Beginners' corner – design your own Colpitts oscillator

ELECTRONICS WORLD

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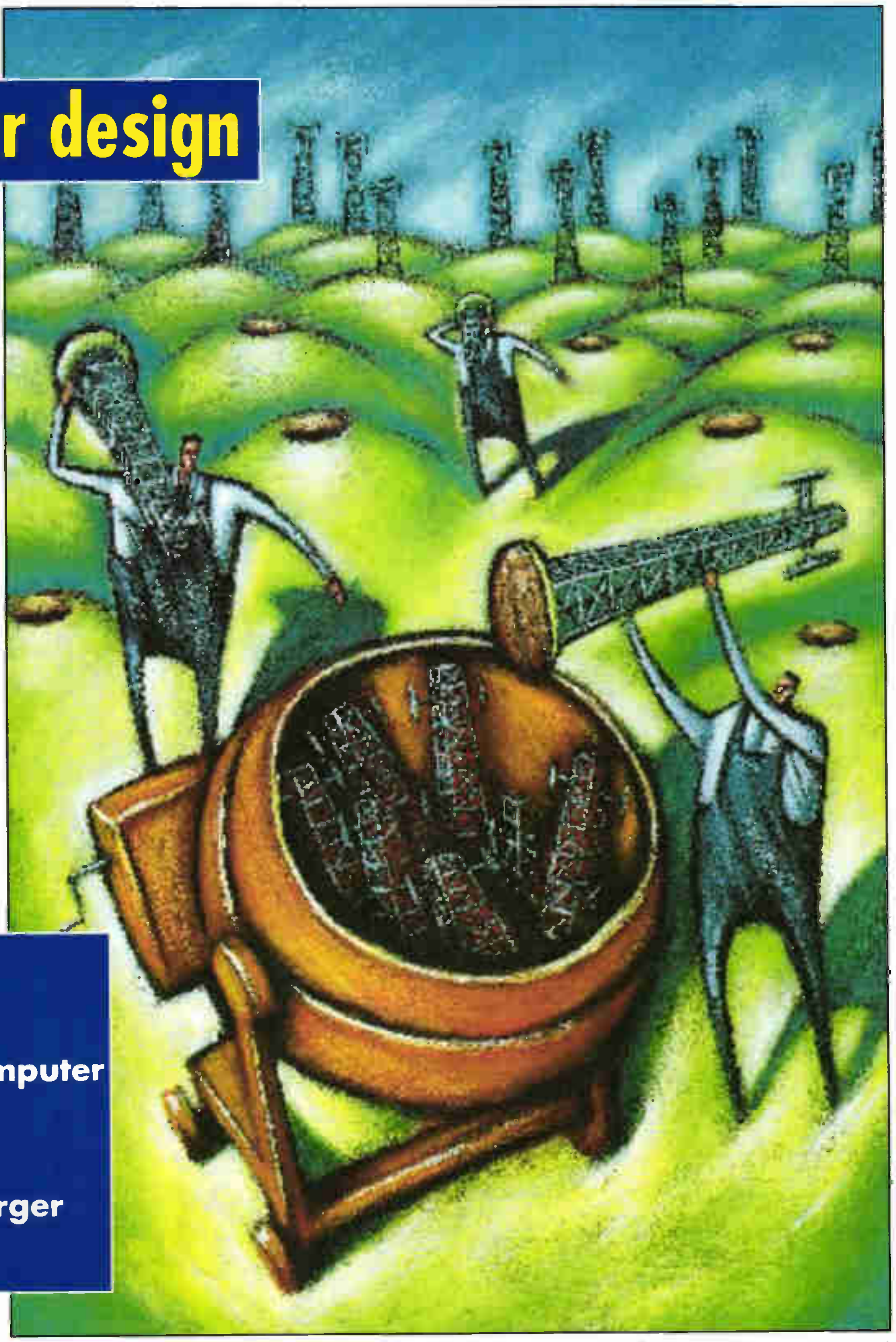

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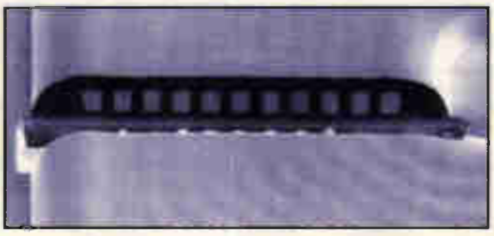
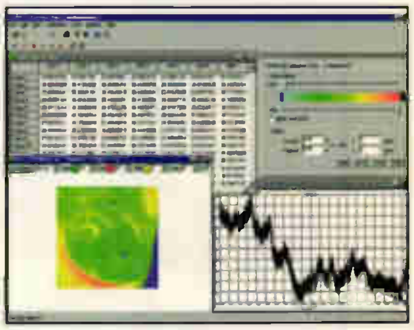


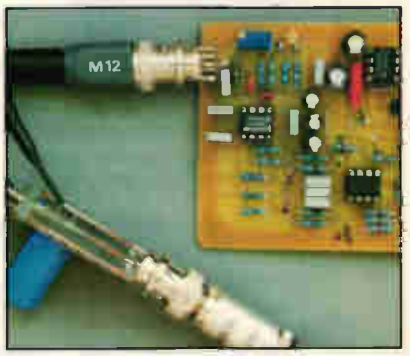
Is the opto-coupler dead? Find out what the benefits of an alternative coupler using micro-machined inductors are on page 677.

Illustration Mark Oldroyd

Ever-changing light patterns: electronics adds a new dimension to art. Read how on page 706.



Optoelectronics spots cancer cells in seconds using a new technique involving just an endoscope, light beam and some DSP. Dave Mitchell reports on page 694.



To evaluate a capacitor for a modern switched-mode power supply, a good test jig is vital. Find out what Cyril Bateman uses on page 696.

October issue on sale 7 September

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Whither WAP?

Shock, horror, hold the front page... sales in the UK of WAP-capable mobile phones – you know, the ones that work as Internet terminals as well – have achieved just 200 000 instead of the projected half million.

But perhaps you didn't know – or didn't even care? That seems to be the reaction of the public at large, even if dealers, manufacturers and pundits are wailing and gnashing their teeth. Building societies and mobile dealers are reduced to giving away WAP phones as a come-on, while the prophets of doom and desperation portray WAP as a bigger techno-flop than the Millennium Dome – as if that were possible. But is WAP really a click too far?

