference across R_1 , as seen at the top of R_2 , by $(1 + R_2/R_1)$. As the voltage across the R_1 , R_2 divider becomes larger, one of the diodes (which one depends on signal polarity) begins to conduct, and the capacitor is rapidly charged through it. This occurs at a voltage difference between the input and output of about $0.6V/(1 + R_2/R_1)$ or about 20mV with the values shown. With the diode forward biased, the time constant of the filter becomes very small, limited only by op amp slew rate or current limit.

To determine the component values for the improved nonlinear filter, consider the noise-reduction requirements of the filter. For example, if you want to filter the noise of a 20V full-scale signal to 0.01% resolution, the peak noise must be filtered to less than 0.01% of 20V, i.e. 2mV peak.

A clamp threshold of ten times this peak (20mV) is an arbitrary but ample threshold. The component values shown in Fig. 2 achieve this. For a 20V step as before, the improvement in settling time is $\ln(0.02/20) = 6.9$ time constants. In other words, for a 20V step, the improved nonlinear filter can improve 0.01% settling time from 9.2 time constants to $(9.2) - (6.9) = 2.3$ time constants – a four-to-one improvement.

In some instances, the input signal may have noise peaks above the 20mV threshold of the nonlinear filter. If the noise of the input signal is greater than the 20mV threshold, the filter will

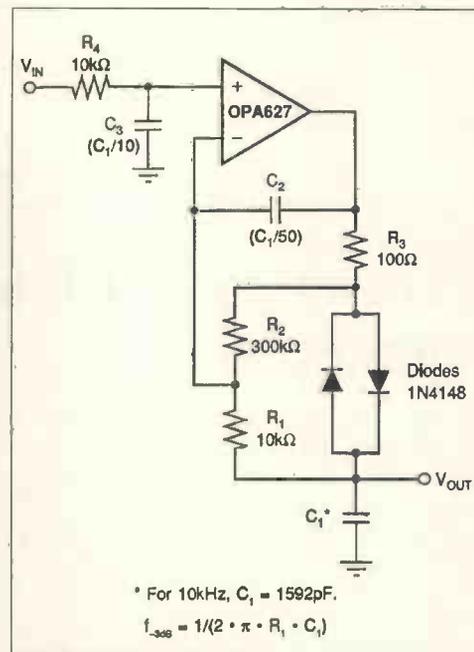


Fig. 3. Improved nonlinear filter with extra networks to ensure stability when driving capacitive loads.

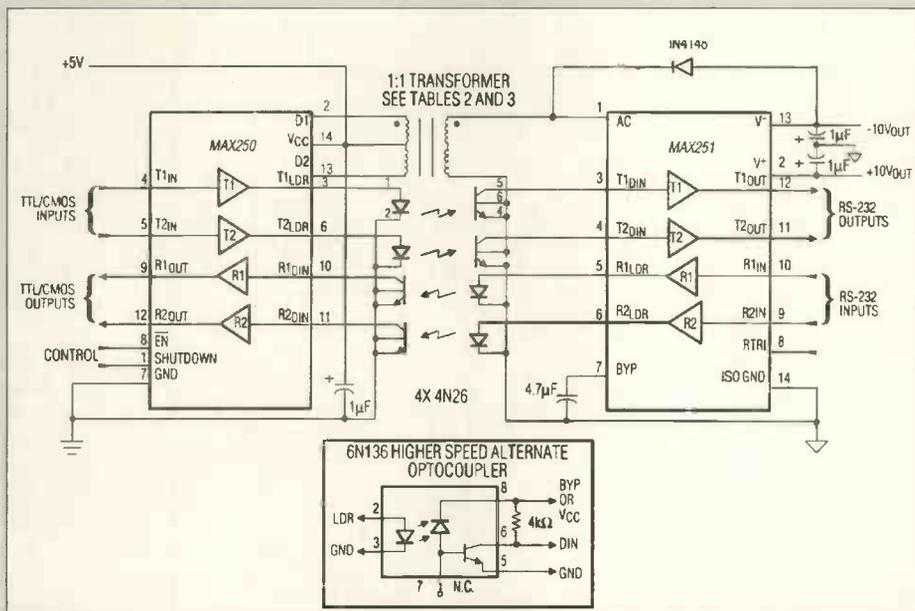
mistake the noise for a step input and fail to filter it out. To prevent this situation an input pre-filter can be added to the nonlinear filter as shown in Fig. 3. The prefilter's bandwidth is set by R_4 and C_3 . To minimize the prefilter's effect on settling time, its bandwidth is set ten times higher than the bandwidth of the nonlinear filter. At this higher bandwidth, the prefilter's effect on settling time is negligible.

Burr Brown International Ltd, 1 Millfield House, Woodshots Meadow, Watford, Hertfordshire, Phone 0923 33837.

Data line driving with isolation

In Maxim's article *What the EIA Specs Don't Tell You*, there is a circuit demonstrating that the MAX250 has a built-in transformer driver to provide power to the RS232 driver on the output side of the isolation barrier. A discussion of the effects of cable capacitance is included in this article, which appears in the company's data book, as are tables comparing EIA232D, EIA423A, EIA422A and RS485 standards. Transformer requirements are discussed in the MAX250 data sheet.

Maxim Integrated Products UK Ltd, 21c Horseshoe Park, Pangbourne, Reading, Berkshire RG8 7JW, Phone 0734 845255.



Pressure sensor controls motor

Feedback information needed to maintain a particular pressure is provided by a piezoelectric pressure sensor in this circuit from Motorola's booklet *Pressure Sensors (BR112/D)*.

Output from the *MPX5100* sensor is compared with a reference from divider *R11-13* to determine whether the *MC33033* controller turns the brushless motor on or off. Output of the sensor varies from about 0.5 to 4.5V for corresponding pressures from 0 to 100kPa. Hysteresis is built into the circuit to avoid chatter.

Also in the booklet are descriptions of the sensor technology, reliability issues and a variety of application circuits, one of which is a microprocessor-based manometer with LCD. A software listing is included. ■

Motorola Ltd, European Literature Centre, 88 Tanners Drive, Blakelands, Milton Keynes MK14 5BP, Phone 0908 614614.

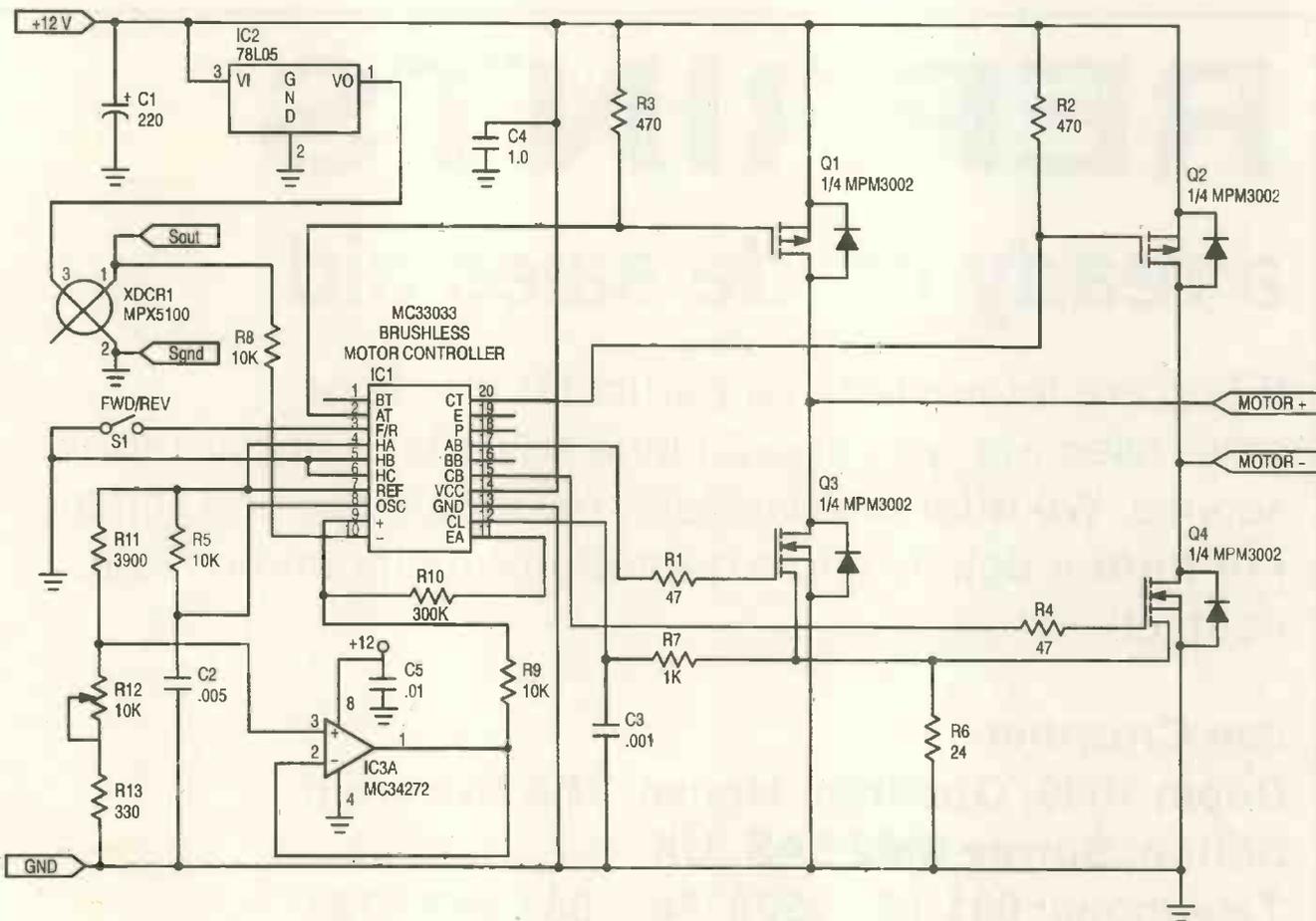
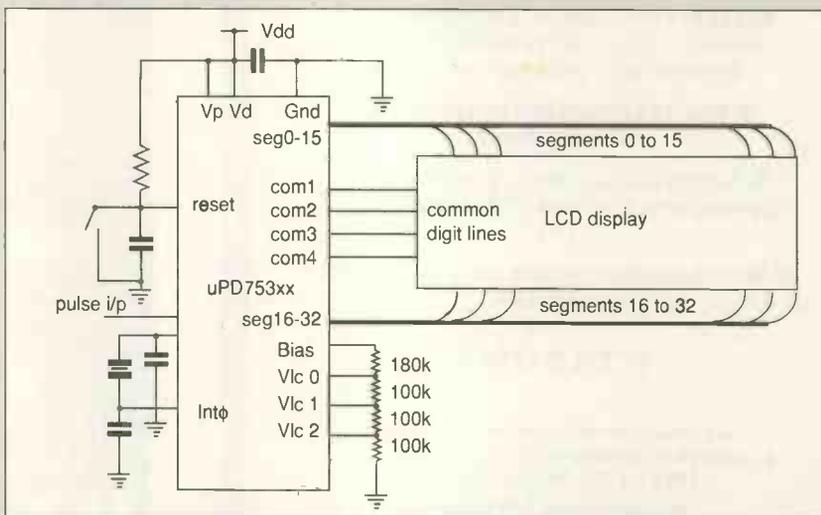
Controller has built-in LCD driver

In order to demonstrate uses of its four-bit microcontroller, NEC has produced note *AN183* describing how the device can form the basis of a general-purpose event counter with LCD.

From the circuit one can see that the μ PD753xx controller has a built-in LCD driver which in this case drives a 112

segment, four-backplane display. A full software listing is included in the description.

NEC Electronics UK Ltd, Cygnus House, Sunrise Parkway, Linford Wood Business Centre, Linford Wood, Milton Keynes MK14 6NP, Phone 0908 691133.





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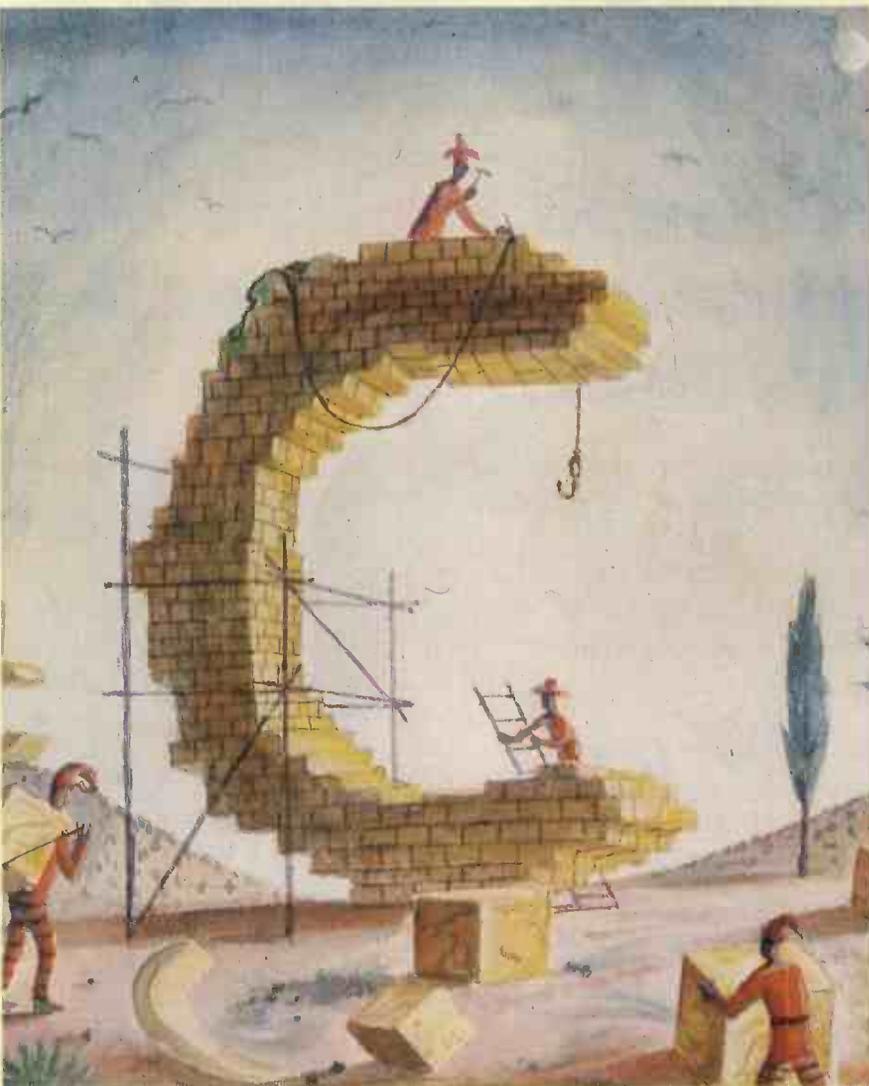
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No perfect waveforms in an imperfect world

Digital engineers who always visualise waveforms from gates and flip-flops as crisp, vertical rises are not living in the real world.

Serialized from Troubleshooting Analog Circuits by Robert Pease.

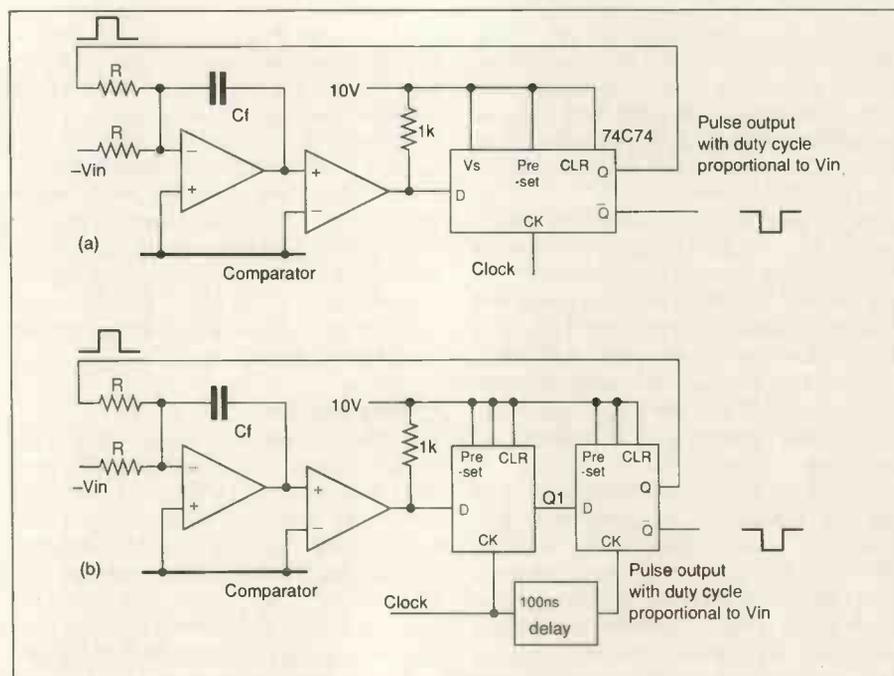
When signals are in a hurry, it can be crucial to remember that waveforms from gates and flop-flops have finite rise-times and delays and that the output of a gate may not change at the same time as the input.

For example, if the data input of a D flip-flop rises just *before* the clock pulse is applied, the output goes high. If the data input rises just *after*, the output goes low. But if the D input moves at the wrong time, the output can show "metastability", hanging momentarily halfway between high and low and taking several dozen nanoseconds to finally decide which way to go. Or, if the data comes a moment earlier or later, the result might be an abnormally narrow output pulse – a "runt pulse."

Feeding a runt pulse to another flip-flop or

counter can easily cause it to respond falsely and count to a new state that might be illegal. So they should be avoided – and make sure that you don't clock flip-flops at random times. Fig. 1a contains an example of a D flip-flop application that can exhibit this problem. When the comparator state changes at random times, it will occasionally change at precisely the wrong time – on the clock's rising edge – making the output pulse narrower or wider

Fig. 1a). Runt pulses cause problems in this simple A-to-D. The comparator state changes at random times. Occasionally, the state will change at precisely the wrong time – on the clock's rising edge – making the output pulse narrower or wider than normal. The problem can be solved by using two flip-flops with the clocks separated by a delay.



than normal. In certain types of A-to-D converters, the effect can cause non-linearity or distortion. A good solution is to use a delayed clock to transfer the data into a second flip-flop (Fig. 1b).

A classic example of a glitch (another name for a runt pulse) occurs when a ripple counter, such as a 7493, feeds into a decoder, such as a 7442. When the counter makes a carry from 0111 to 1000, for a few nanoseconds the output code will be 0000, and the decoder can spit out a narrow pulse of perhaps 6-8ns in duration corresponding to 0000. Even on a good scope, such a pulse can be just narrow enough to escape detection. If the decoder were merely feeding a led display, the sub-microsecond light pulse would never be seen, but if the decoded output goes to a digital counter, a false count can occur.

Digital system engineers often use logic analysers, storage scopes, and scopes with very broad band widths to look for glitches or runt pulses and the conditions that cause them. In analog systems, a logic analyser may not be an option, but these nasty narrow pulses often do exist, and must be considered.