It depends on your outlook. Undoubtedly the product has been oversold if the following nonsense is typical of the sales pitch for WAP – and no, I won't embarrass the perpetrator!

"I'm in the pub, where I order a round of drinks, paying for them from my mobile and join my friends.

"We gather round one mobile to watch highlights of what movies are currently playing at the local cinema. We make a reservation through the mobile and check out the view from our seats on line.

"I also order some food to be ready for me to pick up when I arrive. I like the film so much that I buy and download the soundtrack to my mobile while I am watching.

"After the movie I am too tired to walk home so I call a taxi. I allow the taxi company to find my location and I watch the taxi approach on a map on my mobile. The taxi is paid automatically when I get out and the lights in the house come on as I approach.

"As I settle down with a glass of wine, the soundtrack I bought at the cinema is transferred to my hi-fi to end a great night out!"

Perhaps life really is like that on planet WAP – but not down here.

Clues to WAP's poor performance to date are not hard to find. While many people find the ability to make phone calls on the move convenient and reassuring, it's less clear whether they have the enthusiasm, requirement or the patience to download stock quotes or e-mail or make travel bookings and other transactions over a connection that's slow – at 9600 bit/s – expensive – at 5p or 10p a minute – and hardly a showpiece of legibility or graphic presentation.

For WAP browsing online, content must be redesigned from scratch, without even the benefit of open standards. WAP technology is proprietary and sold under licence, controlled by a group called the WAP Forum embracing the main companies involved.

The marketing of WAP has been less than scintillating too. There's abundant evidence that successful extraction of customers' money demands creating an awareness of need or desire; few products sell themselves.

Even when the merchandise itself is appealing, a price that's pitched higher than the public is prepared to pay will restrict sales to all but 'must have' early adopters. Pocket calculators, microwave ovens, video recorders, home computers, CD players – the list is endless of new technologies that took a while to catch on.

It doesn't help that today's offerings are less than compelling. Even the Carphone Warehouse – arguably Britain's most successful independent purveyor of mobile communication – is forced to concede:

"There is a lot of current hype in the media about what WAP is, and while the possibilities in the very near future are exciting and limited only by the imagination, WAP is currently a predominantly text-based view of the Internet. Improvements to the current mobile networks and in screen and battery technology will all enhance the WAP experience, eventually delivering rich multi-media content in the not too distant future."

Full marks for honesty!

To write off WAP entirely would be wrong, though – its time will eventually come when prices fall and product specifications improve. Combining a WAP-enabled phone with an electronic organiser, palmtop PC, portable FM radio, MP3 player and TV would make a far more compelling proposition. And then, finally, a mass market could be born.

Mobile phones are a classic examplar; those clumsy featureless 'bricks' that cost £1000 back in 1985 were scarcely a runaway success, yet fifteen years on more than half the population now owns a cellphone and a sleek, highly featured and affordable one to boot. Commoditisation is the name of the game.

And that's it. Sorry, what's WAP? Yes, I should have defined the term at the outset but nobody else bothers to. It stands for Wireless Application Protocol, a name that conveys precisely nothing to the world at large.

Years ago, successful marketeers knew that to sell an obscurely titled technology, you had to give it a meaningful name. Touch-Tone, Video Plus and Walkman all give a clue to their function – but not WAP. Time for a re-launch with a new name perhaps? ■

Andrew Emmerson

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UK videophone setback

The launch of the world's first GSM mobile videophone has been delayed for up to six months, according to the UK companies behind the project.

The videophone from Orange was most recently scheduled for launch this summer but is now not expected to see the light of day until the last quarter of this year.

"The launch date is set for sometime in Q4," said an Orange spokeswoman, who refused to comment on why the project was delayed.

According to another member of the entirely UK-based project team, working prototypes of the videophone were delivered to Orange earlier this year and have also been on display at last week's Tomorrow's World Live exhibition in London.

In May of this year the company was still saying it would launch in the summer when high-speed data capability was added to the network in the form of HSCSD (high speed circuit switched data). HSCSD is set to roll-out at the end of July when corporate trials are completed.

A group of UK technology companies was chosen to work on the design and manufacture of the videophone last year and Spring 2000 was set as the date for launching the innovative product.

Cambridge Consultants has led the overall technical project management including both the industrial and mechanical design of the handset and creating the user interface.

ISDN videophone firm Motion Media, Microsoft system integrator NMI and the University of Strathclyde have been working on various different aspects of the project, with manufacturer Celestica.

The development work has included video compression software, video specific application software, control protocols, implementation of the user interface and the design, development and integration of the electronics.

HSCSD achieves 57.6kbit/s data

Internet on TV available this year

Digital terrestrial TV company ONdigital will be offering Internet on TV to its subscribers this year.

An Internet box, around the size of a large paperback book, which connects to the digital set-top box and a telephone line is needed to receive the service.

French company Netgem has been chosen as a technology partner and will be providing the Internet box which uses a system based on the Linux operating system.

Pace Micro Technology, which supplies ONdigital's set-top boxes, already has an Internet TV product on the market, but an ONdigital spokesman said the French supplier was chosen because its, "technology seems to work particularly well."

The Internet pages are instantly reconfigured and the system apparently makes the pages readable on a TV screen from 10-15 feet away.

"I haven't seen anything else like it," said the spokesman. Viewers will be able to click straight from programmes to the Internet while the TV picture remains in the corner of their screen.

"ONdigital will offer the easiest and most cost effective way of getting on to the Internet," said Stuart Prebble, ONdigital's CEO. "No dish, no cable, no computer. Simply a television and an ONdigital subscription."

The box will initially connect to the telephone line via a 56k modem but will be capable of connection to ADSL technology later. Details of price and the launch date will be made available in the next few weeks. There are plans to integrate the Internet box inside the digital set-top box.

Pace licensed its Web technology to Alba for use in its Internet browsing TV in March this year.

Display technology relies on ink-jet printing

Seiko-Epson and Cambridge Display Technology (CDT) have demonstrated a full-colour, active-matrix light-emitting polymer (LEP) display.

The display, which is a prototype, measures 64 by 64mm, has 200 by 150 pixels and 16 levels of grey scale – achieved through a combination of time modulation and the use of sub-pixels.

The display is notable, not only for being the first of its kind, but for being the first display made using ink-jet printing techniques. In fact, the relationship between the two companies was started when Seiko-Epson approached CDT realising that LEP technology would be a new outlet for Epson's printing expertise.