Another fact to remember about digital ICs is that many cmos ICs have the same pinouts

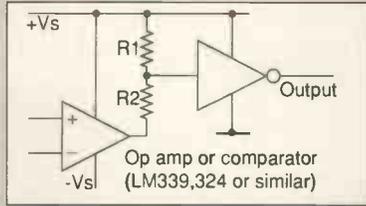


Fig. 2. Driving logic from an op amp operating from the usual large supply voltages requires an attenuator between the amplifier and the logic IC. The following equations show how to calculate the attenuator ratios.

$$\text{For CMOS, } \frac{R_1}{R_2} \equiv \frac{V_S}{-V_S}$$

Example: $+V_S = 5V, -V_S = -5V, R_1 = R_2 = 10k$

For TTL choose

$$\frac{-V_S}{R_2} = \frac{+V_S}{R_1} + 0.16mA$$

Example: $+V_S = 5V, -V_S = -5V, R_1 = 4.7k, R_2 = 2.2k$

For TTL choose

$$\frac{-V_S}{R_2} = \frac{+V_S}{R_1} + 0.16mA$$

Example: $+V_S = 5V, -V_S = -5V, R_1 = 4.7k, R_2 = 3k$

Of course, the unbuffered parts are faster with light capacitive loads, but the buffered ones are faster with heavy loads. So in a critical application, remember that substituting different vendors' parts can mess up a circuit.

Be careful when interfacing from linear ICs into digital ones. For example, an LM324 running on a single 5V supply does not have a lot of margin to drive cmos inputs. But an op amp running on ± 5 or $\pm 10V$ would need some kind of attenuation or resistive protection to avoid abusing the logic-device inputs (Fig. 2).

Likewise, it is considered bad form to overdrive the inputs of digital ICs just because they are protected by built-in clamp diodes. For example, you can make a pulse generator, as per Fig. 3, but it is considered poor practice to drive the inputs hard into the rail and beyond, as

will be the case if the capacitance is more than $0.01\mu F$ or the power supply voltage is higher than 6V. The circuits in Fig. 4 do as good a job without overdriving the inputs.

I have been cautioned that some LS-TTL parts such as DM74LS86 and 74LS75 are very touchy when their inputs are pulled below ground – even micro-momentarily – and give false readings for a long time. It sounds to me that there are probably currents being injected into tubs, as with LM339: Thus, these are unhappy parts if the inputs are overdriven much below ground.

Probing questions

A number of years ago, I was watching the negative transition of an ordinary TTL gate, and I was concerned by the way it was overshooting to $-0.4V$. I set up an attenuator with $1pF$ in the input leg (Fig. 5), and was astounded to see that if I looked at the waveform with an ordinary ($11pF$) probe, the overshoot occurred. But if I disconnected the probe from the gate output and connected it to the attenu-

ator output, the overshoot went away.

So, even when using a fairly high-impedance probe, be prepared for the possibility that looking at a signal can seriously affect it – even when it is as mundane and supposedly robust as a TTL output. You should be prepared to build your own special-purpose probes to allow you to see what is really going on.

D-to-A converters are generally docile

D-to-A converters are pretty simple machines, and usually give excellent results with few problems. If the manufacturer designed it correctly and it is not being misapplied, a D-to-A will not usually cause much grief.

But one area where D-to-As can cause trouble is with noise. Most D-to-As are not char-

Fig. 4. Addition of resistors to the circuit of (Fig. 3a) helps reduce overdrive, but the addition of diode clamps in the shunt leg of the attenuators (b) is more effective. If two two-input Nand gates are available, circuit (c) is the best implementation.

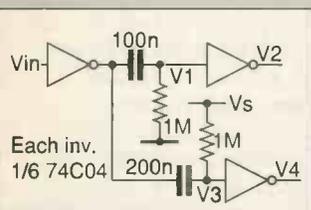
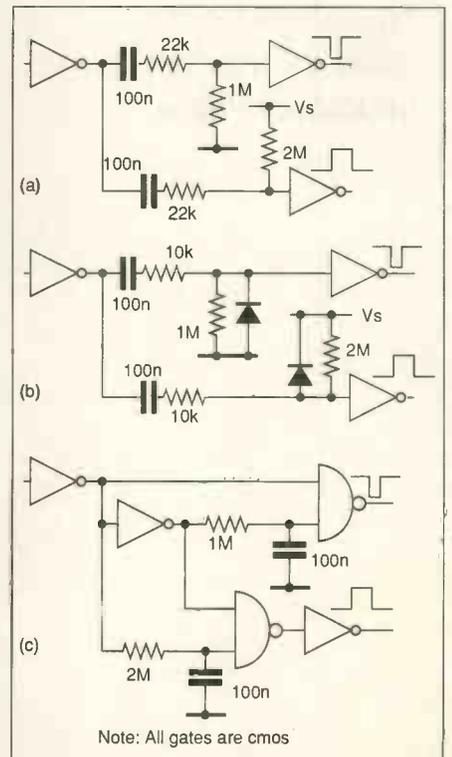
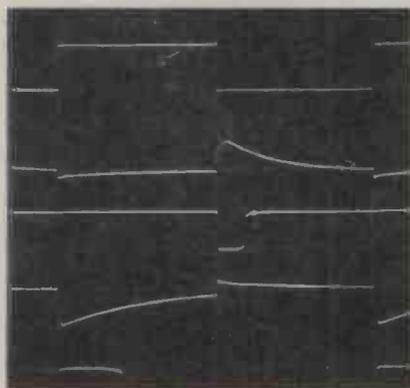


Fig. 3a) Cmos pulse generator, not recommended because, with the values shown, it overdrives the gate inputs excessively – as a waveform (b) indicate.



as TTL parts. For example, the 74193, 74LS193, and 74C193 have the same pinouts. On the other hand, some of the older cmos parts have pinouts that differ from those of similarly numbered TTL devices. The 74C86 pinout is the same as the 74L86 but differs from the 7486. Beware!

Similarly, some cmos devices have many – but not all – of their functions in common with those of their TTL counterparts. For example, the 74C74 has the same pinout and 95% of the same functions as the TTL 7474. Both follow, mostly, the same truth table, except that when both the preset and clear inputs are pulled low, the TTL device's outputs (Q and /Q) both go low, whereas the cmos outputs both go high.

In some cases there is a choice of buffered gate (CD4001BN), an unbuffered gate (CD4001), an unbuffered inverter (MM74HCU04), or a buffered inverter (MM74HC04). Sometimes, specifying a part number will bring an unbuffered part from one vendor and a buffered one from another.

acterised or guaranteed to reject high-frequency noise and spikes on the supply voltages. In some cases, DC rejection can be 80 or 100dB, but high-frequency noise on a supply can come through to the output virtually unattenuated.

So the system must be planned carefully. It might be a good idea, in a critical application, to use a completely separate power-supply regulator for a precision D-to-A. At least plenty of good power-supply bypass capacitors should be added right at the power supply pins – ceramic and tantalum capacitors.

Sometimes feeding signals to a D-to-A without passing them through buffers, allows the noise, ringing, and slow settling of the digital signals to pass through to the analogue side and show up on the D-to-A output. Nobody has a spec for rejection of the noise on D-to-A bit lines in either the high or low state: maybe vendors should specify this parameter, because some D-to-As are good; others are not.

I even recall a case where I had to preload the TTL outputs of a modular D-to-A's internal storage register with a 2k Ω resistor from each line to ground. Otherwise they would overshoot when going high and then recover with a long slow tail, an attenuated version of which would then appear on the D-to-A output.

On-chip buffers at a D-to-A's input can help cut down feed-through from the bit lines to the

analogue output, but will not completely eliminate it. The bus can move around incessantly, and capacitive coupling or even PC-board leakage will sometimes cause significant crosstalk into the analogue world. Even IC sockets can contribute to this noise. If it could be proved that such noise would not bother a circuit, it could be forgotten. The problem is that meaningful measurements of such effects can only be made on an operating prototype – computer modelling is not going to simulate this.

Multiplying D-to-As are popular and quite versatile. But a multiplying D-to-A's linearity can be degraded if the output amplifier's offset voltage is not very close to zero. Degradation of linearity has been estimated at 0.01% per millivolt of offset. Fortunately, low offset op amps are pretty cheap these days – at least a low offset op amp is cheaper than a trim-pot.

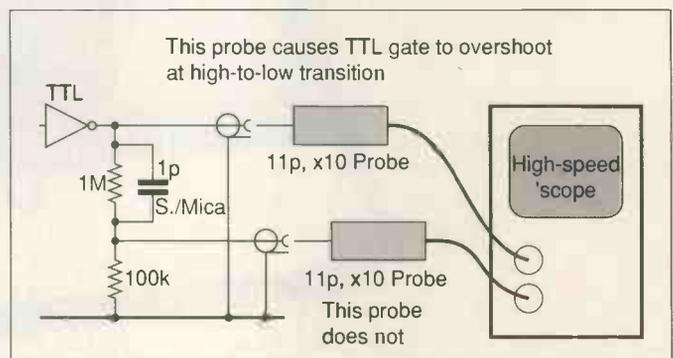
Another imperfection of any multiplying D-to-A is its AC response for different codes. Put in a 30kHz sine wave as the reference and no-one should really be surprised if the gain from the reference to the output changes by more than 1LSB

when going from a code of 1000 0000 to a code of 0111 1111.

In fact, if the frequency is above 5kHz, a 0.2% or larger error might result because the multiplying D-to-A's ladders, whose attenuation is a linear function of the input code at DC, become slightly non-linear at high frequencies due to stray capacitance. Non-linearity can be 0.2%, and the phase change as the input code is varied can exceed 2°, even with a 5kHz reference. So don't let these AC errors in multiplying D-to-As come as a surprise. ■

Another problem with D-to-As is the output

Fig. 5. An ordinary high-impedance probe can cause TTL outputs to appear to overshoot when observed, but not when they are not being observed. The effect can be eliminated by making a very-high impedance probe that presents only a 1pF capacitive load.



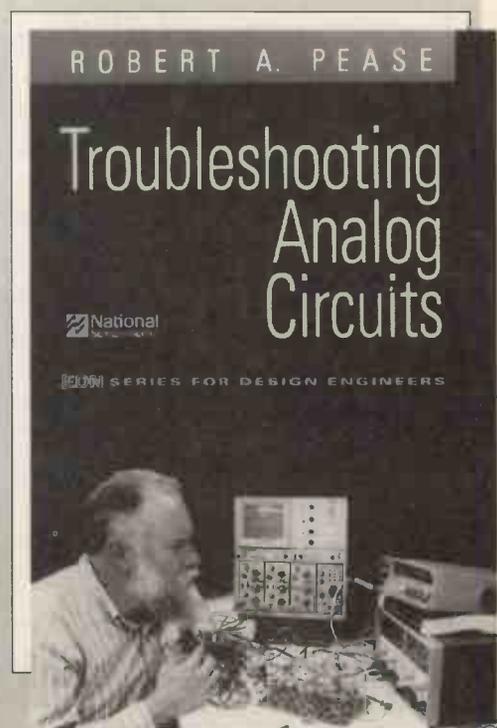
Troubleshooting Analog Circuits

In this book Bob Pease brings together many of the techniques he has developed over the years to expedite debugging and trouble-shooting analogue circuits.

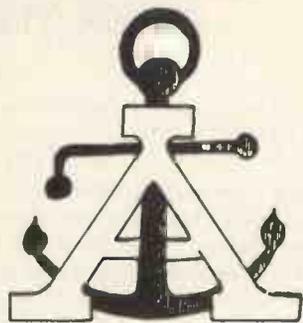
Based on his popular series in the US's EDN magazine, the book also contains new and updated material. Pease's approach to problem identification and isolation makes the book a useful aid to any analogue or digital engineer – whether experienced or not.

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8GHz transistor array. Four field-configurable transistor arrays by Harris in the *HFAXXX* series contain 8GHz n-p-n and/or 5.5GHz p-n-p devices, which are designed for analogue circuitry. Each transistor can work independently, being isolated by means of the Harris UHF1 trench and bonded-wafer isolation technique, their characteristics tracking with temperature. *Pspice* models are available. Harris Semiconductor (UK), 0276 686886.

A-to-D & D-to-A converters

12-bit A-to-D. Analog's *AD1674*, an upgrade from the *AD574/674*, is claimed to be the only device in a 28-pin dip or SOIC to include a wide-band (1MHz) sample-and-hold amplifier. Conversion rate is 100ks/s/10 μ s; THD typically -90dB; signal-to-noise plus distortion -70dB and intermodulation distortion -80dB. Three-state buffers allow direct connection to external digital circuitry. Analog Devices, 0932 253320.

Discrete active devices

Protected mosfet. A 60V power mosfet from Zetex operates in avalanche mode during an over-voltage state, absorbing a repetitive 5mJ. Continuous current handling is 0.6A. With a drain current of 1mA, the 4206AV gate-source threshold is between 1.3V and 3V, on resistance at 10V being 1 Ω . Operating temperature -50°C to 150°C. Zetex plc, 061 627 4963.

Digital signal processor

Fast DSP. Although otherwise identical to Analog's *ADSP-21020* family, *ADSP-21020KG* is about four times as fast as other devices that cost roughly the same. It features 0.58ms execution time for a 1024-point complex FFT, an operation that can take up to 2ms with other devices. Other features remain the same as in the earlier version. Analog Devices, 0932 253320.

Linear integrated circuits

Quad audio switch. Meant for audio signal routing, Analog's *SSM-2404* quad switch IC contains four single-pole, single-throw bilateral switches in its 20-pin plastic or SOIC package. An on resistance of typically 45 Ω , combined with 0.0065% THD from 2V RMS at 1kHz, off-state isolation and cross-talk of -100dB and -94dB, and wide-band noise density of 0.6 μ V peak combine the advantages of analogue switching with better audio performance. Analog Devices, 0932 253320.

Vari-gain amplifiers. 35MHz bandwidth, -60dBc total harmonic distortion and 1.4nV/ \sqrt Hz input noise are features of Analog's *AD600* and *AD602* dual-channel variable-gain amplifiers for AGC application. Gain is continuously variable over a 40dB range by a linear control voltage with a scale factor of 32dB/V. Gain ripple is \pm 0.2dB. Amplifiers can be cascaded to give a total control range of 80dB. Analog Devices, 0932 253320.

Modulation IC. Generating AM, FM, PM or compound carriers at UHF, the RF2402 from RF Micro Devices contains all necessary components including differential amplifiers for base-band inputs, a 90° hybrid phase splitter, limiting local-oscillator amplifiers, mixers, combining amplifier and output amplifier. Anglia Microwaves Ltd, 0277 630000.

45MHz op-amps. Dual and quad op-amps by Linear Technology are in SO8 (dual) or the 16-lead SO package. Settling time is 90ns for a 10V step and the devices are stable into any capacitive load. Minimum open-loop gain is 70dB into 500 Ω from a trimmed input offset of 2mV. Slew rate is 400V/ μ s. Micro Call Ltd, 0844 261939.

Rail-to-rail op-amps. With what is claimed to be a unique combination of rail-to-rail input and output, National's *LMC6482* and *6484* CMOS op-amps accept single supply lines of between 3V and 15V. Input offset voltage is 750 μ V and common-mode rejection ratio a minimum of 70dB. The devices work into loads down to 600 Ω . National Semiconductor, 01049 81 41-10 35 50

RF synthesisers. The first in a series of RF products by National is the PLLatinum series of phase-locked-loop RF synthesisers operating at

frequencies up to 2GHz. Dead zone and noise are reduced and power consumption is 18mW. Supply voltage is 2.7V-5.5V. National Semiconductor, 01049 81 41-10 35 50

Single-chip AM/FM receivers.

TEA5710/1/2 by Philips are single-chip receivers with integrated mosfet RF stages and IF/mixer/demodulator stages for AM and FM reception. In addition, *TEA5711/2* have stereo decoders and the *5712* an IF counter output for use with PLL frequency synthesisers. AM sensitivity is 1.6mV/m for 26dB s:n, tuning range being up to 30MHz. Sensitivity on FM is 2 μ V at 26dB s:n. Philips Semiconductors Ltd, 071 436 4144.

Precision op-amps. To condition low-level signals from battery-powered sensors with high accuracy, Texas has the *TLC2272* and *TLC2274* dual and quad CMOS op-amps. They are packed in Texas's Thin Shrink Small Outline Packages with a height of 1.1mm and are intended to amplify signals from automotive devices, temperature and pressure sensors, etc. A rail-to-rail output swing and a 950 μ V input offset, accompanied by low noise and high input impedance, allow the devices to replace more expensive instrumentation amplifiers. Texas Instruments, 0234 223252.

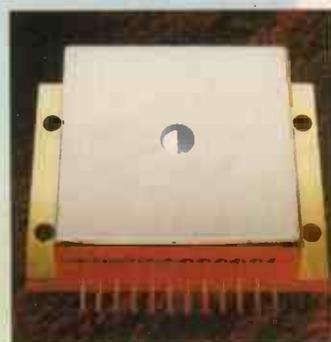
Logic building blocks

Pulse-delay chip. Digitally programmable delay between an input trigger edge and an output edge is provided by Analog's *AD9505* pulse-edge delay vernier chip. Delay is between 1.5ns and 25ns and is controlled by a built-in 8-bit D-to-A converter to 10ps resolution. At 60MHz, differential non-linearity is \pm 1LSB. Incremental delays are programmable on the fly. Analog Devices, 0932 253320.

Low-power clock. Clock circuits needing only 5 μ A at 32kHz are realisable using the Harris *HA7210* crystal-controlled oscillator chip in 2-7V systems. It is programmable from 10kHz to 10MHz and is provided with a disable pin to switch the output to a high-impedance state for test or when several oscillators are in use. Harris Semiconductor (UK), 0276 686886.

Memory chips

Video field memory. Oki's *MSM514221A* is a 1Mbit video field memory with a cycle time of 30ns and



Laser diode. Sony's *SLD323V* near-infrared laser diode emits 800mW through a 100 μ m aperture at 790-830nm - enough optical density for material processing and medical applications. Power taken is 1.3A at 3V, threshold current being 0.3A. Sony Semiconductor Europe Ltd, 0784 466660.

is designed for first-in, first-out serial data access in digital or high-definition television involving special effects or freeze frame. Typical current consumption is 50mA. Highland Electronics Ltd, 0444 236000.

16Mbit dram. *HY5116XXX* series drams by Hyundai are in CMOS silicon-gate technology and offer access times down to 60ns. They need a 5V supply and interface directly with logic such as Schottky TTL. Hyundai Electronics Europe, 081 741 8634.

Mixed-signal ICs.

Image compression. A chipset by Zoran allows the development of full-motion video and still-image compression products, including computer boards for multi-media application, digital still cameras and peripherals. The set includes a discrete cosine transform processor and an image coder/decoder perform compression complying with the JPEG baseline standard. Emphasis is on low cost.

Optical devices

3W laser diode. A GaAlAs laser diode from Sony, the *SLD3252T* has maximum optical output of 3.3W in the 790-830nm band. It is contained in a 12-pin flat pack with integral photodiode, thermistor and thermoelectric cooler so that its emission wavelength is adjustable by

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Oscillators

Oscillators. Oscillators in the new Champion Technologies range crystal types compatible with TTL, CMOS and ECL, SM clock oscillators working up to 40MHz, voltage-controlled oscillators, temperature-controlled types and high-frequency TTL clocks for 130MHz working. Acal Electronics Ltd, 0344 727272.

varying the temperature. Sony Semiconductor Europe Ltd, 0784 466660.

Power semiconductors

Power-factor correction. Micro Linear's *ML4830* is a power factor corrector for fluorescent-light ballasts. It uses average-current sensing, with a current-fed multiplier and over-voltage protection, producing power factors of better than 0.99 at over 99% efficiency. The ballast section allows programmable pre-heat and tube out-of-socket interrupt times, controlling light output by means of tube current feedback and either frequency or pulse-width modulation. Ambar Components Ltd, 0844 261144.

Battery power controller. Replacing five ICs normally used for power conversion and regulation in Battery-powered equipment, Micro Linear's *ML4862* has a built-in DC-to-DC converter and regulator. Battery voltage from 5.5V to 20V is controlled to 5V or 3.3V, depending on configuration. The chip provides mosfet drive to turn off unused sections of circuit and will disconnect the battery on connection of an AC adaptor. Ambar Components Ltd, 0844 261144.

Mosfet power switch. Hitachi's *2SJ317* is a p-channel mosfet intended use in portable equipment, in which some of the circuitry can be switched off when not needed to reduce power consumption. On resistance of this 12V, 2A device, which has a 2.5V gate drive, is 0.28 Ω at 4V. Surface-mounted, the mosfet measures only 4.5mm by 2.5mm by 1.5mm. Hitachi Europe Ltd, 0628 585000.

5A step-down converter. Maxim's *MAX724* pulse-width-modulated DC-DC converter has an internal 6.5A switch and is optimised for step-down working. It takes a 8V-40V input (60V for the /H version), the output being settable to any voltage between 2.5V and the input voltage by two resistors. Maxim Integrated Products Ltd, 0734 845255.

Voltage dropper. Semtech's *EZ Dropper* provides a constant 1.7V or 2V drop while passing currents 0.1A to 1A. Two versions give 3V and 3.3V outputs from 5V input, while the third provides an adjustable output from 2.7V to 3.3V, external components extending the range of adjustment to 1.5-10V. No filtering is needed, since the devices produce no EMI or spikes. Semtech Ltd, 0592 773520.

Switching regulator. *UC2575* regulators contain all the active functions needed to form a buck switching regulator in one chip, giving $\pm 3\%$ tolerance and driving a 1A load. A 40V output is standard, but there is also a 60V version, the *UC2575HV*, both types producing 5V, 12V or 15V outputs. Unitrode (UK) Ltd, 081 318 1431.

PASSIVE

Passive components

Ceramic PSU capacitors. *MF30* capacitors from AVX are high-value multilayer types for power input and output filters in resonant converters, DC-DC converters and other PSU types. In 50V, 100V and 200V DC versions, capacitance covers the range 1 μ F to 100 μ F, with very low ESR and ESL figures. AVX Ltd, 0252 336868.

Ceramic discs. *Cera-Mite* high-voltage disc ceramic capacitors cover the 100pF to 0.01 μ F range of values at voltages from 1kV DC to 10kV DC in seven series, including the EIA class III types. The components are tested to 4kV AC and conform to most of the international standards for coupling, by-pass and RFI/EMI filters.

Two of the series possess ultra-stable capacitance over a range of temperature, frequency and voltage variation. Acal Electronics Ltd, 0344 727272.

SM inductors. Toko's D7 range of low-profile, surface-mounted inductors are less than 5mm high. Inductance values from 1 μ H to 470 μ H are available in current ratings 0.195-3.12A. There is also a magnetically shielded version with an external ferrite ring. Coil terminations are separate from the mounting pads. Cirkit Distribution Ltd, 0992 444111.

HF crystals. Crystals in the *AB HC49/SD3* range from Flint comes in 164 frequencies from 3.2MHz to 48MHz, with others up to 70MHz on request. The devices withstand infrared, wave and vapour-phase soldering for surface-mounting. Flint Distribution, 0530 510333.

Surge suppressors. Designed to protect automotive electronics against huge transient, Harris's TVS chips will take a 24.5V jump start for five minutes and up to 25J of transient energy in less space than a conventional zener diode. *V18AUMLA1210/1812/2220* are rated at 3j/10j/25j and the surface mounting eliminates package-lead inductance for quick response. They have a symmetrical breakdown and go into a shorted state when over-stressed. Harris Semiconductor (UK), 0276 686886.

Chip inductors. Surface-mounted chip inductors from Murata exhibit a 2% tolerance, allowing many oscillator types to dispense with adjustment components. *LQS33N* coils have a Q of over 80 and values from 1 μ H to 100 μ H, also possessing an integral ferrite shield, which assists with high mounting density. Murata Electronics (UK) Ltd, 0252 811666.

Electrolytics. ST electrolytic capacitors made by Nichicon operate from -55°C to 105°C, having a load life of 32,000 hours at 55°C. Values are 0.1 μ F to 220 μ F at working voltages of 6.3V to 50V. Leakage current is 3 μ A and maximum ripple 130mA. Nichicon (Europe) Ltd, 0276 685393.

EMI cores. For both round and flat cables, the Tokin ferrite cores for noise suppression are hinged and non-hinged types working up to 300MHz with nylon or stainless steel clamps. Pedoka Electronics Ltd, 0493 440047.

SMPS capacitors. Vitramon's *VK50* and *VK60* series boxed radial capacitors are meant for use in switched-mode power supplies, delivering best ESR and ESL performance between 100kHz and 1MHz. Ceramic dielectrics provide

tolerances to $\pm 5\%$ and capacitances up to 18 μ F. Vitramon Ltd, 0628 524933.

Displays

Bicolour led. Dialight's series *550-3605* circuit board indicator has two leds, green and yellow, in a right-angle 5mm package. Luminous intensity is 5mcd at 10mA. BLP Components Ltd, 0638 665161.

LC monitor. Meant for outside-broadcast and ENG use, the *CM-L500* active-matrix thin-film transistor liquid-crystal video monitor by Hitachi Denshi measures 102mm by 77mm, which gives 720 by 479 pixels, and works for two hours on an *NP-1* NiCd battery or can use a 12V source. Video input is NTSC or pal at 1V pk-pk into 75 Ω . Horizontal and vertical resolution is 350 tv lines, pixels being arranged as RGB triangles. Hitachi Denshi (UK) Ltd, 081-202 4311.

Filters

Satellite Tx/Rx filters. Intended for use in satellite communications, Matthey's new range includes base-band video filters, group-delay equalisers, pre-emphasis and de-emphasis networks and a 70MHz band-pass filter for IF shaping. Matthey Electronics, 0782 577588.

Hardware

Mount adaptors. Chip-specific adaptors from Genalog convert surface-mounting versions of to the pin grid array format with pin compatibility, so that cheaper quad flat packs can be used on a board designed for PGAs. Other adaptors are available to convert from SM packages to other types of through-hole mounting. Genalog Ltd, 0580 753754.

Liquid dispenser. *Auto Tube* from Fisnar dispenses liquids or pastes from a manufacturer's "toothpaste" tube without mess or waste. A foot valve or timer control the dose and the unit is hand-held or can be bench mounted. I Fisnar Inc., 0101 201 796-1477.

Controlled fans. Intelligent temperature control of Papst's *6200 NT* cooling fans reduces noise and wear. Normally, the fans run at between half and full speed, only reaching full speed in exceptional conditions. Maximum free air flow is 410 cubic metres/hour with a noise pressure of 52dBa in the optimum speed range. All electronics are housed within the fan. Papst plc, 0264 342200.

Heat sinks. Penguin coolers by Wakefield are direct-mounting heat sinks for use with Intel, AMD and Motorola microprocessors and

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chipsets. There is also an adhesive-attached range of sinks with extremely flat bases that need no thermal compound. Warth International Ltd, 0342 315044.

Instrumentation

Digital phase meter. At frequencies from 5Hz to 500kHz, the Clarke-Hess 6000 digital phase meter measures phase angle to a resolution of 10millidegrees, automatically selecting amplitude and phase range. Signal and reference channels accept 10mV-350V inputs, regardless of waveform, and there is an optically isolated IEEE-488 interface. Lyons Instruments Ltd, 0992 467161.

Waveform generator. Any desired shape of waveform is available from

the Philips *PM5150* arbitrary waveform generator. It has a library of 20 waveforms, but also 31K of non-volatile memory for storing waveforms defined by the user. Waveforms are synthesised to 12 bits and can be defined by drawing by keypad or mouse and stored, whereupon adding subtracting and multiplying can supply composite waveforms. They can also be down-loaded from a DSO or produced on a PC and loaded from there. Philips Test & Measurement, 0923 240511.

Audio analyser. UPD, a versatile digital/analogue audio analyser by R&S, has its own PC. There are built-in generators with less than 115dB distortion and two-channel measurement for stereo. The FFT analyser samples at 32-768kHz, inputs and outputs are serial to 32 bits and parallel to 27 bits. New functions are simply loaded from software, with no extra hardware needed. Rohde & Schwarz UK Ltd, 0252 811377.

1.3GHz counter. The *SC-230* microprocessor-based frequency counter from Saje includes RS232 communication as standard. Frequency range is 5Hz-1.3GHz at a sensitivity of 10mV. Multi-gating provides readings of minimum, maximum, average or difference and a hold facility; RPM is also measured. Display is a backlit LCD. Saje Electronics, 0223 425440.

Tek TDS 400 improvements. At no extra cost, Tektronix offers



enhancements to its *TDS 400* series of digitising oscilloscopes. Roll mode enables the capture of low-speed events; template generation and test allows limit specification for displayed waveforms; and record length is increased from 5000 to 15000 points, with 60,000 as a low-priced option. Further options include an RS232/Centronics interface with spooler, a printer pack and an FFT pack. Tektronix UK Ltd, 0628 486000.

1GHz oscilloscope. With 5K memory per channel, 8-bit flash converters, pattern and glitch trigger, the models *9320* and *9324* digital oscilloscopes by LeCroy also offer a range of waveform-processing capability; options include FFT, enhanced resolution, waveform maths and averaging. The internal floppy drive and memory card interface are both dos-compatible and an internal graphics printer gives a screen dump in 10s. LeCroy Ltd, 0235 533114.

Literature

Multimeter/software. A brochure from AVO describes a range of multimeters and the *Avovins* software which allows the multimeters to download to a PC via an adaptor, the data being communicated by way of infrared links. AVO International, 0304 202620.

Amplicon 1993. Amplicon Liveline's first 1993 catalogue is 10% updated and now shows which products are subject to BS5750 quality checks, the company being accredited in 1992. New this year are the 200 series data acquisition system, OWS and TW DC-DC converters, a new series of DAP boards and software and some new software packages. Amplicon Liveline Ltd, (Free)0800 525 335.

Power supplies. In 160 pages, Gresham's Power Supply Guide provides data and applications information on a range of AC/DC, DC/DC and DC/AC supplies, including some from Power General and Philips. Gresham Power Electronics Ltd, 0722 413060.

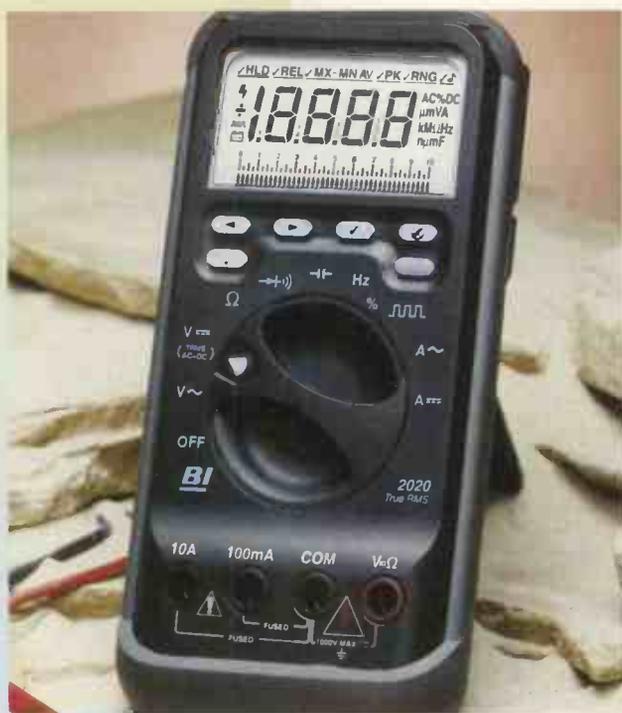
Sensors. Laser-based analogue sensors, optical-fibre sensors and photosensors are all covered in Matsushita's shortform colour catalogue, which features the new LM laser sensor range with the glass lens for accuracy. Resolution accuracy of

the LM 300 is 0.2µm. Matsushita Automation Ltd, 0908 231555.

Power supplies

DC-DC converters. Working from 20-60V input, OWS DC-DC converters from Amplicon come in the 2in by 2in standard EMI/RFI-shielded case, supply 25W and provide 5V or 12V output. With continuous short-circuit protection and remote on/off control, output voltage accuracy is ±1%, stabilisation and regulation ±1%, isolation voltage is 50V DC and switching frequency 300kHz. Amplicon Liveline Ltd, (Free)0800 525 335.

Lithium batteries. Custom-made, high-energy density lithium-thionyl chloride batteries are available from Battery Engineering in a range of sizes from 7mm to 64mm diameter in discharge capacities of 0.08 to 135 amp-hours. They are made in stainless steel with hermetic seals, and come with a variety of terminations. These batteries retain up to 80% of capacity after 10 years and operate at temperatures between -55°C and 200°C. Battery Engineering Inc., 0101 617 361-7555.



BARGAINS – Many New Ones This Month

AMSTRAD 3' DISK DRIVE Standard replacement or, why not have an extra one? £20, Order Ref. 20P28.

SILENT EFFORTLESS MOVEMENT with our 14mm ballrace complete with removable spindle. 4 for £1, Order Ref. 912.

6-12V AXIAL FAN is a Japanese-made 12v DC brushless axial fan, 93mm square. Its optimum is 12 but it performs equally well at only 6v and its current then is only 100mA. £4, Order Ref. 4P65.

FM CORDLESS RADIO MIKE, hand-held battery-operated, a good performer with a really professional look. Has usual shaped body and head and is tuneable to transmit and be picked up on the FM band of any radio. Yours for only £8.50, Order Ref. 8.5P1.

PUSH BUTTON EDGE SWITCH gold-plated contacts, 1A 50V. Top push counts up, bottom push counts down. BCD switching. Switches clip together. 2 for £1, Order Ref. 915.

POWER SUPPLY WITH EXTRAS mains input is fused and filtered and the 12V dc output is voltage regulated. Intended for high-class equipment, this is mounted on a PCB and, also mounted on the board but easily removed, are 2 12V relays and a Piezo sounder. £3, Order Ref. 3P80B.

ULTRASONIC TRANSDUCERS 2 metal cased units, one transmits, one receives. Built to operate around 40kHz. Price £1.50 the pair, Order Ref. 1.5P/4.

100W MAINS TRANSFORMER normal primary 20–0–20 at 2.5A, £4, Order Ref. 4P24. 40V at 2.5A, £4, Order Ref. 4P59. 50V at 2A, £4, Order Ref. 4P60.

PHILIPS 9" HIGH RESOLUTION MONITOR black & white in metal frame for easy mounting, brand new, still in maker's packing, offered at less than price of tube alone, only £15, Order Ref. 15P1.

16-CHARACTER 2-LINE DISPLAY screen size 85mm x 36mm, Alpha-numeric LCD dot matrix module with integral microprocessor made by Epson, their Ref. 16027AR, £8, Order Ref. 8P48.

INSULATION TESTER WITH MULTIMETER internally generates voltages which enable you to read insulation directly in megohms. The multimeter has four ranges. AC/DC volts, 3 ranges DC milliamps, 3 ranges resistance and 5 amp range. These instruments are ex British Telecom, but in very good condition, tested and guaranteed OK, probably cost at least £50 each, yours for only £7.50, with leads, carrying case £2 extra, Order Ref. 7.5P/4.

MAINS 230V FAN best make "PAPST" 4½" square, metal blades, £8, Order Ref. 8P8.

2MW LASER Helium Neon by PHILIPS, full spec. £30, Order Ref. 30P1. Power supply for this in kit form with case is £15 Order Ref. 15P16, or in larger case to house tube as well £18, Order Ref. 18P2. The larger unit, made up, tested and ready to use, complete with laser tube £69, Order Ref. 69P1.

½ HP 12V MOTOR – THE FAMOUS SINCLAIR C5 brand new, £15, Order Ref. 15P8.

SOLAR CHARGER holds 4 AA nicads and recharges these in 8 hours, in very neat plastic case, £6, Order Ref. 6P3.

AIR SPACED TRIMMER CAPS 2–20pf ideal for precision tuning UHF circuits, 4 for £1, Order Ref. 818B.

45A DOUBLE POLE MAINS SWITCH mounted on a 6x3½ aluminium plate, beautifully finished in gold and with pilot light. Top quality, made by MEM, £2, Order Ref. 2P316.

MAINS ISOLATION TRANSFORMER stops you getting "to earth" shocks. 230V in and 230V out. 150watt upright mounting, £7.50, Order Ref. 7.5P/5 and a 250W version is £10, Order Ref. 10P97.

MINI MONO AMP on PCB. Size 4" x 2" with front panel holding volume control and with spare hole for switch or tone control. Output is 4 watts into 4-ohm speaker using 12V or 1 watt into 8-ohm using 9V. Brand new and perfect, only £1 each, Order Ref. 495.

AMSTRAD POWER UNIT 13.5V at 1.9A enclosed and with leads and output plug, normal mains input £6, Order Ref. 6P23.

DC VOLTAGE REDUCER 12V-6V @ 500ma on quite small pcb with heat sink. £1, Order Ref. 916.

80W MAINS TRANSFORMERS two available, good quality, both with normal primaries and upright mounting, one is 20V 4A, Order Ref. 3P106, the other 40V 2A, Order Ref. 3P107, only £3 each.

PROJECT BOX size approx 8" x 4" x 4½" metal, sprayed grey, louvred ends for ventilation otherwise undrilled. Made for GPO so best quality, only £3 each, Order Ref. 3P74.

12V SOLENOID has good ½" pull or could push if modified, size approx 1½" long by 1" square, £1, Order Ref. 232.

15W 8-OHM 8" SPEAKER & 3" TWEETER made for a discontinued high-quality music centre, gives real hi-fi, and only £4 per pair, Order Ref. 4P57.

GOODMANS 20W, 5", 4ohm speaker, made for Ford. £3, Order Ref. 3P145.

20W, 4 OHM TWEETER. £1.50, Order Ref. 1.5P9.

6½", 12W MAX, 8 ohm, Japanese made (Sanyo) speaker, £1.50, Order Ref. 1.5P11.

6½", 8 OHM SPEAKER with built in tweeter, £1, Order Ref. 897.

0-1MA FULL VISION PANEL METER 2¾" square, scaled 0100 but scale easily removed for re-writing. £1 each, Order Ref. 756.

PROJECT BOX a first-class, Japanese two-part moulding size 95 x 66 x 23mm.

This is nicely finished and very substantial. You get 2 for £1, Order Ref. 876.

AM/FM RADIO CHASSIS with separate LCD module to display time and set off alarm. This is complete with loudspeaker but is not cased. Price £3.50. Order Ref. 3.5P5.

2, 3 AND 4-WAY TERMINAL BLOCKS the usual grub screw types. Parcel containing a mixture of the 3 types, giving you 100 ways for £1, Order Ref. 875.

FULLY ENCLOSED MAINS TRANSFORMERS with 2m 3 core lead terminating with a 13A plug. Secondary rated at 6v 4A. Brought out on a well insulated 2 core lead terminating with insulated push on tags. £3, Order Ref. 3P152. Ditto but 8A, £4, Order Ref. 4P69.

SWITCHED BC CORD GRIP LAMP HOLDERS. Always useful. A good make, 3 for £1, Order Ref. 913.

2M 3-CORE LEAD terminating with flat pin instrument socket, £1, Order Ref. 879. Ditto but with plug on the other end so that you could use this to extend an instrument lead. £1.50, Order Ref. 1.5P10.

INFRA RED RECEIVER CONTROLLER, made by Thorn to channel switch their T.V. receivers. Mounted on panel with luminous channel indicator, mains on/off switch, leads and plugs all yours for £2, Order Ref. 2P304.