Producing the display has pushed ink-jet technology to its current limit. Printing 30μm features with 30μm droplets has required special treatment of the printable surface as well as custom print heads.

Our reporter present at the demonstration reports that the display had no problems with fast action, was acceptably bright for shaded viewing, could be viewed easily from the side and had good LCD-like colours, but not as well saturated as CRT colours.

"The display uses our previous generation of polymers," explained CDT technical director Jeremy Burroughs, "We now have polymers that can match PAL television colours, but they are not suitable for ink-jet printing yet."

The companies are aiming the displays firmly at the mobile-phone and PDA markets, and expect production within two years. Beyond this, the companies are confident they have the technology to push into all other display areas, including TVs. "We can make 20 or 30in, no problem," said Mr Shimoda, head of R&D at Epson, "the feasibility is very high, the advantage of light emitting polymer is high."

TiePieScope HS801 PORTABLE MOST

ABRITARY WAVEFORM GENERATOR-
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- The HS801: the first 100 Mega samples per second measuring instrument that consists of a MOST (Multimeter, Oscilloscope, Spectrum analyzer and Transient recorder) and an AWG (abritary waveform generator). This new MOST portable and compact measuring instrument can solve almost every measurement problem. With the integrated AWG you can generate every signal you want.

- The versatile software has a user-defined toolbar with which over 50 instrument settings quick and easy can be accessed. An intelligent auto setup allows the inexperienced user to perform measurements immediately. Through the use of a setting file, the user has the possibility to save an instrument setup and recall it at a later moment. The setup time of the instrument is hereby reduced to a minimum.

- When a quick indication of the input signal is required, a simple click on the auto setup button will immediately give a good overview of the signal. The auto setup function ensures a proper setup of the time base, the trigger levels and the input sensitivities.

- The sophisticated cursor read outs have 21 possible read outs. Besides the usual read outs, like voltage and time, also quantities like rise time and frequency are displayed.

- Measured signals and instrument settings can be saved on disk.This enables the creation of a library of measured signals. Text balloons can be added to a signal, for special comments. The (colour) print outs can be supplied with three common text lines (e.g. company info) en three lines with measurement specific information.

- The HS801 has an 8 bit resolution and a maximum sampling speed of 100 MHz. The input range is 0.1 volt full scale to 80 volt full scale. The record length is 32K/64K samples. The AWG has a 10 bit resolution and a sample speed of 25 MHz.The HS801 is connected to the parallel printer port of a computer.

- The minimum system requirement is a PC with a 486 processor and 8 Mbyte RAM available. The software runs in Windows 3.xx / 95 / 98 or Windows NT and DOS 3.3 or higher.

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CIRCLE NO. 150 ON REPLY CARD

Is electronic product gazumping back due to scarcities?

Gazumping could be back in the semiconductor market after a period in which shortages have – surprisingly – not translated into higher prices.

Normally, when the silicon cycle turns up, prices increase sharply. However, in the present up-cycle prices have stayed broadly flat.

Asked last week why this is so, Motorola Semiconductor CEO Fred Tucker, replied: "I wish I could figure that out".

Tucker added: "Price increases can be so detrimental to a customer relationship. Could you give them a price increase and get away with it? Yes – but boy would it affect the relationship".

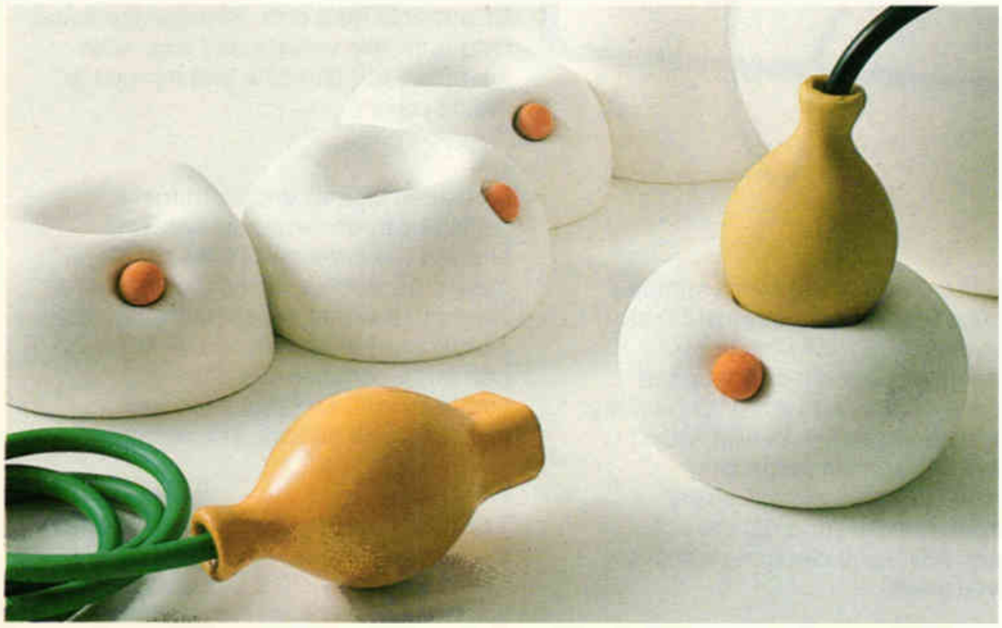
Four bad years have made chip manufacturers nervous about offending customers. But there are anecdotal signs from the industry (not related to Motorola) that things may be changing.

"We placed an order for 5000 serial EPROMs back in March for delivery in 22 weeks. Confirmed, acknowledged, everything," our reporter was told by Drew Hoggatt, managing director of access control system manufacturer Paxton Access, "We got to around week 19, and were told by the franchised distributor that we would not receive our EPROMs – ever."

The distributor told Hoggatt there was a supply shortage. "The supply shortage was of course a euphemism for 'You've been gazumped'," commented Hoggatt.

The suspicion was that someone in the supply chain had allocated the devices elsewhere – quite possibly at a better price.

"To add insult to injury," added Hoggatt, "the distributor offered us some devices which 'they found on the grey market' at five times the price on our purchase order to them. We could of course address the same market ourselves and we bought 10k devices at a 40 per cent premium on our original order unit price."



Giving art a plug... The annual design showcase at the Royal College of Art came round recently. Among the things of beauty there were things with practicality as well – particularly in the industrial design section. This Ergo Plug concept comes from Stephen Waldron who worked at Dyson Appliances last year. Waldron is on sj.waldron@virgin.net. You can read more about the exhibition at www.rca.ac.uk.