HIGH QUALITY KEY SWITCH, single pole on/off or change-over through panel mounted by hexagonal nut. Complete with 2 keys. Regular price £3, our price £1.50, Order Ref. 1.5P12.

DIGITAL MULTI TESTER M3800, single switching covers 30 ranges including 20A ac and dc, 10 meg input impedance, ¾ LCD display. Currently advertised by many dealers at nearly £40, our price only £25, Order Ref. 25P14.

ANALOGUE TESTER, input impedance 2K ohms per volt. It has 14 ranges, ac volts 0-500, dc volts 0-500, dc current 500 micro amps at 250 milliamp, resistance 0-1 meg-ohm, decibels 20 ±56dB. Fitted diode protection, overall size 90x60x30mm. Complete with test prod's, price £7.50, Order Ref. 7.5P8.

2" 50 OHM LOUDSPEAKER, replacement for pocket radio, baby alarm, etc. Also makes good pillow 'phone. 2 for £1, Order Ref. 905.

LCD CLOCK MODULE, 1.5v battery-operated, fits nicely into our 50p project box. Order Ref 876. Only £2, Order Ref 2P307.

AMSTRAD KEYBOARD MODEL KB5, very comprehensive, has over 100 keys, £5, Order Ref. 5P202.

SENTINEL COMPONENT BOARD, amongst hundreds of other parts, this has 15 ICs, all plug in so don't need de-soldering. Cost well over £100, yours for £4, Order Ref. 4P67.

9V 2.1A POWER SUPPLY, made for Sinclair to operate their 128K Spectrum Plus 2. £3, Order Ref. 3P151.

SCREWDRIVERS – pocket sized. Will save you having to worry where you left the last one! 10 for £1, Order Ref. 909.

STEPPER MOTOR BARGAIN This is just a mini motor, 12v operated and 7.5° step angle. Offered at the very low price of only £1, Order Ref. 910.

STANDARD CASSETTE MOTOR for 9v recorder players. This is brushless and has internal electronics to facilitate speed changes and reverse. £1.50 each, Order Ref. 1.5P14.

CAR SOCKET PLUG tubular construction to take small pcb etc. £1, Order Ref. 917.

THIS COULD SAVE YOU EXPENSIVE BATTERIES an in-car unit for operating 6v radio, cassette player, etc. from car lighter socket. £2, Order Ref. 2P318.

READY BUILT .5W FM TRANSMITTER tested and working. Very compact unit, with electret microphone, 3v operated. £6, Order Ref. 6P29. Will fit, with batteries, in our project box, Order Ref. 876.

METAL PROJECT CASE nickel plated, size 1½x5½x2½", so ideal to take the Philips laser with its power supply or just a power supply. Has Instrument type mains input plug, output socket and built in on/off switch. £7.50, Order Ref. 7.5P9.

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First reach your conclusion!

The father of a deeper insight into electrical phenomena and a precursor of Coulomb, or a scientific fraud?

Leonid N Kryzhanovskiy assesses the contribution of Charles Stanhope.

Experimental philosophy

Developing the methodology of Sir Isaac Newton (1643-1727), life-long President of the Royal Society of London from 1703, Francis Hauksbee (c 1666-1713), curator and demonstrator of experiments of the Royal Society, wrote in the preface to his book:

"The Learned World is now almost generally convinc'd, that... there's no other way of Improving Natural Philosophy, but by Demonstrations and Conclusions founded upon Experiments judiciously and accurately made"¹.

This remains the guide-line of modern scientists. The situation is presented more specifically by Richard F Feynman (1965 Nobel Prize in physics):

"In general we look for a new law by the following process. First we guess it. Then we compute the consequences of the guess to see what would be implied if this law that we guessed is right. Then we compare the result of the computation to nature, with experiment or experience, compare it directly with observation, to see if it works. If it disagrees with experiment it is wrong. In that simple statement is the key to science"².

The author is with the Popov Central Museum of Communications, St. Petersburg.

In his book published in London in 1779, Charles Stanhope presents an experiment⁴ which was at the core of much of his thinking on electrostatic induction.

An uncharged conductor AB of circular cross-section is suspended horizontally on insulators (silk strings or ribbons) a certain distance apart from a small charged ball O located at the same height (in Stanhope's set-up, a metal ball was connected by a stud to the prime conductor of an electrostatic generator). It has long since been recognised⁵ that if the ball O is charged, say, positively, the end A of the rod AB is found to be charged negatively while the other end, B , positively. Stanhope assumed an electrical neutral N to exist between A and B . Using an electroscope in the form of two small cork balls fixed at the ends

The great tragedy of science - the slaying of a beautiful hypothesis by an ugly fact. TH Huxley

of a linen thread suspended at its middle point, he looked for the position of N , at which the electroscope threads stayed plumb, for different values of OA .

By this series of experiments, Stanhope wanted to confirm his hypothesis that the "density of electricity of the electrical atmosphere" in which a body is immersed varies inversely as the square of the distance from the body creating this atmosphere⁶.

In modern terms: the linear density $\tau(x)$ of the charge induced on the conductor AB by a point charge located at the point O is inversely proportional to the square of the distance x from the point O to a point P on the rod, $\tau(x) \approx 1/x^2$ ($x = OP$, $OA \leq x \leq OB$).

Stanhope represents the actual charge on the rod as a fictitious linear charge (we use this approximation in all the calculations below) and the prime conductor-stud-ball system, as a

Pedigree

Viscount Charles Mahon (1753-1816) became a Fellow of the Royal Society in 1772, an early honour favoured by the fact that the 19-year-old Charles had solved a prize-winning problem on a pendulum clock proposed by the Learned Society of Copenhagen³. His case was also helped at the Royal Society by the support of his influential father.

During the War of the Spanish Succession, Charles's grandfather Earl Stanhope had received the title of Mahon for the seizure of the port bearing this name on the Isle of Minorca. After the death of his father in 1786, Charles became the third Earl Stanhope.

From 1780 until his succession to the peerage in 1786, he was member of Parliament. Educated under the opposing influences of conservative Eaton and democratic Geneva, he took in politics the radical democratic side. Being "in a minority of one" in the House of Lords, he seceded from Parliament from 1795 to 1800.²¹²⁵

His Geneva preceptor was George-Louis Lesage (1724-1803). An adherent of obsolete mechanistic views of electricity, Lesage, nevertheless, was perhaps the first to put forward the idea of an electrostatic telegraph¹⁶ which was as progressive as impracticable.

point charge (in the numerical method below, we take into account the actual configuration of the system).

Stanhope's method may be characterised as: first equations, then physical ideas. Such an approach was used to advantage in the early 20th century in the creation of quantum mechanics and the theory of relativity.

Starting from

$$\tau(x) \approx \frac{1}{x^2}$$

and

$$\int_{OA}^{ON} \tau(x) dx = \int_{ON}^{OB} \tau(x) dx$$

Stanhope obtained

$$AN = OA \cdot AB / (AB + 2 \cdot OA) \quad (1)$$

by simple manipulations reducible to a compact form⁷.

Without presenting his experimental results, Stanhope merely declares that all of his experiments in finding the position of the neutral are in "excellent agreement" with his theory. He gives a table of results calculated from the above equation, and this table may well be taken for a table of experimental data. He mis-

leads the reader by writing that it was not easy to determine experimentally the position of *N* with an accuracy of a hundredth or even an eightieth of *AB*. The reason given by him for the difficulty is quite clear: in the vicinity of *N*, at both sides, the response of the electro-scope was very weak⁸.

Correct criticism

The declared accuracy was out of the question, as we shall show.

But many were dissuaded from criticising "one of the best mathematicians of Europe" –

a description found on the title-page of the French translation of Stanhope's book. Among the victims of an uncritical approach to his work was Thomas Young (1773-1829) who even attributed to Stanhope the discovery of the Coulomb law⁹ – the equation $\tau(x) \approx 1/x^2$ being apparently similar to the famous law published six years after the appearance of Stanhope's book.

Some contemporaries of Stanhope rightly criticised his work. For example, John Robinson (1739-1805) experimentally disproved Stanhope's results (unfortunately,

Solving Stanhope's problem

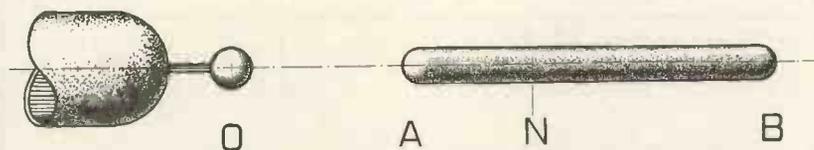
Stanhope gives the complete geometry of his experimental arrangement a metal ball, 4.5in. (0.114m) in diameter, with *O* as its centre, is joined by a stud 0.75in (0.019m) in dia, to a prime conductor in the form of a cylindrical envelope, 1ft (0.30m) in dia and 6ft (1.83m) long, terminated in hemispheres. Total length of the prime conductor-stud-ball system was 6ft 9in (2.06m). Rod *AB* is 3.75in. (0.095m) in dia and 3ft 4in (1.02m) long, and also terminated in hemispheres. The clear distance between the ball and the rod was varied between 4ft and 4in. For smaller distances, sparking occurred, from which it can be concluded that the output of Stanhope's electrostatic generator could be about 200kV¹¹.

"Stanhope's problem" has been solved at St Petersburg Technical University under the guidance of GV Podporkin using the charge simulation technique¹, a method implemented by the Triton program¹³. Podporkin took the earth into account, and the rod was assumed to be suspended 1.5m above a conducting plane. To solve the problem, the prime conductor, stud and rod were divided into 7, 3 and 13 parts, respectively, and all assumed to be cylinders; the sphere was taken as such.

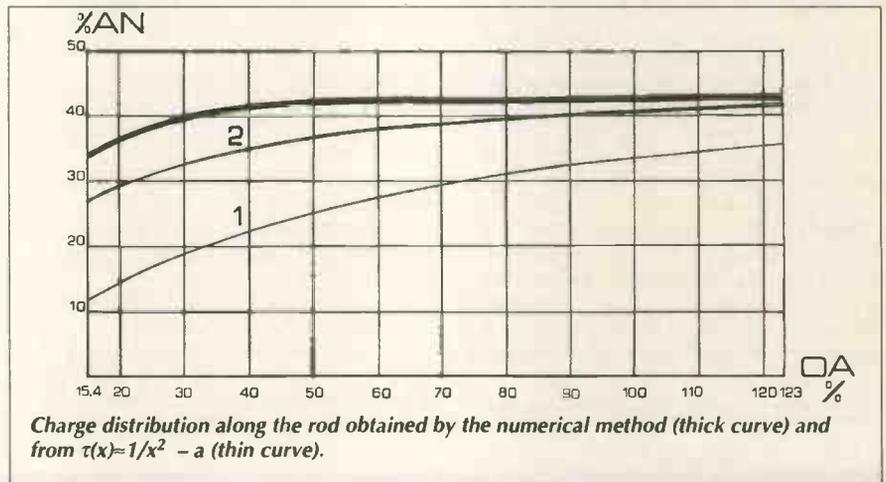
Distribution of the charge induced along the rod has been computed in a piecewise-constant approximation. After smoothing the bar graphs obtained for different ball-to-rod spacings, the charge distribution was found to be always of the form shown in the graph (thick curve).

With the clear distance between the ball and the rod varied from 10 to 120cm, which corresponds to *OA* varying from 15.7 to 125.7cm, or from 15.4 to 123% of *AB*, the position of the neutral, *AN*, has been found to vary from 35.3 to 44.5cm, or from 34.6 to 43.5% of *AB*, as shown in the graph (thick curve).

The computed results fit reasonably well into the equation which obtains if Stanhope's idea of the "inverse-square law" is reconciled with the electrical neutrality of the rod as a whole.



Stanhope's experimental arrangement. Point *N* was proposed as electrically neutral, confirming to Stanhope the inverse square law for distance and charge density. He claimed to verify his hypothesis with a crude electro-scope.



Charge distribution along the rod obtained by the numerical method (thick curve) and from $\tau(x) = 1/x^2 - a$ (thin curve).

$$AN = [OA \cdot (OA + AB)]^{0.5} - OA \quad (2)$$

We can explain this by letting the charge distribution along the rod be represented by a curve of the form given by $1/x^2$. The curve must cross the X axis at *N* so as to satisfy the neutrality condition of the rod. Thus, $\tau(x) = 1/x^2 - a$ for $OA \leq x \leq OB$. So "Stanhope's curve" must be displaced vertically by *a* (thin curve, Fig. 2). To find *a*, put down the neutrality condition, with equal magnitudes of the total charges on *AN* and *NB*,

$$\int_{OA}^{ON} \left(\frac{1}{X^2} - a \right) dx = - \int_{ON}^{OB} \left(\frac{1}{X^2} - a \right) dx$$

Integrating and then solving the equation for *a* yields

$$a = 1/[OA \cdot (OA + AB)]^{0.5}$$

By substituting *a* in $1/x^2 - a$ and equating this to zero, we find

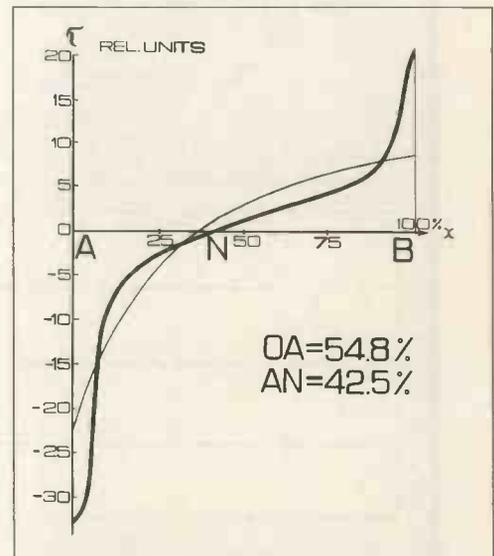
$$x = ON = [OA \cdot (OA + AB)]^{0.5}$$

$$\text{So } AN = ON - OA.$$

The lower graph shows the *AN* vs *OA*

dependencies calculated from Eqs. (1) and (2) (curves 1 and 2 respectively), in addition to the dependence shown by the above numerical method (thick curve).

Note that Eq (1) gives not the position of the electrical neutral but the position of the centre of masses of the rod *AB* with respect to *O*. That is how far Stanhope was led away by the analogy between static electricity and gravitation.



Null-point position obtained by the numerical method (thick curve), and from Eqs (1) and (2) respectively.

Robinson also did not publish his own experimental data)¹⁰.

Still, Stanhope's experiment was a landmark on the way to the quantification of electrical

Stanhope's inventions

Stanhope invented a method of securing buildings from fire (which, however, proved impracticable), the printing press and the microscopic lens which bear his name, a monochord for tuning musical instruments, a steam carriage and a steamship for conveying coal from Newcastle to London (which, however, said Lord Holland, "would have consumed its cargo before it could have reached its destination")¹⁷ and contrived mathematical machines. He projected a canal to link up his Devonshire estate with the Bristol channel, taking the levels himself, and experimented with new methods of raising and lowering the barges. He made a new kind of cement, more durable than ordinary mortar, and slates and tiles composed of tar, chalk and fine sand.

As for his mathematical machines, he claimed that they were capable of not only making calculations but also performing complicated logical operations and drawing faultless conclusions. Thus, Stanhope anticipated the quantification of electrical science and the development of computers.

science and so deserves to be analysed from the modern standpoint.

Scientific "misconduct"

By the modern standard, Stanhope's account of his experiment is a serious scientific misconduct. He thought he knew the answer and just "cut the corner", which is usually the basis for scientific fraud¹⁴. Nevertheless, Stanhope's work stimulated a deeper insight into and a quantitative approach to electrical phenomena so that, in a way, Stanhope may be regarded as a precursor of Coulomb.

Stanhope was never properly engaged in electrical research again – with the exception of a case study of electrocution by lightning presented by him to the Royal Society in 1787.

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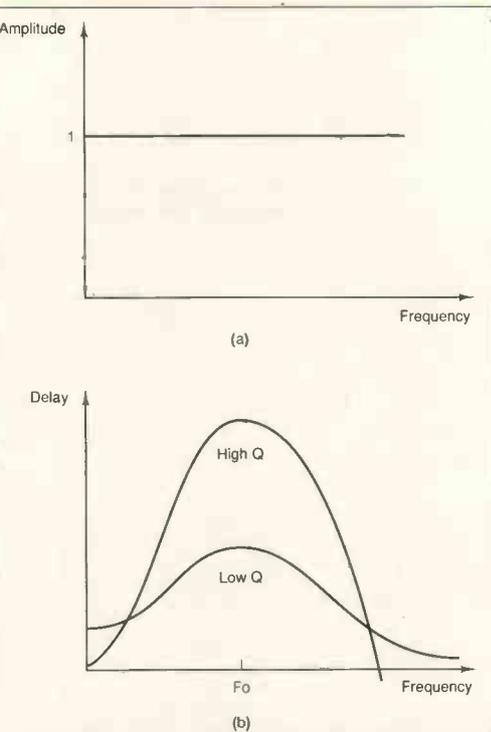
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All pass for universal SC filters

In a previous article, the MAX293 preconfigured SC device was used to design an elliptic, variable frequency low-pass filter. Now Bashir Al-Hashimi shows how to design all-pass filters using universal SC devices.



Upper. Amplitude response of an all-pass filter. Lower. Delay response of 2nd-order all-pass filter.

Designers wanting to define a custom filter type and specification need to use universal switched capacitor filters. Commercially available universal SC devices consist of one to four 2nd-order filter sections, with each section configured to yield low pass, high pass, band pass notch or all-pass filters. But though preconfigured SC filters are available for low pass, high pass, band pass and notch filters, they are not available to produce all-pass networks.

Universal SC devices contain one to four 2nd-order filter sections per chip. The 2nd-order filter is normally based on the state-variable configuration¹. In its basic form, this circuit consists of two integrators and a summing amplifier. By adding external resistors, this configuration can be used to generate low pass, high pass, band pass, notch or all-pass filtering functions.

Second-order filters are characterised by H_o , F_o and Q . H_o defines the filter gain while Q represents the filter quality factor. F_o defines the cut-off frequency of low pass and high pass filters, and may also represent the centre (resonant) frequency of band pass, notch or all-pass filters.

In universal SC filters these design parameters are set using a combination of a clock and external resistors. The different ways of connecting the resistors are called the filter operating modes. Usually, commercial universal SC devices have up to seven different operating modes which determine the complexity and/or flexibility of the final filter.

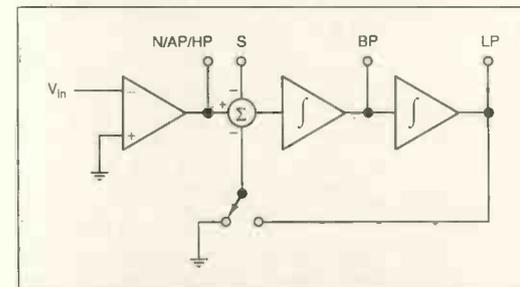
For example, mode 1 in the MAX265 universal device from Maxim, allows the simplest implementation of low pass and band pass filters using only two external resistors. But the limitation of this mode is that F_o and Q of the filter cannot be tuned. Low pass and band pass filters with tunable characteristics need mode 3, with the tuning ability achieved at the expense of introducing two more resistors.

Other modes allow different compromises of filter type, flexibility and performance to be achieved.

Design example

The design process using universal SC filters can be carried out in the following steps:

- Choose the appropriate mode.
- Define the 2nd-order filter design parameters (H_o , F_o , Q) – determined by the characteristics required of the filter being designed.
- Choose the appropriate (F_{clk}/F_o) ratio and calculate the values of external resistors required. Resistor values may be obtained from the tables or the design software supplied by the IC manufacturer.