UK lags Europe with Internet broadband

The UK is slipping well behind Europe, and Europe slipping behind the US and Asia in upgrading the Internet to broadband capability, it appears from speakers at the DSL Summit in Colorado Springs.

The UK could be the last country in Europe to open up to free competition for the upgraded products. Whereas the EC has recommended that all countries in the EU should have free competition by the end of the year, the UK won't have it until July 2001.

While BT has the field to itself, for the next 12 months, it will charge £40 a month for the service, on top of the BT line rental charge, whereas speakers at the DSL Summit agreed that $40 (£25) a month was the figure which would actively encourage consumers to pay for broadband services.

Korea, Taiwan, Singapore and Hong Kong expect 4.5 million users in 12 months.

BSkyB must free up encryption technology

In a move to further open up the digital TV market, watchdog Oftel has decreed that encryption technology used by satellite TV company BSkyB should be made openly available.

The encryption technology is provided by Sky Subscribers Services (SSSL) and is used in BSkyB's digital set-top boxes to provide interactive services over digital televisions.

The watchdog has decided that SSSL is in a dominant position in the market and so in the interests of competition it must allow other companies to have fair access to its services so they can provide interactive services to their customers.

"As digital TV becomes more widespread, it is important that different companies can provide new and exciting interactive services to consumers," said David Edmonds, Oftel's director general. "But it is also important to ensure regulation in this area is only imposed when justified, otherwise competition and innovation in a fast-moving market could be stifled."

The ruling will be reviewed in the first half of 2001.

● The UK is leading the world in converting to digital TV, according to analyst company Strategy Analytics.

By the end of the year it, expects 29% of homes to have switched to digital. The US will follow with 24% while in France and Spain 15% will have converted.

The company forecasts there will be 56 million homes around the world watching digital TV by the end of this year. By the end of last year the figure was 34.4 million homes with satellite taking the major share of 77%. Cable achieved 21% and terrestrial just 2%.

Isolated drivers made by micromachining

A magnetic alternative to optocouplers has been announced by Analog Devices.

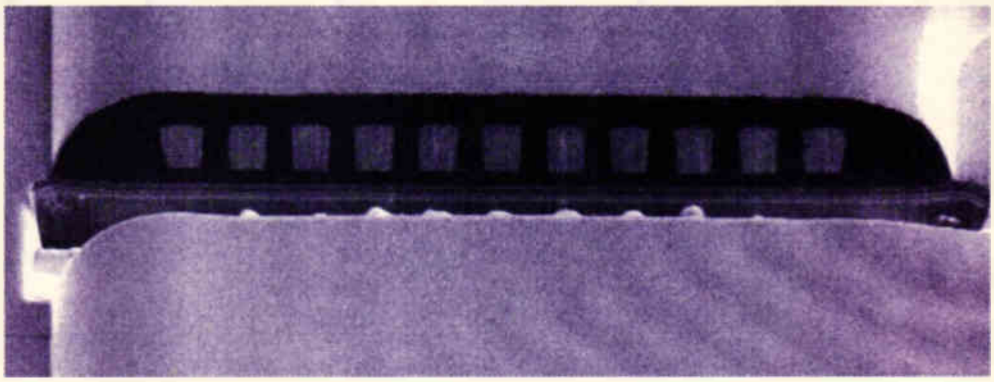
The 'μmIsolation technology' uses micromachining – hence 'μm' – to add coils to the die, isolating driver and receiver circuitry.

"We use what's called set-reset technology," said Ronn Kliger, business development manager at Analog Devices. The driver side of the chip looks for edges on the input. "Whenever it sees an edge it sends a short pulse to the top coils," said Kliger.

The receiver side of the die measures pulses using a comparator for detection. In order to distinguish between rising and falling edges there are two pairs of coils.

"With two pairs of coils we can send two bits of information," said Kliger.

Pulses sent to the coils are around 2ns long – a good achievement from the 0.6μm CMOS process. This length enables the isolator to cope with input data rates above 100Mbit/s.



	Optocoupler	μmIsolator
Data rate (Mbit/s)	25	100
Propagation delay (ns)	40	10
Transient immunity (kV/μs)	10	25
25Mbit/s consumption (mW)	95	30

To successfully couple the 2ns pulses across the coils, Analog Devices had to achieve a high inductance. To do this the firm uses a separate process for the micromachining.

A standard CMOS process can only manage coil depths of 2μm, while the firm's process can manage 8 to 10μm. Thus the 300μm diameter copper coils have an inductance of around 100nH.

South Korea scraps mobile phones subsidies

Mobile phone operators in South Korea have thought the unthinkable and agreed to government plans to scrap the subsidises on mobile phones in the country.

Faced with the enormous cost of investment in third-generation mobile phone services, the Korean operators have said they can no longer afford the subsidies which keep handset prices down and cost them 30 per cent of sales.

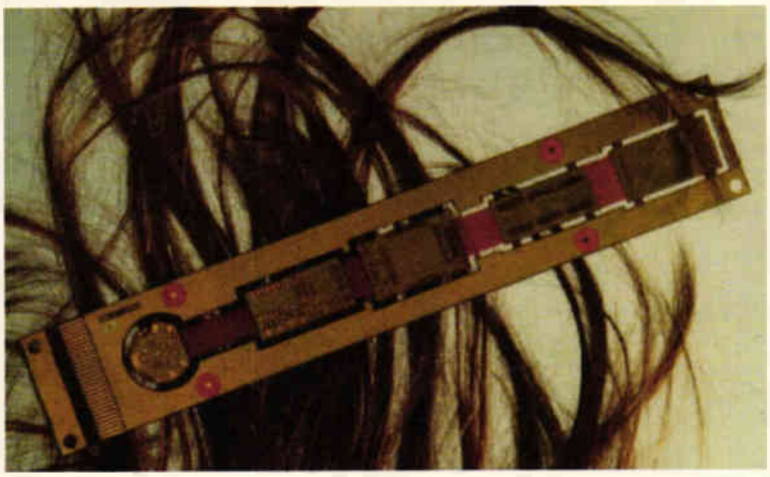
Korean plans to scrap subsidies on mobile phones could send shock waves through the world's mobile phone markets.

Low-cost mobile phones, which are heavily subsidised by network operators, are a way of life in most of the world's mobile phone markets, but South Korea has broken ranks by scrapping handset subsidies this month.

But similar moves are seen as highly unlikely in this country. "It is in no one's interest to stop subsidies," said a spokeswoman for UK operator Vodafone. "We are certainly not considering it."

What seems to worry the Korean government is the high level of foreign silicon in each subsidised mobile phone. It believes too high a proportion of the financial benefit of the subsidies paid by Korean operators is going to foreign chip suppliers.

For example, a trendy new handset valued at $200 is made of $100 to $120 worth of imported components, including $30 to $40 worth from Qualcomm's Mobile Station Modem 3000 chip and $10 worth of Intel flash memory.



Hairing aid... Micro-via PCB technology called DYCOstrate has been used to implement a programmable hearing aid by Swiss firm Sulzer Microelectronics. This hearing aid is the smallest of its type, the firm claims. The extreme packaging needs of a hearing aid led the firm to use the flexible substrate which can be folded up after manufacture. DYCOstrate is available in the UK through Rigid and Flex.

Wage rises for engineers stay low

Engineering pay settlements have remained low with the average pay settlement being 2.5 per cent over the last three months. The latest survey findings from the Engineering Employers' Federation (EEF) shows that nearly one in seven settlements were pay freezes in the three months to the end of May 2000. The EEF puts the low level down to companies facing "difficult economic conditions".

Modulator scheme for 25Tbit/s involves time-domain and wavelength multiplexing

Researchers in an inter-university project have accumulated £12.5m in grants and investment to develop 100Tbit/s fibre-based data communication techniques.

Behind the project is a modulation scheme that combines both time domain and wavelength-division multiplexing.

The modulator requires a source of 'hyper-short' 10fs pulses of 800nm light, although slightly broadened pulses of 1500nm light are also being considered.

Being so short, these pulses have a wide spectral content and can therefore be broken into individual 'colours' using a prism. This process is called 'spectral slicing'.

The sliced components – and you may have to get out your book of Fourier transforms here – are longer than the original pulse. The new width depends on how thinly the spectrum is sliced, but ten ways yields pulses of around 100fs, according to the University of St Andrews which is heading the project.

The wavefront of the spectrally separated 100fs pulses impinges on a multi-channel high speed modulator block featuring one channel per colour. The individual modulators block or pass the 100fs pulses to add digital bits to the wave front.

The multicolored, modulated wavefront is recombined into a single pulse with a second prism and a lens. The spectral content and duration of this pulse depends on the data content for any given pulse.

"These pulses," says Dr Tom Brown, assistant programme director at the University of St Andrews, "are suitable for transmission over local and metropolitan networks, but not transatlantic communication", adding "It could also be used between two adjacent computers or even within a PC".

At the receiving end, a third prism splits the pulse again and the re-sliced multi-coloured wave front hits an array of photodetectors which recover the original data.

It is reasonable to predict, says the university, that this modulator will provide the potential for ten separate 2.5Tbit/s channels, thus constituting a total transmission rate of 25Tbit/s. "We therefore believe that it will be necessary to develop advanced femtosecond-based technologies if data rates up to 100Tbit/s are to be made available," said the University in a statement – and this is what the six year project is all about.

There are two choices for the source of 10fs pulses. "We have some ideas for a semiconductor laser. The other option is a solid-state crystal laser pumped with a semiconductor laser. We can do the solid-state laser now."

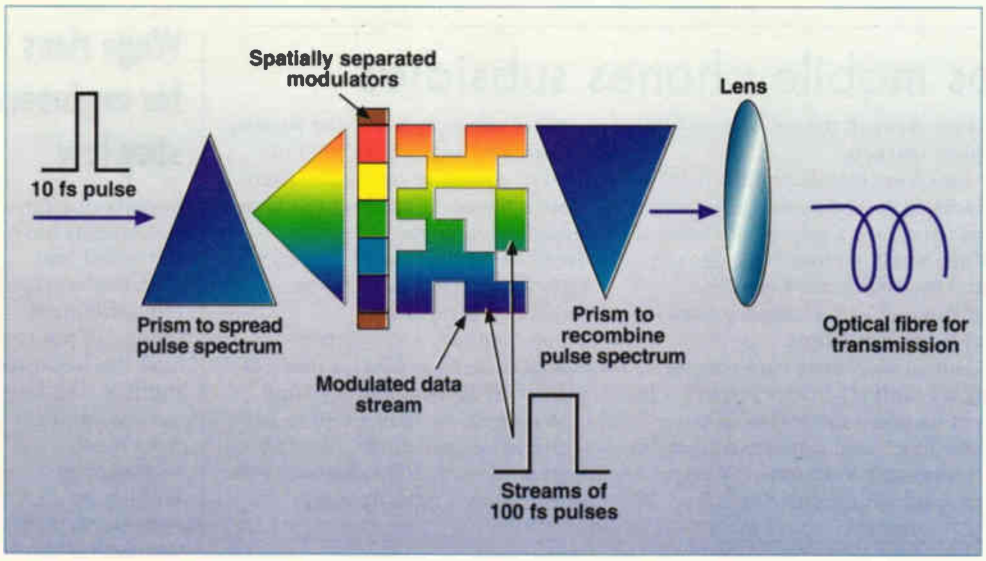
The technology challenge here, according to Brown, is to get the pulse repetition rate for the system under control, "then miniaturise the whole thing and make it rugged and cheap".

The prisms are acting as 'dispersive elements' and could in practice be diffraction gratings or, according to Brown, exotic structures in photonic bandgap materials.

The spatially-separated modulator idea is brand new according to Brown, but applicable technology already exists. Marconi already has suitable lithium-niobate modulators, "well into development, running at high bit rates. This is not hugely exotic technology," he said.

Photodetector technology for the receiver does not exist yet. But, says Brown, "Detectors are running at 40GHz now, but they need good quality pulses that have not been messed about."

Brown says that all of the components can be miniaturised and could be put on a single chip. ■



Government faces tax legal challenge

The government is facing a legal challenge over its controversial IR35 tax crackdown on contract employees which affects many self-employed electronics engineers and IT professionals.

The Professional Contractors Group (PCG), an industry body, is preparing papers to go before the court in the next few weeks requesting a judicial review of the case. The PCG believes IR35 contravenes European law and so has a strong case.