State variable circuit consists of two integrators and a summing amplifier.

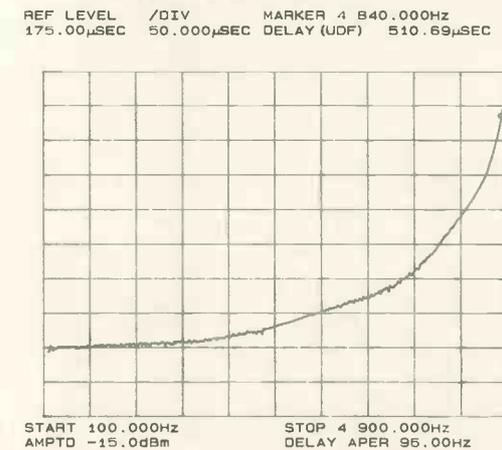


Fig. 1. Group delay response of a 5kHz low pass elliptic filter.

Delay equalisation

All-pass filters have the frequency characteristics shown in the graphs on the previous page. The theoretical amplitude response does not change with frequency while the delay response of 2nd-order all-pass filter has the shape of the lower graph. The filters are used to provide delay equalisation, and for this reason they are often called delay equalisers.

If a signal is to pass through a filter without distortion, the filter delay characteristics must be considered. To minimise signal distortion, the filter must exhibit a linear phase-frequency characteristic. The derivative of the phase function is a measure of the delay (or group delay) through the filter. A typical phase curve for a sharp cut-off filter is shown in Fig. A, with its group delay curve shown in Fig. B. Deviation from "linear phase" gives rise to the variation in group delay response, meaning that signals of different frequencies within the filter pass band will be delayed differently through the filter. The result is dispersion and adds to the ringing caused by Fourier truncation.

Classic "ringing" noticed on a square waveform is a combination of truncation of the Fourier series and non-linear phase as shown in Fig. C.

Minimum possible ringing is obtained by having a filter with constant group delay. This can be achieved by cascading all-pass networks with the appropriate characteristic to "build up" a constant delay from the filter group delay curve. Figure D shows the group delay graph of the filter to be equalised together with the delay responses of the two required all-pass filters denoted by D_1 and D_2 respectively. In this example, equalisation is required such that the total delay graph (filter plus group delay equaliser) is within $80\mu\text{s}$ over the frequency range of 10-5kHz. Adding the three delays graph (filter and the 2 group delay

equalisers) gives the total delay graph shown in Fig. D. Effect of the group delay equalisation on the ringing is shown in Fig. E.

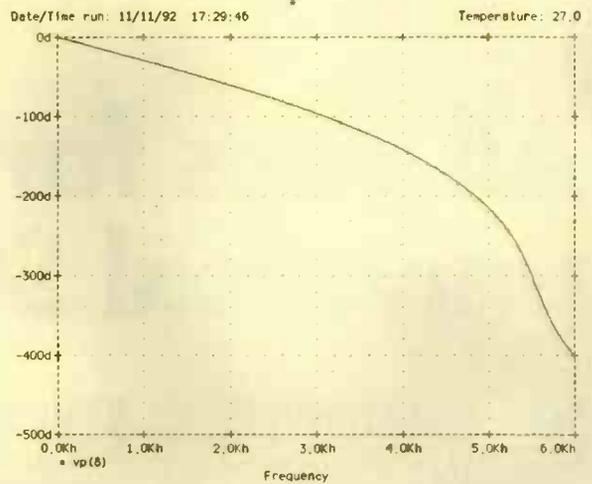


Fig. A. Typical phase curve for a sharp cut-off filter

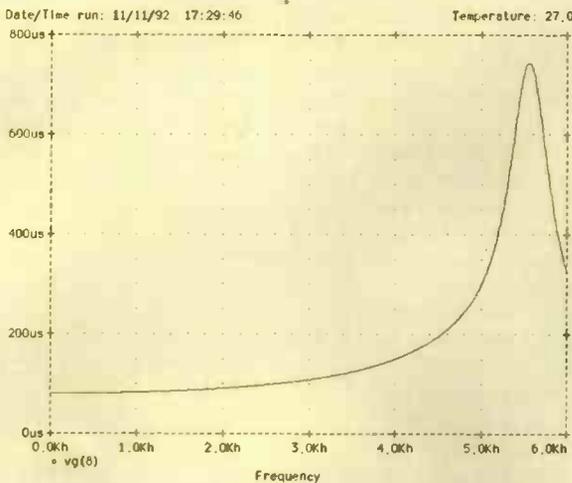


Fig. B. Group delay curve. Deviation from "linear phase" gives rise to the variation in group delay response. Signals of different frequencies within the filter pass band will be delayed differently through the filter.

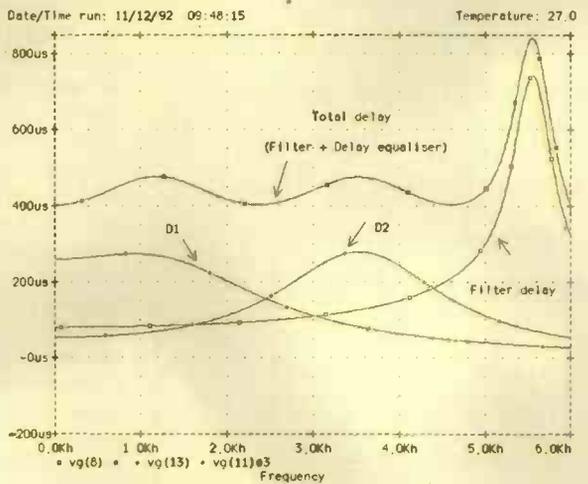


Fig. C. Classic "ringing" noticed on a square waveform is a combination of truncation of the Fourier series and non-linear phase.

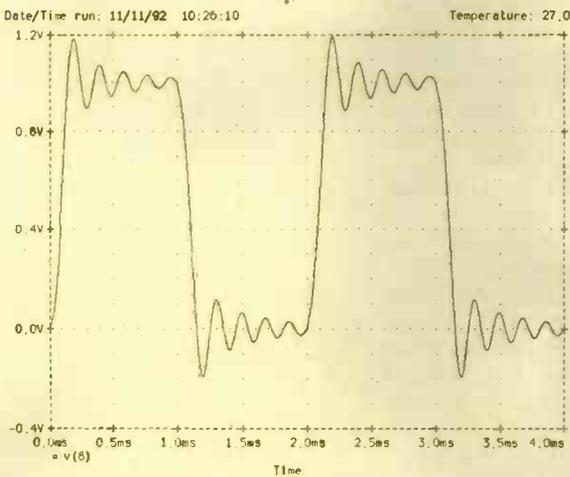


Fig. D. Group delay graph of the filter to be equalised together with the delay responses of the two required all-pass filters denoted by D_1 and D_2 respectively.

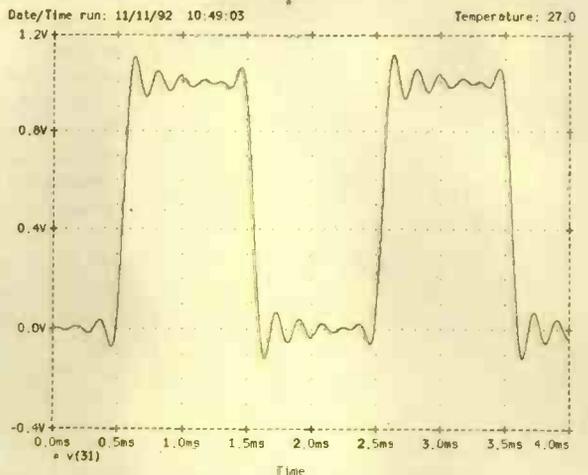
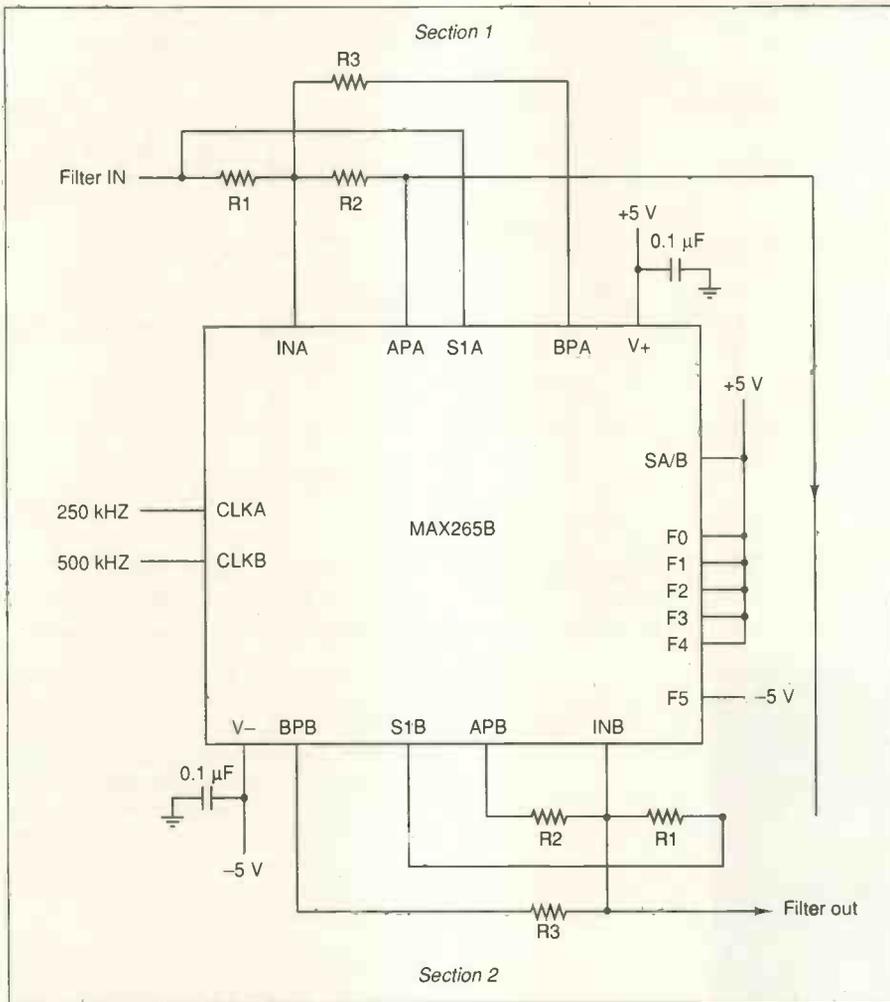


Fig. E. Effect of a group delay equalisation.

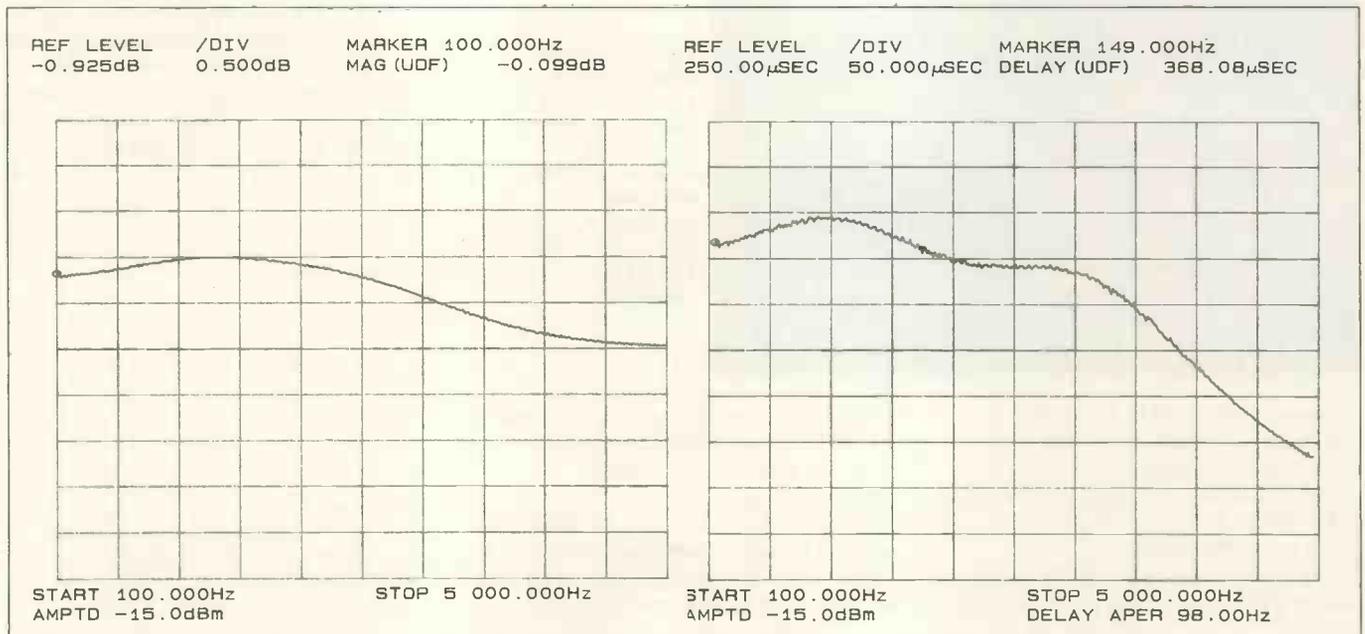


Since preconfigured SC filters are not available to produce all-pass networks, we will look at design of group delay equalisers using such networks.

First, assume that a system requires the group delay ripple of a 5kHz elliptic filter (considered in detail in "Better design with SC filters", EW + WW, May) to be flat within

Fig. 2. Resistors of 1% tolerance were used in practical implementation of the complete group delay equaliser.

Fig. 3a). Amplitude response of the group delay equaliser. It should be a straight line but has 1dB roll off. b) Delay response of the group delay equaliser.



50µs over the frequency range of 10Hz-4.6kHz. The group delay graph of the filter is shown in Fig. 1.

Using an optimisation program² to define the parameters of the all-pass filters required to give the target response, one solution indicates that two 2nd-order sections will satisfy the requirement.

Several commercially available universal SC devices can be used to implement the group delay equaliser.

One is the MAX265 from Maxim, a device containing two 2nd-order filter sections that can be configured to yield the required all-pass filters. Mode 4 of the MAX265 is the only mode available for implementing all-pass filters.

To design 2nd-order all-pass filters, parameters F_o and Q are required, in this case obtained from the optimisation routine mentioned earlier. F_o and Q parameters of each 2nd-order all-pass filter are:

- 1st-section $Q = 0.638, F_o = 1.70\text{kHz}$
- 2nd-section $Q = 1.259, F_o = 3.40\text{kHz}$

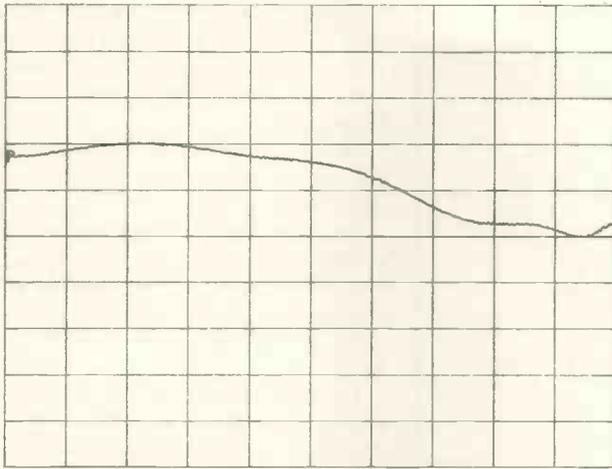
Although, each 2nd-order filter of the MAX265 has its own clock input, they must both share the same (F_{clk}/F_o) ratio, externally programmed by six pins-strapped inputs. The ratio varies approximately over the range of 100 to 200 in 64 steps, given in the data sheet.

So, to implement the design, the (F_{clk}/F_o) ratio of 149.23 is selected from the data sheet. This ratio is chosen so that the required clock signal of each all-pass filter section can be closely derived from the master clock signal (500kHz) used to set the pass band edge of the 5kHz filter considered in the previous article (EW + WW, May).

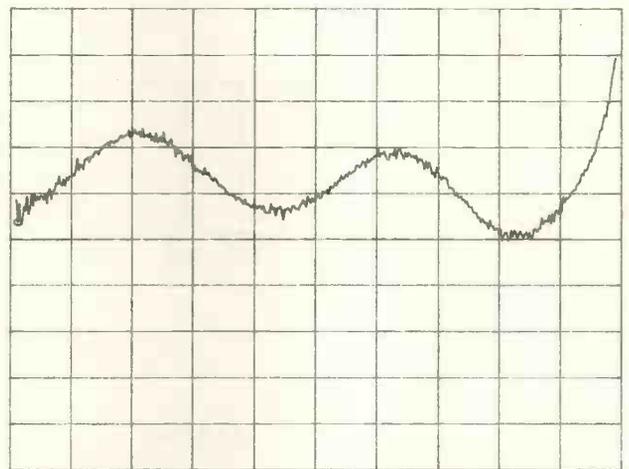
Having decided on the clock ratio, resistor values and the clock signal for each section are obtained using the Maxim filter design software. Given the F_o and Q of each all-pass filter, the program outputs:

REF LEVEL /DIV MARKER 136.750Hz
 -1.250dB 0.500dB MAG (UDF) -0.354dB

REF LEVEL /DIV MARKER 158.750Hz
 528.00µSEC 20.000µSEC DELAY (UDF) 535.50µSEC

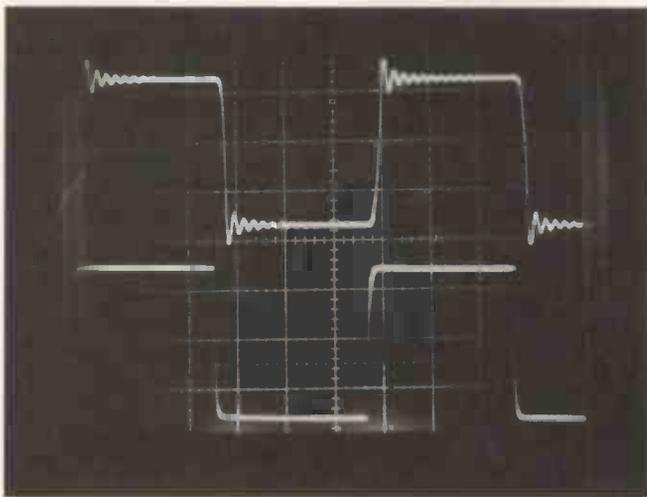


START 100.000Hz STOP 5 000.000Hz
 AMPTD -15.0dBm



START 100.000Hz STOP 4 800.000Hz
 AMPTD -15.0dBm DELAY APER 94.00Hz

Fig. 4a). Amplitude response of filter and group delay equaliser combined. b). Delay response of the filter and group delay equaliser combined.



Left Fig. 5. Top trace: 5kHz elliptic filter output waveform. Bottom trace: 500Hz input square

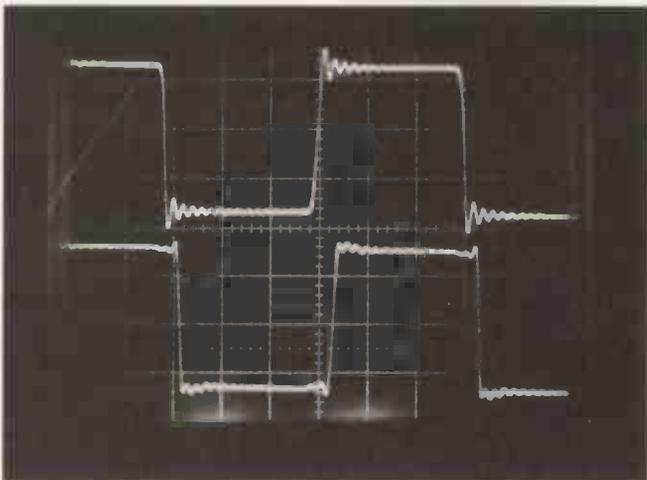


Fig. 6. Waveform after group delay equalisation. Top trace: 5kHz elliptic filter output waveform without any group delay equalisation. Bottom trace: With group delay equalisation.