"We believed all along it [IR35] was unfair, wrong and would never work," said a PCG spokeswoman. "We're now following it up because the government looks intent on pushing this through. The challenge in the courts is because we think it is legally wrong too."

The PCG has also urged the government to abandon IR35 after a loophole was discovered which would allow large foreign-owned companies to gain a tax advantage over their smaller UK competitors by employing cheap, foreign labour.

It is claimed that foreign workers brought here on 'fast-track visas' to tackle the skills shortage are being paid a low wage and then rented to clients for an inflated fee. The difference can be treated as pre-tax profit and re-invested by big companies whereas small UK consultants must treat any profit as salary and cannot re-invest it in their business.

The legislation is intended to tackle tax avoidance by some self-employed staff working for larger electronics firms on temporary contracts.

A new low-IMD mixer

Chris Trask's new series-shunt feedback active mixer offers clear advantages over both the common Gilbert Cell active mixer and diode-ring mixers. With lower local-oscillator power requirements, low distortion, and higher saturable output power, this new mixer is highly suitable for low-power high-performance communications systems. Yet it's possible to implement the design on the kitchen table!

Mixers are essential building blocks of radio communications systems, being used for modulation, demodulation, and signal frequency conversion. Among the various forms have been transconductance multiplication – dual-gate FETs, pentagrid and heptode vacuum tubes, etc. – diode and switching FET rings, and the transistor tree – also know as the Gilbert Cell.

An inherent undesirable property of mixers has been – and continues to be – intermodulation distortion. Commonly abbreviated to IMD, intermodulation distortion is caused by two adjacent signals interacting. This interaction creates spurious signals that can interfere with adjacent smaller signals. In some cases, IMD can actually cause interference within wideband communications systems themselves.

Eliminating IMD

Overcoming this unwanted characteristic is no small task. Traditionally, the efforts at improving IMD have included using Class III diode ring and switching FET ring mixers that generally require local oscillator signal levels of +17dBm or more. This is an unsuitable solution for field-portable equipment though, where power consumption is an important parameter.

Active mixers, such as Plessey's SL6440 and the Motorola's MC1496, make use of the six-transistor double-balanced transistor tree. Emitter degeneration is applied to the driver transistors to provide some degree of linearisation.

Still more recent methods regulate the leg currents using negative feedback, but they do not encompass all sources of distortion in the mixer.

These sources must include the switching transistors as well if the correction is to be truly effective.

Until now, it has been generally considered that the mixer is an open-ended device. This means that there is no obvious opportunity to apply the traditional linearisation techniques such as feedback.

Notable exceptions have been Plessey's SL6440, devised by Phil Moon, which originally used a clever negative feedback amplifier for linearising the leg currents[1,2].

This linearisation amplifier required p-n-p transistors. Because of the difficulties involved in making suitable p-n-p transistors though, the original circuit was sadly dropped in favour of the more easily fabricated version with emitter degeneration resistors[2].

In more recent years, a US patent by Joseph Heck describes an active mixer in which the leg currents are linearised by way of using transconductance amplifiers[3]. A later circuit by Barrie Gilbert uses Norton current input amplifiers to achieve the same results[4]. These last two are specific embodiments of an earlier generalised method patented by Daniel Talbot[5]. All of these methods produce an appreciable improvement, but they fall short of the mark in terms of complete linearisation.

Presented here is an active mixer circuit that produces substantial IMD improvement. At the same time, it uses far less local oscillator (LO) power than a diode ring mixer of comparable performance.

Table 1. Measured performance data for the new mixer configuration.

Circuit configuration	Gain	Input intermodulation intercept point (IIP_3)	Compression point (P_{1dB})
Mini-Circuits SBL-1	–5.0dB	+19.0dBm	–4.5dBm
Typical Gilbert-cell mixer, see text.	–1.5dB	+17.5dBm	+4.5dBm
1st-generation series-shunt feedback mixer, Fig. 3.	–7.0dB	+21.5dBm	+5.5dBm
2nd-generation series-shunt feedback mixer, Fig. 4.	–3.0dB	+29.5dBm	+10.5dBm

By using a more detailed method of combining the four signal currents in a transistor tree mixer, the amplified intermediate-frequency (IF) signal voltage can be recovered at the mixer output and then used in a traditional feedback topology, thereby including all sources of IMD in the feedback scheme[6,7,8].

In addition, the radio-frequency output is derived from the combiner as well, and with a good degree of IF-RF and LO-RF isolation. Moreover, the circuit can be made to operate from the baseband modulator and demodulator level to VHF frequencies using components that are readily obtainable from common hobbyist sources.

Active mixer basics

Before I get into specifics, to help you appreciate this new concept, I'll examine the currents within a generalised transistor tree active mixer with the aid of **Fig. 1**.

Driver transistors Tr_3 and Tr_6 generate collector currents I_3 and I_6, respectively. These currents each contain a quiescent DC bias current and a differential signal current.

Signal currents of I_3 and I_6 are a direct consequence of the input differential IF signal voltage. Both are chopped, or modulated, by switching transistor pairs Tr_1/Tr_2 and Tr_4/Tr_5, respectively. This results in a foursome of output currents I_1, I_2, I_4, and I_5. These contain components of DC, IF, LO, and RF signals, each with a unique phase relationship. The accompanying panel on double-balanced active mixers gives a more detailed explanation for those so interested.

In the traditional double-balanced active mixer, the currents I_1 and I_5 are combined by connecting the collectors of Tr_1 and Tr_5 together, and the currents I_2 and I_4 are combined by connecting the collectors of Tr_2 and Tr_4.

Combining currents in this way leaves an output differential RF current and effectively cancels the output LO and IF currents. This last cancellation deprives us of the signals needed for the implementation of a feedback linearisation scheme.

Signal combining and recovery

Now consider other combinations of the four output currents, using **Fig. 2** as

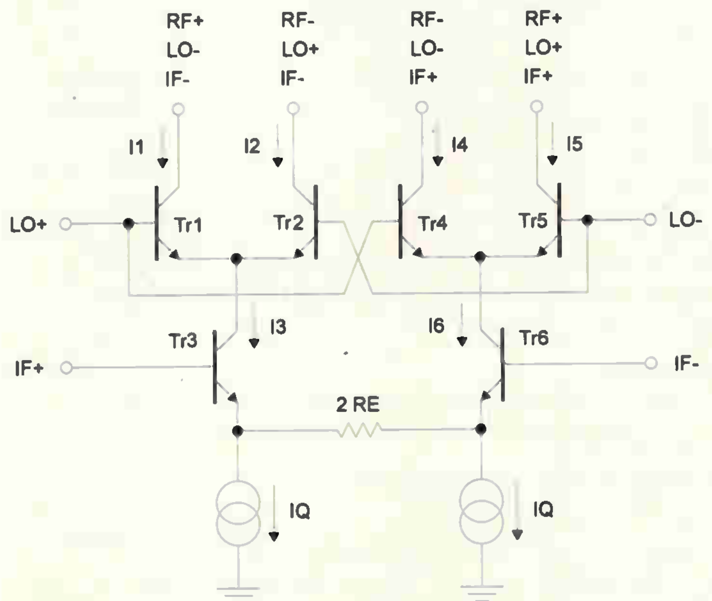


Fig. 1. Currents involved in the basic transistor tree mixer.

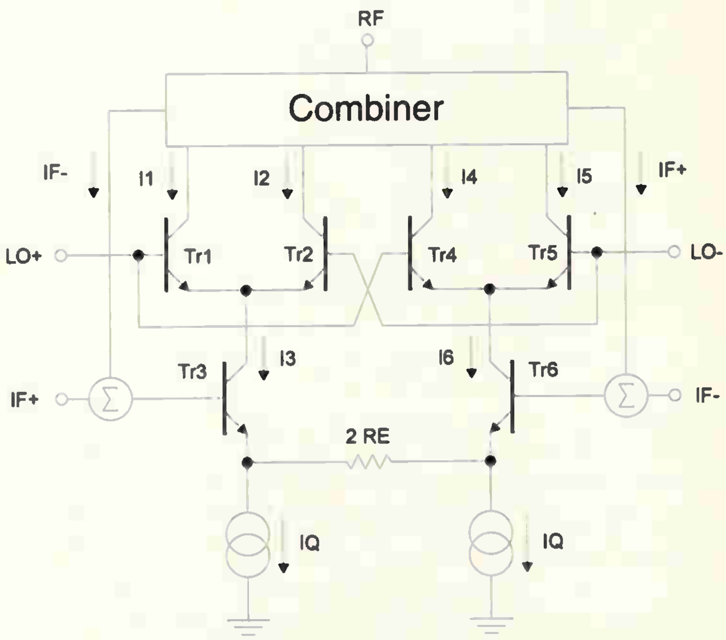


Fig. 2. Feedback mixer signal combining and recovery configuration.

a guide. In a general sense, the desired outcome of the mixer is to have an output RF signal that is a linear combination of the input intermediate-frequency and local-oscillator signals.

By using a saturating LO signal for the switching transistors, you are left with the burden of ensuring that the amplified IF currents at the four switching transistor collectors are a faithful linear reproduction of the input signal voltages. Therefore, an output feedback signal needs to be provided for each input signal to be used in comparison and subsequently correct for any errors.

First, if I_1 with I_2 are summed, the amplified and inverted IF signal current for the left side can be recovered. Similarly, by summing I_4 and I_5, the amplified IF signal for the right side can be recovered. This can be accomplished in a variety of ways. Two of these are described in detail later, after the preliminaries are taken care of.

Once the amplified IF signals are recovered, they can be used as feedback signals for use in the linearisation of the

The double-balanced active mixer

Here, I'll briefly examine of the core element of a double-balanced active mixer, as in **Fig. A**.

Drive transistors Tr_3 and Tr_6 convert the input IF voltage into a pair of differential currents I_3 and I_6,

$$I_3 = I_Q + \frac{A\cos(\omega_S t)}{R_E + r_e} \qquad (1)$$

$$I_6 = I_Q - \frac{A\cos(\omega_S t)}{R_E + r_e} \qquad (2)$$

Here, A is the amplitude of the input IF voltage, I_Q is the quiescent bias current for each leg, and r_e is the nonlinear emitter resistance of Tr_3 and Tr_6, assumed to be equal for both devices.

Fixed resistance R_E is used to establish the mixer signal (and conversion) gain as well as degenerate the driver transistors to provide for some stability over temperature and a small degree of linearisation. These two currents are then passed on to a pair of differential switching transistor pairs, Tr_1/Tr_2 and Tr_4/Tr_5, where an applied LO signal causes each of these currents to be divided into two differential currents. Ignoring the higher-order terms, these currents are:

$$I_1 = \frac{I_Q}{2} + \frac{A\cos(\omega_S t)}{2(R_E + r_e)} - \frac{I_Q\cos(\omega_L t)}{2} - \frac{A[\cos(\omega_S - \omega_L)t + \cos(\omega_S + \omega_L)t]}{2(R_E + r_e)} \qquad (3)$$

$$I_2 = \frac{I_Q}{2} + \frac{A\cos(\omega_S t)}{2(R_E + r_e)} + \frac{I_Q\cos(\omega_L t)}{2} + \frac{A[\cos(\omega_S - \omega_L)t + \cos(\omega_S + \omega_L)t]}{2(R_E + r_e)} \qquad (4)$$

$$I_4 = \frac{I_Q}{2} - \frac{A\cos(\omega_S t)}{2(R_E + r_e)} + \frac{I_Q\cos(\omega_L t)}{2} - \frac{A[\cos(\omega_S - \omega_L)t + \cos(\omega_S + \omega_L)t]}{2(R_E + r_e)} \qquad (5)$$

$$I_5 = \frac{I_Q}{2} - \frac{A\cos(\omega_S t)}{2(R_E + r_e)} - \frac{I_Q\cos(\omega_L t)}{2} + \frac{A[\cos(\omega_S - \omega_L)t + \cos(\omega_S + \omega_L)t]}{2(R_E + r_e)} \qquad (6)$$

There's a great deal of information contained in these four currents. No two are alike, all four containing elements of RF, LO, and IF signals each with a unique variety of phase relationships.