1st-section $R_1 = 20k\Omega$, $R_2 = 20k\Omega$,
 $R_3 = 25.445k\Omega$, $(F_{clk}/F_o) = 149.225$,
 Clk freq = 253.7kHz

2nd-section $R_1 = 30.786k\Omega$, $R_2 = 30.786k\Omega$,
 $R_3 = 20k\Omega$, $(F_{clk}/F_o) = 149.225$,
 Clk freq = 507.38kHz

$N = 011111$ (code for the pin setting for ratio F_{clk}/F_o).

Actual clock signals used for each section are 250kHz and 500kHz respectively. The complete group delay equaliser circuit is shown in Fig. 2.

Resistors of 1% tolerance are used in the practical implementation. Figure 3 shows the amplitude and the delay response of the delay equaliser and Fig. 4 shows the amplitude and

delay response of the filter and the group delay equaliser combined. Equaliser amplitude (Fig. 3a) – theoretically a straight line – has about 1dB roll-off. Achieved group delay ripple is <50µs up to 4.6kHz (Fig. 4b). Variation in practical amplitude and group delay response from the theoretical targets are due to the tolerance in clock signals used to drive the all-pass sections and to the MAX265B itself which has 2% tolerance on F_o and Q .

The group delay equaliser has been designed theoretically to yield <40µs group delay ripple. Figure 5 shows a 500Hz square waveform after passing through the low pass elliptic filter without any group delay equalisation.

After group delay equalisation (Fig. 6) the overshoot has been reduced considerably and the square waveform has a better symmetry.

Choose preconfigured devices

Universal SC devices can be used to produce all-pass filters not available as preconfigured devices. But the design process is significantly more involved for universal SC devices rather than preconfigured devices, and this is true for all filter configurations. So, wherever possible, use preconfigured devices and only resort to universal devices when necessary. ■

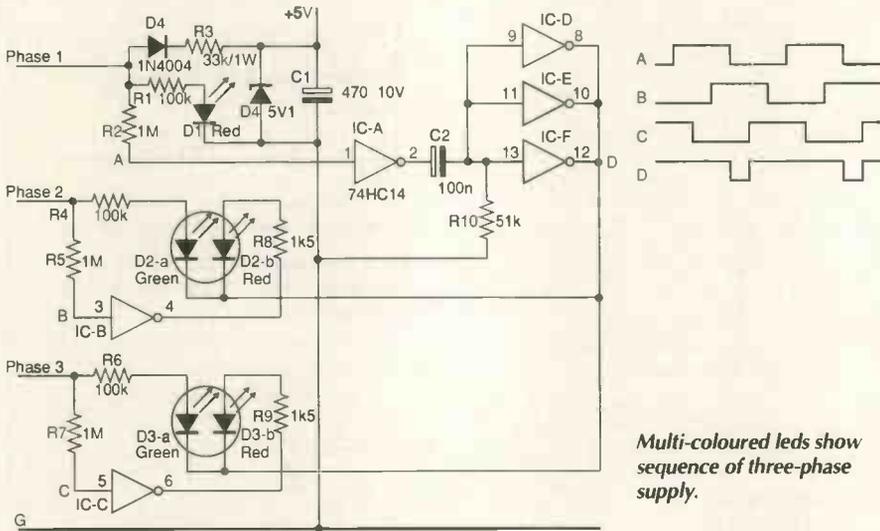
References

1. M E Valkenburg, Analogue Filter Design, Holt-Saunders International, 1982.
2. J.K. Fidler & C. Nightingale, Computer Aided Circuit Design, Nelson, 1978.

CIRCUIT IDEAS

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

Three-phase indicator



Multi-coloured leds show sequence of three-phase supply.

Phase sequence in a three-phase supply is shown by red and green leds.

Phase 1 drives a red led, is rectified to provide 5V for the gates and, after the 74HC14, emerges as 3ms negative-going pulses. Phase 2 drives a green led and is inverted to drive the red led, which is also controlled by the 3ms pulses from point D. Phase 3 is the same as phase 2.

If the phases are in order, phase 2 red led is on and phase 3 red led off so that, since the pairs of leds are in the same packages, phase 2 shows yellow and phase 3 green. If phases 2 and 3 are out of order, phase 2 is green and phase 3 yellow.

Yongping Xia
Torrance
California

DO YOU HAVE A £100 CIRCUIT? AS OF THE JULY ISSUE, EACH MONTH'S TOP CIRCUIT IDEA AUTHOR WILL RECEIVE £100. ALL OTHER PUBLISHED IDEAS WILL BE WORTH £25. WE ARE LOOKING FOR INGENUITY AND ORIGINALITY IN THE USE OF MODERN COMPONENTS

Microprocessor/analogue voltmeter

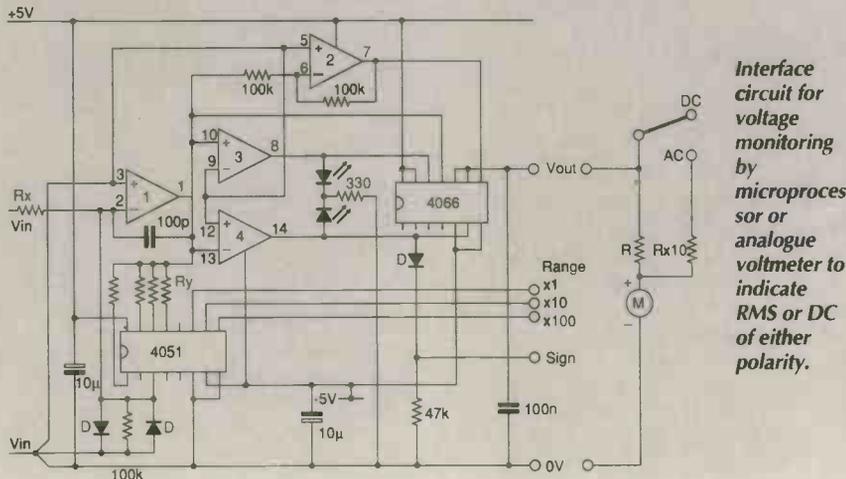
Originally intended as a microprocessor interface for DC, or average, peak and RMS AC voltage monitoring under software control, this will also serve as the core of an analogue voltmeter for DC and RMS measurement. Full resolution from the microprocessor's A-to-D converter is retained for either polarity, a differential type being unnecessary. The circuit provides for automatic polarity recognition and either automatic software or manual range setting. TL074 op-amps 1 and 2 provide inverted

and non-inverted inputs to two fet switches in the 4066, scaled by R_y/R_x . Op-amps 3 and 4 are comparators, which recognise polarity and drive the fet gates in such a way that the output capacitor charges or discharges unidirectionally. The three-coloured led across their outputs shows red or green for polarity and yellow for AC.

For an analogue indication, a 2.5V meter across the output capacitor will serve and manual range-resistor selection eliminates the 4051 multiplexer, which is software controlled when the circuit is used as a microprocessor interface. With R alone in series with the meter, DC or an average-value reading results; to make it read RMS, the 10Ω resistor across it increases the reading by 1.11.

On a practical note, the input limiting diodes must be shielded against light and high $R_{x,y}$ values require care with the layout.

H Maidment
Wilton
Salisbury
Wiltshire.



Interface circuit for voltage monitoring by microprocessor or analogue voltmeter to indicate RMS or DC of either polarity.

Measuring voltage in interference

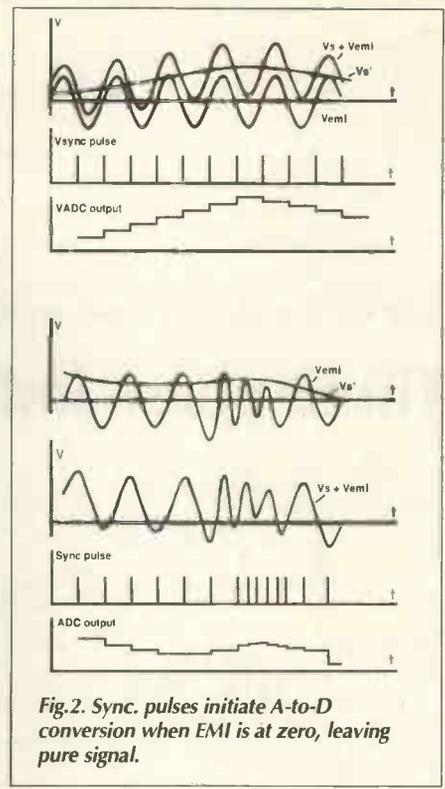
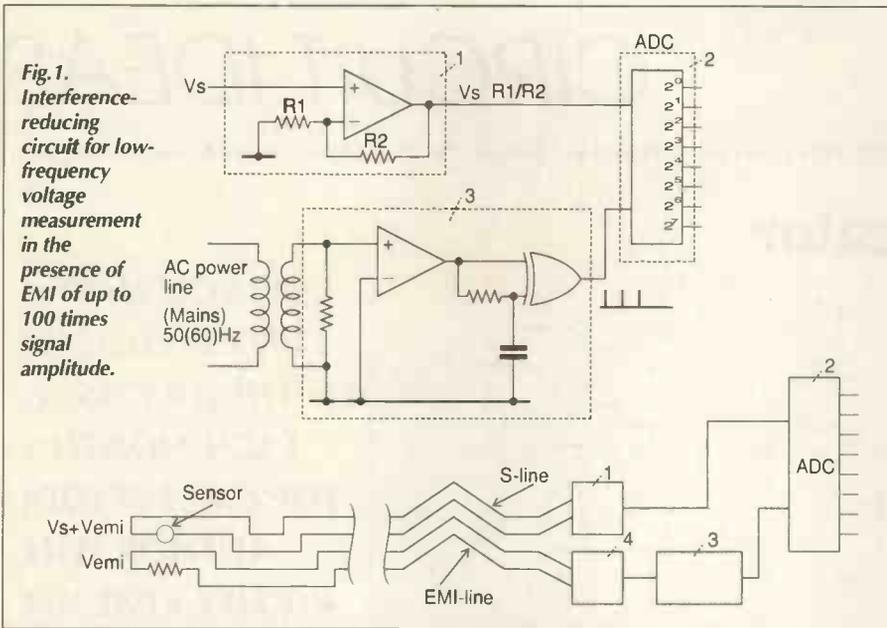


Fig. 2. Sync. pulses initiate A-to-D conversion when EMI is at zero, leaving pure signal.

Although integrating A-to-D converters can reduce the effects of mains-frequency interference when measuring small, low-frequency voltages, other types of interference, the periods of whose fundamental and harmonics are not multiples of A-to-D integration time, still cause trouble. This circuit alleviates this situation.

In essence, a sample of input signal is taken at the moment the interference is at zero, so that the remaining input is purely signal. The A-to-D holds the sample until the next EMI-zero sync. pulse one period of interference later.

Figure 1 is the basic idea when the only interference is at mains frequency. Signal plus noise is amplified and passed to the converter, which only produces a conversion when the mains-frequency sync. pulse appears from the logic circuit. Figure 2 shows the effect.

In the case of random-frequency interference, it is preferable to use a dummy sensor instead of the step-down transformer, reproducing the measuring sensor's resistance, capacitance and inductance, connected by the

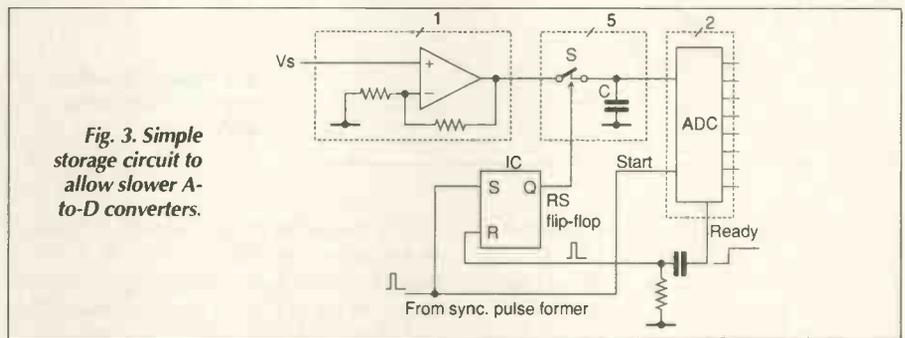


Fig. 3. Simple storage circuit to allow slower A-to-D converters.

same length of line and taken to a similar amplifier. The rest of the circuit is the same as that of Fig. 1. If the interference is at a relatively high frequency, phase shift between the sensor and dummy sensor inputs should be negligible. A TDC11007 converter, with a conversion time of 0.15µs, and an AD509 amplifier worked well in the prototype.

To use a slower converter, the simple capacitive storage circuit in Fig. 3 is feasible. Sync. pulses open the switch and begin the conver-

sion, the converter's ready pulse closing the switch to allow a further reading at the next sync. The capacitor discharge time must not exceed one period.

At low signal frequencies, this circuit is valid when EMI is up to 100 times greater than signal voltage.

N T Lavrentiev
Moscow
Russia

555 has high mark:space

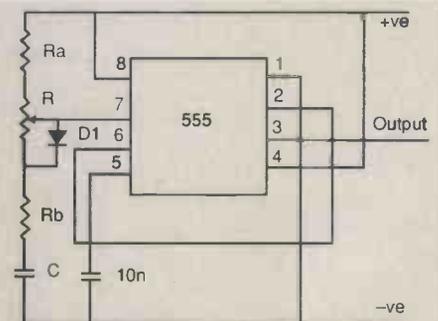
An astable flip-flop using a 555 timer in the usual circuit will not operate with a duty cycle of less than 50%. Diode D₁ shown in the diagram overcomes the problem; it allows a duty cycle from very near zero to virtually 100% while keeping a reasonably constant frequency.

$$T_{(on)} = R_1 C \left(\ln \frac{3-k}{1-3k} - \ln \frac{3-k}{2-3k} \right)$$

and $T_{(off)} = 0.695 R_2 C$, where
 $R_1 = R_a +$ resistance of upper leg of pot.,
 $R_2 = R_b +$ resistance of lower leg of pot.,
 $k =$ diode forward drop/supply voltage.

If the diode drop is very small compared with the supply voltage, k can be neglected, giving $T_{(on)} = 0.695 R_2 C$, duty cycle $R_2 / (R_2 + R_1)$ and frequency $= 1.44 / C(R_1 + R_2)$.

Cyril WW Palihawadana
Sana'a
Republic of Yemen



Adding the diode to a 555 multivibrator circuit allows a much larger range of duty cycles - from virtually zero to 100% - with a fairly constant frequency.

Coarse/fine D-to-A audio attenuator

Two 8-bit digital-to-analogue converters are combined in this circuit to obtain an effective resolution of 14 bits, offering coarse and fine control.

Ignoring $R_{3,4}$ in the diagram, $R_{1,2}$ will combine the output currents of the AD7528 dual A-to-D converter. Resistors $R_{1,2}$ attenuate the input by the required amount; for example, if $R_1=63R_2$, the reference voltage of DAC B is 1/64 that of DAC A and the range of DAC B is 4 least-significant bits of DAC A. A V_{in} of 2.56V gives a DAC B reference voltage of 40mV and an LSB of 156µV.

Problems exist, however, with the basic arrangement. The output would not be monotonic, since the ratio of $R_{1,2}$ would need to be perfect and the input offset voltage of the output amplifier would contribute a code-dependent noise gain term. Additionally, there is the loading of DAC B input R across R_2 , although this can be eased by making the resistance a low value. Then again, the temperature coefficients of $R_{1,2}$ will not match that of the internal ladder.

All this is avoided by the inclusion of R_3 in series with the internal feedback resistor. Making this value $R_{1,2}^*$ means that DAC B reference voltage is only a function of the ratio $R_1:R_2$. Resistor R_3 should also have a temp.comp. similar to that of $R_{1,2}$. Resistor R_4 compensates for R_3 .

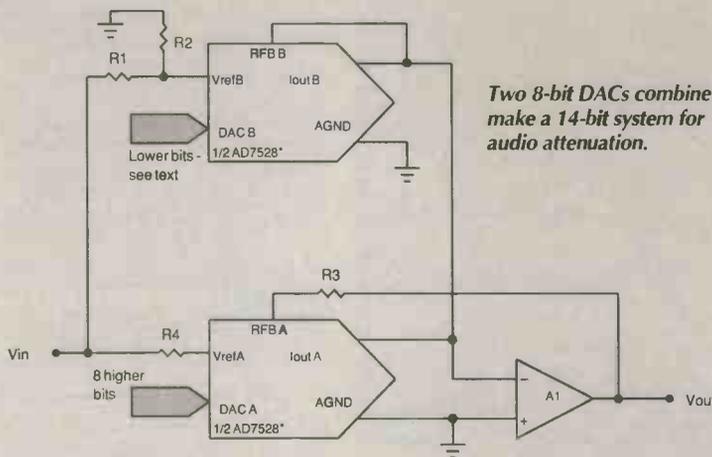
Output voltage is now

$$V_{out} = -D_A V_{IN} - D_B V_{IN} [R_2 / (R_1 + R_2)],$$

$D_{A,B}$ being fractional representations of the input code N in decimal; that is, $D_A = N_A / 256$.

*Brokaw, Paul. Input resistor stabilises MDAC's gain, EDN, Jan. 7, 1981, p.210.

John Wynne
Analog Devices
Limerick Ireland



Two 8-bit DACs combine to make a 14-bit system for audio attenuation.

Current source has 170V voltage compliance

Using a three-terminal adjustable voltage regulator chip, the TL783C, this circuit arrangement will drive switchable currents from 25mA to 100mA into a load varying between a few ohms and up to 1.7kΩ. TL783C maintains a nominal and stable 1.25V between OUT and ADJ pins, which is used to drive current through resistors

switched to give 25,50,75 and 100mA output to the load. Current into the ADJ pin is a very stable 60µA and, since this is referred to the output, it sets a minimum output current of 60µA and therefore a maximum load resistance. The rest of the circuit is designed to manage this.

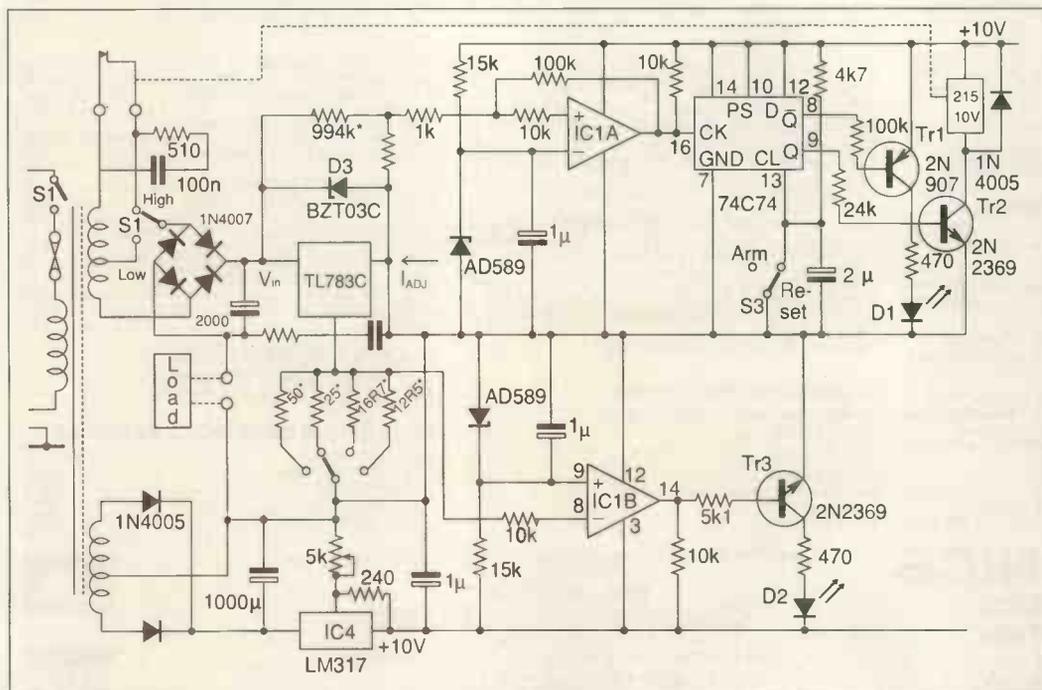
In the low position of S2, maximum

input/output voltage and dissipation are no problem, but an increasing load resistance reduces $V_{in}-V_{out}$ and $V_{out}-V_{adj}$. Comparator IC1b turns on Tr_3 and the led illuminates to show out-of-range at about a 1kΩ load resistance at 100mA.

With S2 at high, a load resistance of less than 700Ω or a short circuit would destroy the circuit, although the load can go up to 1.7kΩ. In this situation, the 3W zener across the TL783C limits $V_{in}-V_{adj}$, the schmitt IC1a detecting the limiting voltage and clocking the flip-flop. Both Tr_2 and Tr_3 turn on, the relay disconnects the input and led D_1 lights. A reset to the flip-flop reinstates the circuit when order is restored.

Voltage compliance in the low mode is 110V; taking into account both modes, 170V.

JJ Meyer
ITODYS
Paris.



Voltage regulator chip used in a 25-100mA current source with voltage compliance up to 170V.

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COMPARATIVELY BETTER THAN THE OP-AMP?

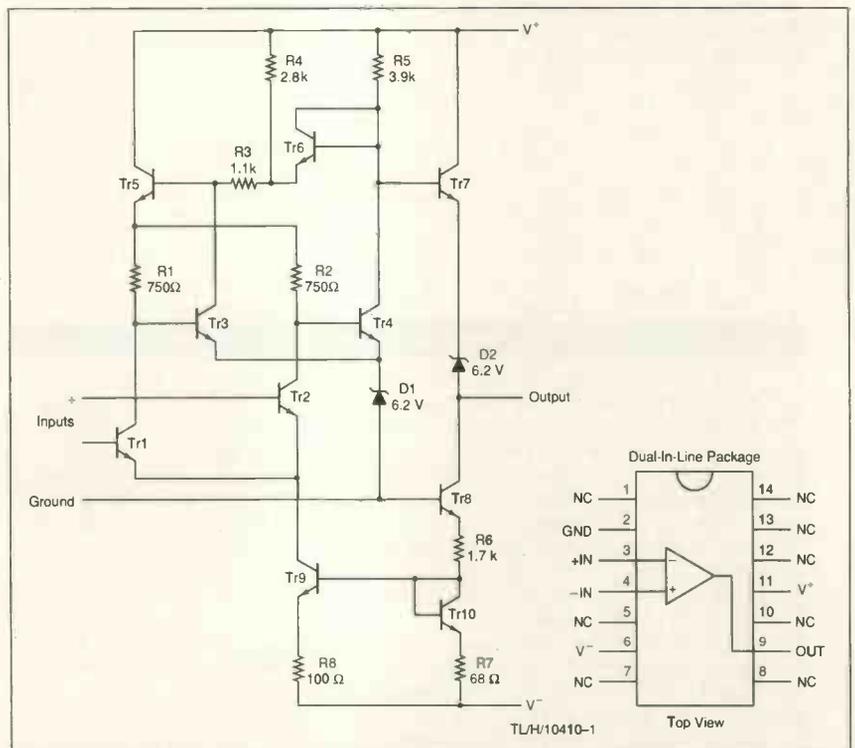
Comparators have much in common with op amps. But Ian Hickman shows why the comparator is the better choice in some applications.

Most ICs can be classified simply as either analogue or digital. But some – such as A-to-D/D-to-A converters – straddle the divide between the two. One of the earliest chips to fall into this category is the comparator, a device whose sole function is to signal as quickly and accurately as possible whether an analogue input voltage is above or below an arbitrary threshold. As such it may be regarded as a 1-bit A-to-D, and indeed at least one is integrated inside a conventional A-to-D converter IC, and many in the case of a flash converter. The analogue input voltage is applied to one input pin and a reference voltage to the other. Output is high if the voltage at the non-inverting input is higher than at the inverting input, and vice versa, just like an op amp.

But unlike an op amp, a comparator is designed to operate for long periods with the inputs “toggled”; ie at a difference of several or many volts – a condition which can cause unpredictable long-term changes in input offset voltage in some op amps. Furthermore, comparators are designed to respond rapidly when the input crosses the threshold. An op amp can take a long time to recover from the toggled condition, as various internal stages are cut off or saturated.

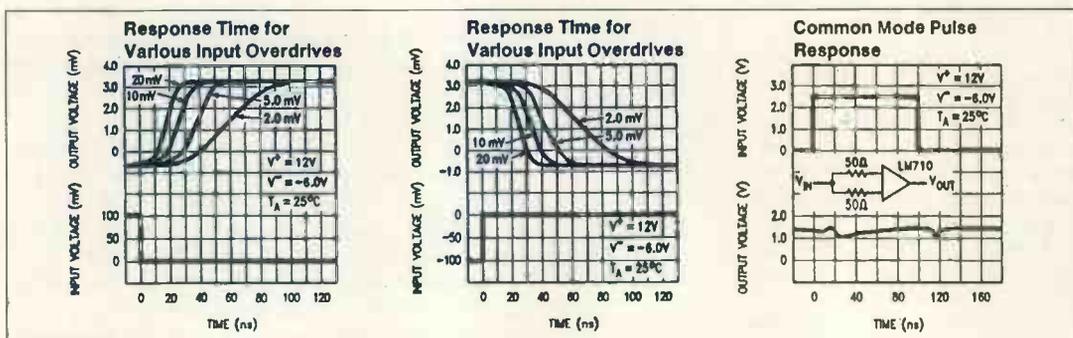
Design precautions

Useful as they are, precautions are necessary when designing comparators into a circuit. A commonly encountered problem is oscillation when the two inputs dwell at, or very near, the same voltage. Oscillation can occur when the input passes through the reference voltage with a slow-moving input such as a ramp waveform



(Fig. 2a). The reason is not difficult to see. A comparator has a gain similar to the open loop gain of an op amp. But whereas the latter is operated with a gain defining network, the comparator is actually used open

Fig. 1. 710 comparator.



Performance on positive- and negative-going inputs (at the inverting input pin) for various degrees of overdrive.

Like an op amp, a comparator exhibits a high CMRR, common mode rejection ratio. (National Semiconductor Corporation)

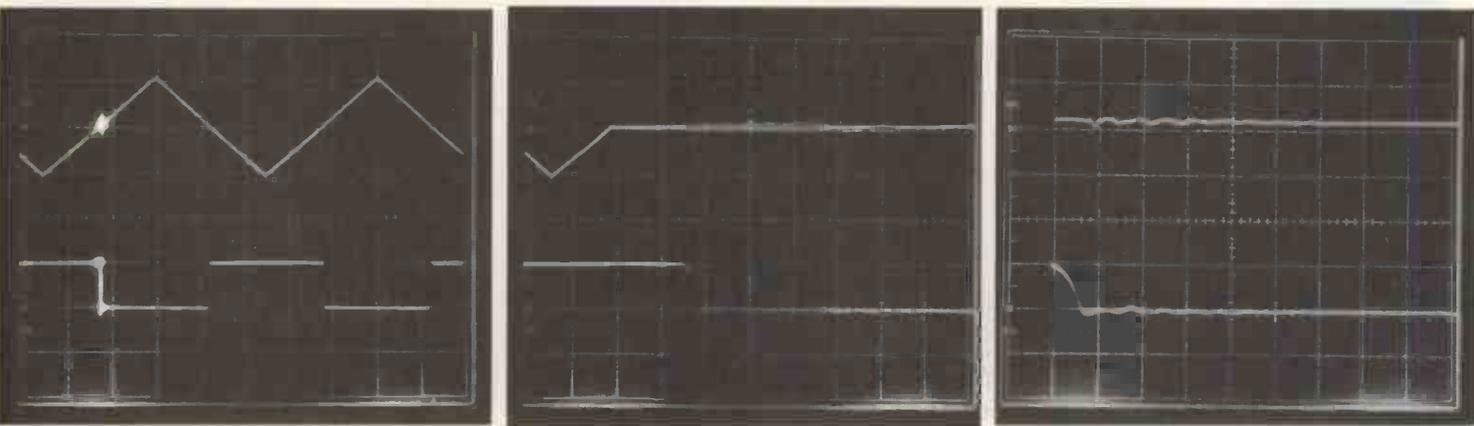
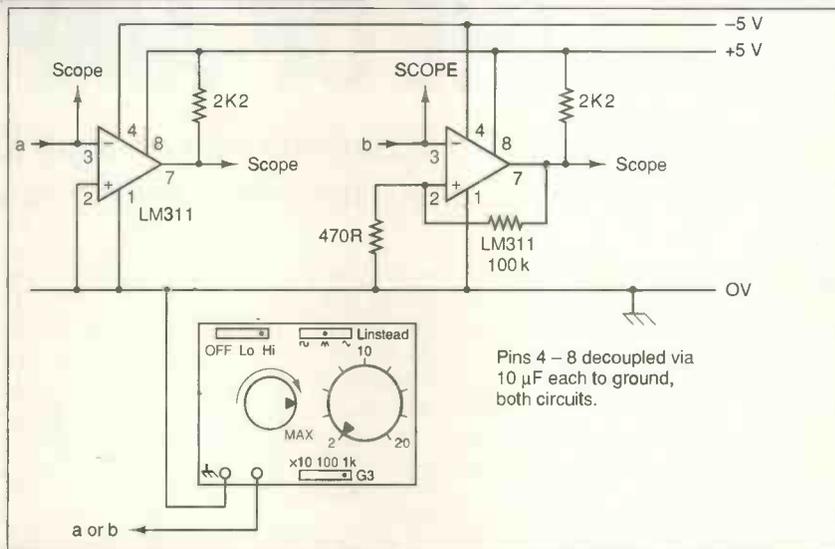


Fig. 2a). (Left). Positive and negative ramp-voltage input (upper trace, 0.2V/div) applied to an LM311 comparator, and resultant output (lower trace, 5V/div). "A" timebase (100µs/div) intensified by B (1µs/div). (Right). Same traces on MIX timebase (A switching to B at start of intensified period). In addition to multiple transitions at the output, hash kickback is visible on the input. b) (Far right). As a), but with hysteresis applied to the comparator (A timebase, 0.1µs/div), showing single clean transition. c) Circuits relating to a) and b).



Comparator choice

One of the earliest IC comparators was the SN72710 or µA710, still available under type numbers such as LM710. By comparison with many more modern devices this is still one of the fastest (Fig. 1) and it has a low impedance output designed to interface directly with TTL. With the input switching from 100mV one side of the reference to 5mV the other side (ie 5mV "overdrive"), the response time to the nominal TTL threshold of 1.4V is typically 40ns for both positive- and negative-going inputs. Other types show distinctly different delays.

Like an op amp, a comparator exhibits high common mode rejection. But since it is usually operated with one input tied to ground or some other reference voltage, this is usually of little significance.

Despite its speed, the 710 does have some disadvantages; it has no facility for adjusting the input offset to zero (found on many more modern types) and needs +12V and -6V supply rails. The 711 contains two comparators each generally similar to the 710, with outputs internally ORed and brought out to a single pin.

Typical of a later generation of comparators is the LM111/211/311 family, with input bias currents hundreds of times lower than the 710 and with much wider ranges of input, differential input and supply rail voltages - up to 30V or more. This family can be used with the usual ±15V supply rails found in much analogue circuitry. But since the input common mode range includes the negative rail, it can also be used with a single +5V supply. The family also offers an offset null facility and an NPN open collector output structure which conveniently interfaces with the internal input pull-up of TTL. As a result, many comparator outputs can be "wire ORed" - or should that be "wire NORed"?

Nothing comes for free of course, and the price in this case is a response time of 170ns for negative-going inputs and nearer 200ns for positive (with the output loaded by a 500Ω resistor to +5V).

Two further tricks are possible: if the open collector output is connected directly to the positive supply rail and the output taken between ground pin (the emitter of the NPN output transistor) and the negative supply rail, then, in a dual supply circuit, something approaching a rail-to-rail output swing is obtained. Note that using the output stage in this manner as an emitter follower instead of as an inverter reverses the function of the two input pins, the invert pin becoming non-invert and vice versa. A 50V, 50mA output transistor rating enables the device to drive lamps, solenoids or relays directly while, one of the two input offset adjust pins acts as a strobe input. Pulling it low (for details, see the data sheet) disables the device.

loop. Stray capacitance from output back to the non-inverting input - including that internal to the device - can cause oscillation by providing AC-coupled positive feedback. Another cause can be disturbances on the supply rail(s) due to current spiking as the output stage switches.

The usual solution is to apply a touch of deliberate DC-coupled positive feedback, the resulting hysteresis ensuring a clean single transition at the output (Fig. 2b).

Some comparators have a small degree of hysteresis built-in: 500µV for the Analog Devices AD790 which also incorporates a low impedance output stage especially designed to avoid supply-rail-disturbing current spikes during output transitions - a feature it shares with a number of other modern comparators. Disadvantage to incorporating hysteresis is that the switching level on the input waveform differs slightly for positive- and negative-going inputs and in some applications this may not be acceptable. Until a few years ago there was no simple way round the problem other than the clumsy expedient of using two separate comparators, one to detect the positive-going threshold and another for the negative. The faster the comparator - and ECL comparators are very fast - the more likely becomes oscillation on slow-moving inputs. As an example of the speeds involved, the VC7697 dual ultra-fast ECL comparator (from VTC Inc, pin compatible versions from other makers) has a typical input to output propagation delay of just 1.4ns, permitting signal processing applications at up to 600MHz. Latch-enable inputs enable each half of the device to work in either track-hold or sample-hold mode.

Fast comparators make a ground plane construction mandatory and even then, special precautions to avoid oscillation may be necessary, up to the ultimate step of breaking the loop. Loop breaking is how the Elantec

EL2019 clocked comparator works. A master/slave flip-flop is interposed between the internal comparator output and the output pin of the device. State of the inputs is held in the master stage but only transferred into the slave section on the rising edge of a clock pulse. Any disturbance to the input from a transition at the output

subsides before the next clock pulse, so the loop is effectively broken and a clean, single transition – indicating that the input has just passed the threshold – is assured (Fig. 3).

The time when the input passes the threshold is of course quantised to the nearest clock pulse. But a high speed clock and a slow slewing input makes this of little significance.

Without hysteresis, the positive and negative thresholds are exactly equal and the relative polarity of the input is indicated with perfect accuracy for input changes in either direction – subject to the finite gain of the comparator section and the time quantisation. The Elantec EL2018 is generally similar but has a transparent latch (for use in track/hold applications) instead of a master/slave flip flop. Both devices have tri-state outputs, controlled by chip select inputs. Voltage gain is 103dB and the 4V output swing corresponds to less than 30µVrms uncertainty at the input, more than adequate for a 16-bit A-to-D converter.

Comparators in use

Comparators have hundreds of uses in addition to their application in A-to-Ds. One is in the very common requirement to indicate whether a voltage rail is within its specified limits or not. The basic arrangement is the “window comparator” (Fig. 4a) where one comparator compares the input with the lower threshold, and another with the upper. If either output goes low, the rail voltage is out of limits – signalled to the system controller and/or indicated by a lamp or led. By arranging that the

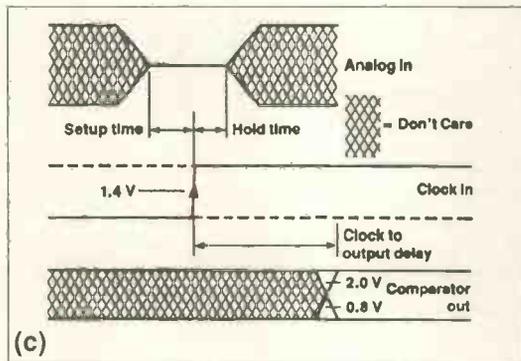
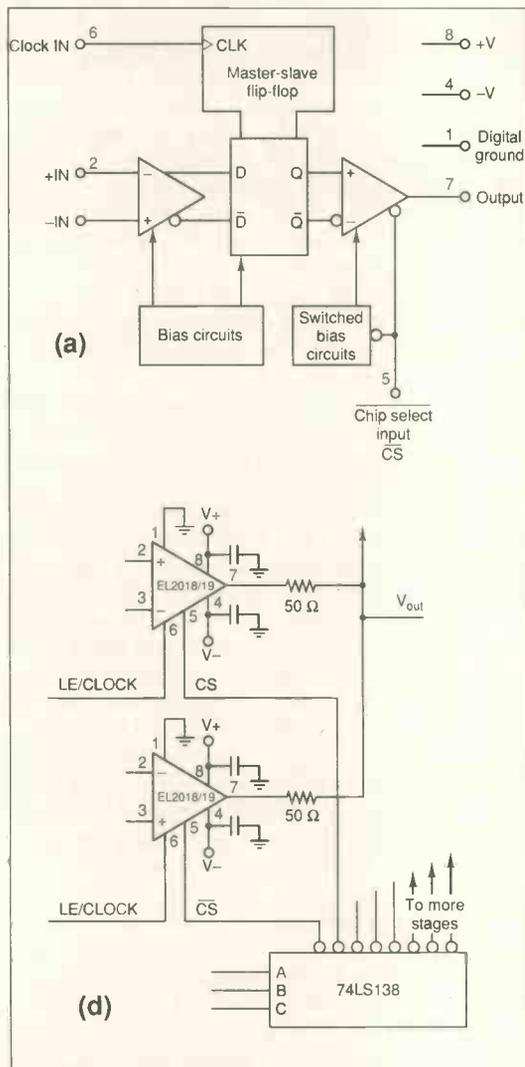
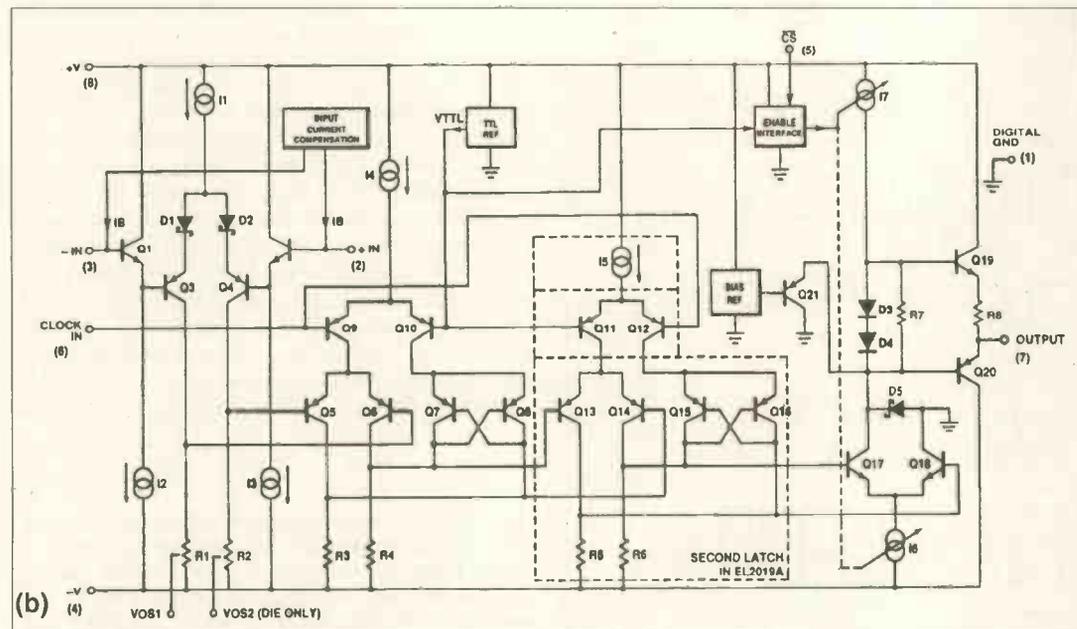


Fig. 3a) Elantec EL2019 fast, high voltage comparator with master slave flip-flop offers typical set-up and hold times of 12ns and -3ns respectively. b) The three state TTL compatible output stage draws a constant supply current, preventing supply line glitches. c) This feature, plus the fact that the output can change state only following a positive clock edge, prevents the possibility of oscillation. d) Using the chip select bar three-state feature to monitor several inputs.(Elantec Inc)



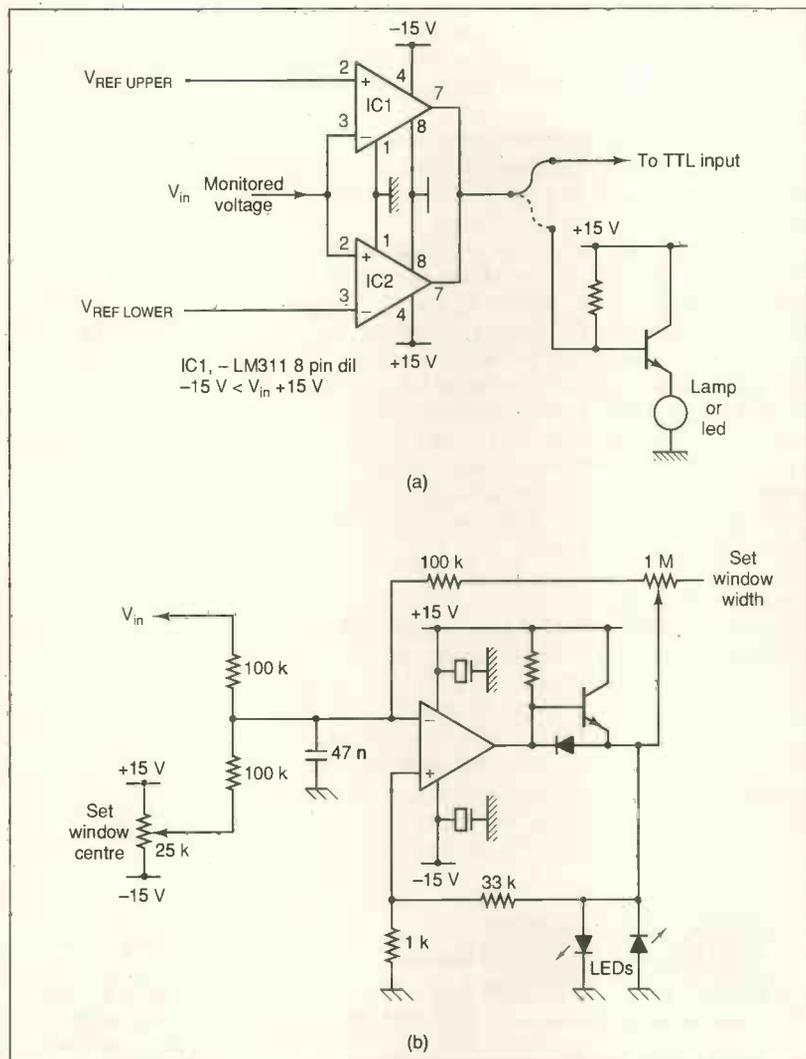


Fig. 4a). Basic window comparator, with logic output (low indicates out of limits) or lamp indication (lamp out indicates out of limits or lamp fault). Dedicated window comparator chips are also available.

b) Window comparator using a single comparator: either led out indicates voltage out-of-limits (or possibly led failure).

lamp is normally lit, then the lamp-out condition indicates trouble, either voltage out of limits or lamp failure.

A dual/quad comparator such as the LM393/LM339 can provide one or two such circuits in a single package. With a little ingenuity, a quad comparator can furnish four window comparators (Fig. 4b) and a single comparator indicates in- or out-of-window by oscillating or not. In-window, both leds are lit, one or other dimming slightly as the corresponding limit is approached. When the limit is reached the cessation of oscillation is abrupt, the appropriate led suddenly extinguishing and indicating whether the voltage is too high or too low (or possibly lamp failure).

Another application uses a comparator in a circuit for producing symmetrical square waves from an asymmetrical input (Fig. 5). A combined op-amp/comparator such as the LM392 is convenient-but will limit operation to a modest frequency range, since this handy eight pin device does not have offset null facilities for either the op amp or the comparator section. Using a separate op amp and comparator with offset adjust facilities enables the operational frequency range to be considerably extended: upwards - where the peak to peak output from the integrator becomes small; and downwards, by using higher values for R_2 and R_3 , given a fet or cmos input op amp. Note that neither edge of the equal mark/space ratio output square wave will coincide with either edge of the asymmetrical input square wave. The circuit operates on the assumption that the comparator output swing has equal peak positive and negative values. For equal rail voltages this is the case, to a close approximation.

Waveform generators

Comparators are also useful items in waveform generators of all sorts (Fig. 6). Waveform generator ICs such as the 8038 are very convenient for fixed frequency operation, but the circuit shown is much more flexible where a wide frequency range is required. There is also the additional advantage that the frequency is directly proportional to the value of the frequency setting potentiometer, rather than inversely proportional.

With the aid of a ten-turn digit dial, the 10K frequency setting resistor R provides a direct digital readout of frequency. 10% or even 20% tolerance range capacitors C can be used, since the top frequency on each range can be individually set by the appropriate 4K7 preset pot. Given equal supply rail voltages, the triangle and square wave outputs are both DC free, ie symmetrical about earth, while the output frequency is independent of the actual value of the (equal) supply rail volts. A comparator is used for IC_2 on account of its rapid response - the slow recovery of an op-amp from the toggled state would spoil the linearity of the frequency read-out on the top frequency range, highlighting the

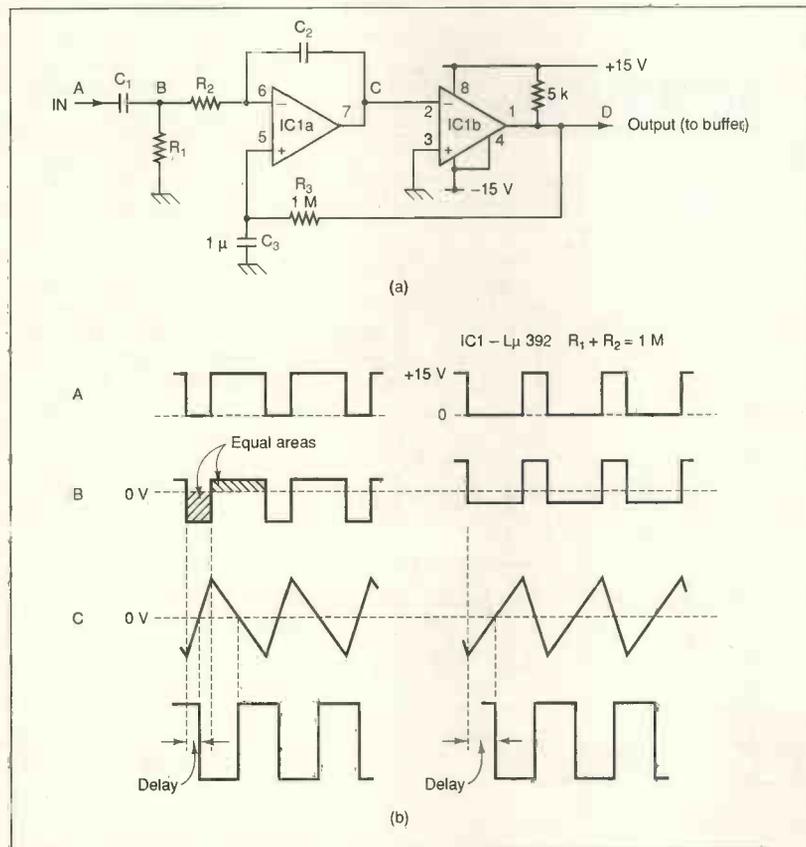


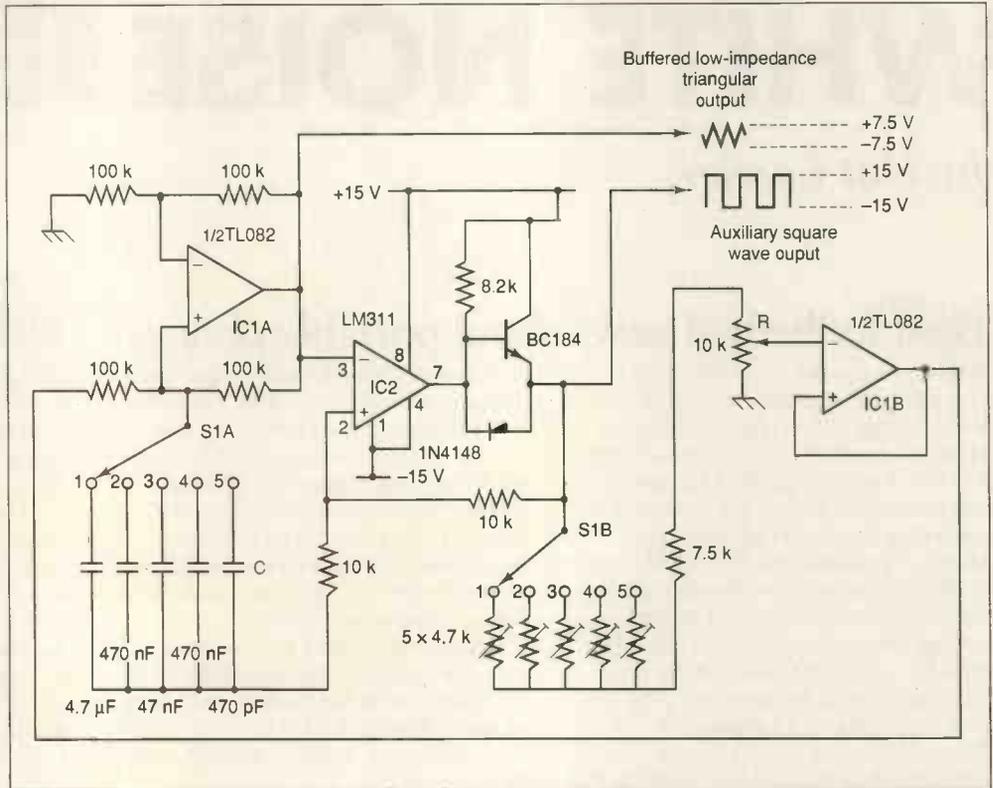
Fig. 5a). Circuit for converting asymmetrical square waves into symmetrical ones. C_1R_1 should be large enough to pass the lowest input frequency F_{min} with minimal sag. R_2C_2 should be such that at F_{min} IC_{1a} output almost reaches the rails. R_3C_3 should be much greater than $1/F_{min}$. b) Waveforms associated with the above circuit.

principle of "horses for courses". An op-amp can be used as a comparator (and vice versa), but in general the results are not optimum.

High voltage clamping

An interesting exception to this rule is the Comlinear CLC501 high-speed current-feedback op amp, which features a unique capability for output voltage clamping. The device has the usual eight pin dil pinout except that pins 5 and 8 are clamp pins V_{low} and V_{high} . These internally buffered pins may be set, for example, to limit the output swing to the standard TTL levels. Clamping avoids saturation, resulting in a very fast recovery time from overload of just 1ns. ■

Fig. 6. Circuit of a simple square- and triangle-wave function generator with digital readout of frequency from ten-turn digit dial associated with potentiometer R.

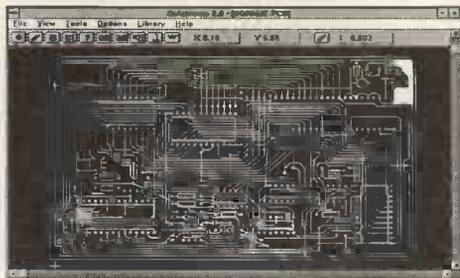


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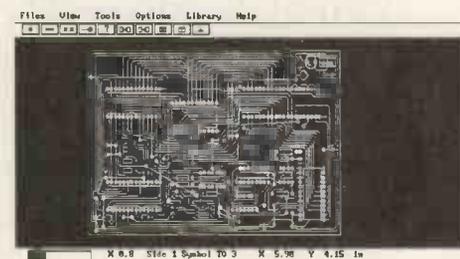
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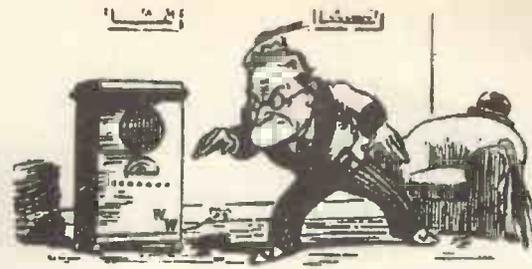
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hearing aids can be designed to incorporate immunity to high localised RF fields. But who's going to pay for retrofitting existing hearing aids. Is it the fault of the aids? After all they are only supposed to pick up acoustic sound waves, not to act as radio receivers (other than for LF induction loops). But then, when they were designed, no-one foresaw the hazard of high level RF fields impinging upon them.

The situation is similar to the amateur-radio-transmitter-interferes-with-TV syndrome, except that a TV is supposed to receive radio signals, so it could reasonably be expected to be designed to receive only the required ones.

Pity that hearing-aid wearers won't be able to hear all the inevitable arguments above the sound of the roaring in their ears.

Video phone highlights lack of vision

Congrats are due to the people at Marconi, whose video phone system is being adopted around the world, boosting its chances of becoming a *de facto* standard. Such success scored by a UK electronics company is all the more creditable in the light of the absence of

UK government support on anything like the scale common in other European countries. UK governments seem to see any financial encouragement for industry as automatically falling under the heading of propping up lame ducks. Other European countries - and even more so Japan - see investment as a strategic necessity for the

future. In the US, the government can not give money away to its electronics industry. President Clinton's plan for a \$17 billion four-year technology support programme for the industry is coming up against almost unanimous disapproval. Industrialists don't want it, being under no illusions as to where the money would come from, namely taxes that reduce the strings-free private capital available to their companies in the first place. The money is supposedly there anyway, from the "peace dividend". Better, they say, that the US government use it to reduce its huge annual deficit. Nice to have the choice though isn't it.

Thought please

Computer games, needing little more than hair-trigger rapid reflexes, might be more active than just watching tv, but surely there are more challenging pursuits on which tomorrow's adults (including engineers) can cut their intellectual teeth? What we want is not a race of instant button pushers, but people with a reasoned, questioning outlooks on life.

One engineer I knew was given the job of measuring the performance of a batch of passive LC filters. He was using a LO/SLMS set-up, a level oscillator plus selective measuring set. One of the parameters he was measuring was insertion

loss at top and bottom of the pass-band, and he solemnly recorded the results for each filter in the batch. In many cases, the recorded insertion loss was -0.1 or even -0.2dB! But that was what the instruments said. Needless to say, the problem was wrong terminations, as so often the case with filters.

A more inquisitive mind would have asked where the "extra" power was coming from - an attitude more likely to be fostered by the constructive pastimes of yesteryear such as *Meccano*, railway or aero modelling than by the likes of *Super Nintendo*.

Dumbo decibel

A well-publicised pronouncement from one of the business organisations recently flagged up excessive time spent on meetings as one of the contributory causes of our industrial malaise. Certainly the main task for a middle manager when things aren't going too well is often to be seen to be doing *something*, and holding a meeting can fit the bill. But another cause which might have been mentioned was the limited technical background of many managers.

I recall an antenna distribution system where the signal was to be split between so many listening posts that the level at each would be too low, even with the 20dB antenna preamplifier to be supplied. The manager in charge, looking for a solution, was told by a mischievous engineer that another similar preamp could always be added in series with the first, after all 40dB gain should be more than enough. His response: "Oh, can you add dBs, just like that?"

From Russia with love

Last year, AT&T Co announced it was hiring one hundred scientists working on optical fibre research at the Russian Academy of Sciences in Moscow. The Russian scientists will be paid the same as before the deal - about \$60 per month - which must represent real bargain basement research. Corning Inc has said it also was planning to hire about the same number of scientists, working on basic research in glass. Both companies say that the Russians would retain any resultant patent rights within Russia and would get bonuses for outstanding work.

It makes a change from financing low cost revolutions in South America.

Virtual (harsh) reality

The patent rights relating to virtual reality previously owned by the US company VPL Research are now the property of the French company Thomson CSF SA. This blow to US efforts to stay competitive in the field came about because VPL had pledged the patents as collateral for a one million dollar loan from Thomson CSF, on which it subsequently defaulted.

I suppose it's just another form of inward - oops - virtual investment. ■

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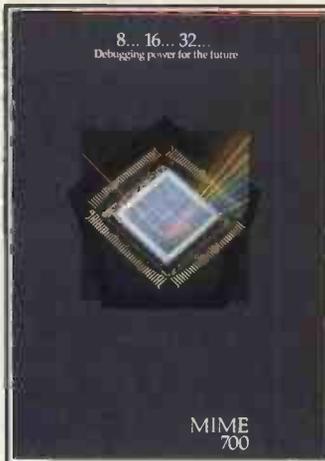
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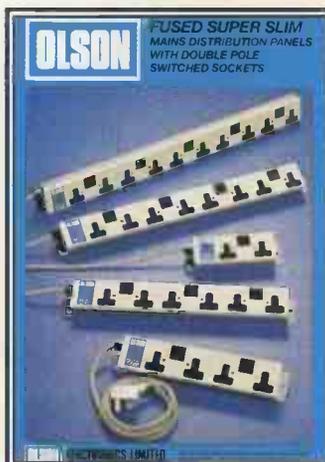
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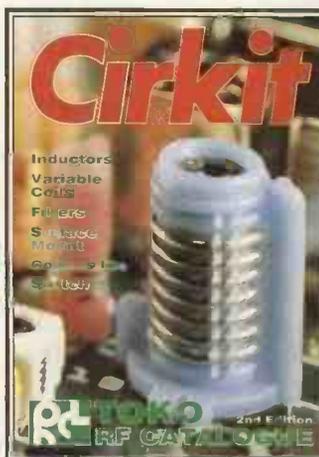
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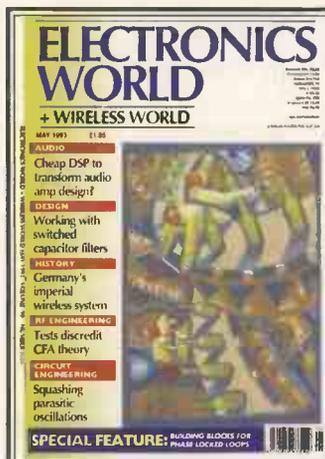
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