In equations 3 to 6, the first term is the quiescent bias current, the second is the amplified IF signal current, the third is the LO current, and the fourth is the mixing product RF current. By making various combinations we can recover one and eliminate the other two, the most obvious of which is the combining of I_1 with I_5 and I_2 with I_4, illustrated in Figure 1, which creates a differential pair of RF output currents I_7 and I_8, cancelling both the IF and LO signals currents:

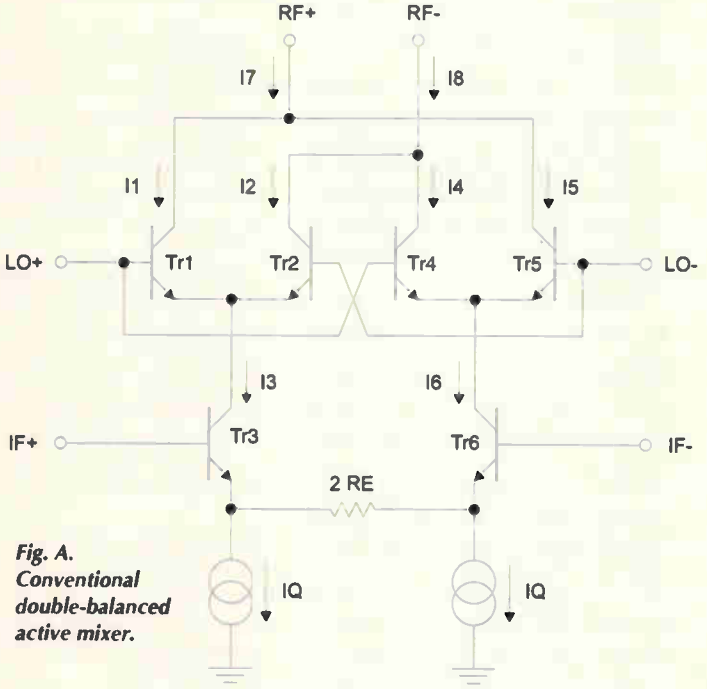


***Fig. A.* Conventional double-balanced active mixer.**

$$I_7 = I_Q + \frac{A[\cos(\omega_S - \omega_L)t + \cos(\omega_S + \omega_L)t]}{2(R_E + r_e)} \qquad (7)$$

$$I_8 = I_Q - \frac{A[\cos(\omega_S - \omega_L)t + \cos(\omega_S + \omega_L)t]}{2(R_E + r_e)} \qquad (8)$$

This is the very basis of the double-balanced active mixer, which was originally patented in the form of a synchronous demodulator by Howard Jones in 1966[1]. It was later given the name Gilbert Cell after its subsequent use in a later patent for an analogue multiplier in 1972 by Barrie Gilbert[2].

The obscurity of the work by Jones has been detrimental in his being recognised as the original inventor of this circuit. A more detailed explanation can be found in references 3 and 4 below, as well as numerous other works on semiconductor circuit theory and design.

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mixer as a whole.

The earlier combining of I_1 with I_5 and I_2 with I_4 is still a practical choice for recovering an output RF signal while at the same time cancelling the undesired IF and LO signals at the output. But it would be preferable to do this and still accommodate the recovery of the amplified IF signals.

An alternative method is to take the difference of I_1 and I_2. This yields a positive RF signal and a negative LO signal on the left side. A similar difference between I_5 and I_4 on the right side also yields a positive RF signal, but now the LO signal is positive.

By adding these two signals together you arrive at a positive RF output and a cancellation of the LO signal, which is the desired outcome.

With the amplified IF signals recovered, you can now consider an appropriate feedback amplifier topology. For baseband modulation, a pair of operational amplifiers can be used to compare the input IF signal with the feedback IF signals. Then, the amplified

The series-shunt feedback amplifier

There's a handful of circuit topologies that are suitable for RF feedback amplifiers. Perhaps the most straightforward of these is the series-shunt feedback amplifier[1,2].

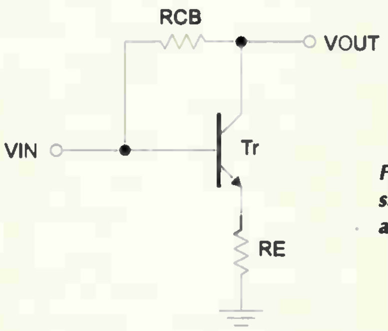

Fig. B. Basic series-shunt feedback amplifier.

This circuit, shown in **Fig. B**, is a good example of how the simplest of inventions can be the most profound. These simple wide-band amplifiers are easily designed using the following simple relationships for input/output resistance and gain:

$$R_{in} = R_{out} = \sqrt{R_{CB} \times (R_E + r_e)} \quad (1)$$

$$G = 1 - \sqrt{\frac{R_{CB}}{R_E + r_e}} \quad (2)$$

where R_{CB} is the collector-base feedback resistance, R_E is the emitter feedback resistance, and r_e is the incremental non-linear emitter resistance of the transistor.

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Hybrid transformer

The hybrid – or bridge – transformer[1,2] has been with us for well over a century. It was originally used in telephony to enable two-way communication over a single pair of wires, with earth being the required third connection.

As shown in **Fig. C**, the hybrid transformer is simply a two-winding transformer with a centre-tapped primary winding.

The various currents and voltages conform to the following relationships,

$$I_B = I_A + I_C \quad (1)$$

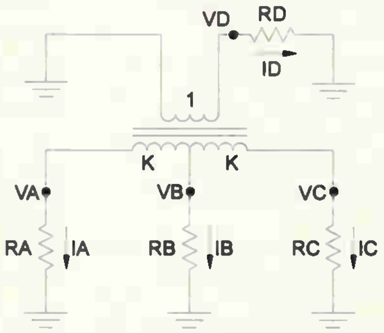

Fig. C. Hybrid transformer – originally used in telephony to enable two-way communication over a single pair of wires.

$$I_D = (I_A - I_C) \times \mathrm{K} \quad (2)$$

$$V_B = \frac{V_A + V_C}{2} \quad (3)$$

$$V_D = \frac{V_A - V_C}{\mathrm{K}} \quad (4)$$

In other words, the common-mode voltage and current appears at the centre-tap while the differential voltage and current is at the secondary. The differential voltage and current are scaled by the turns ratio K. It also follows that,

$$R_A = R_C = 2R_B \quad (5)$$

$$R_D = \frac{R_C}{\mathrm{K}^2} \quad (6)$$

The hybrid transformer can be found in numerous applications in telecommunications, including repeater amplifiers, crystal lattice filters, amplifier neutralisation, and frequency multiplexing.

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difference can be applied to the driver transistors.

For high-frequency mixer applications, the use of an operational amplifier becomes impracticable. Instead, a very simple feedback amplifier topology commonly referred to as series-shunt can be used. These simple yet effective amplifiers require two additional resistors to the common-emitter transistor amplifier. A detailed discussion is provided in the accompanying panel.

First-generation series-shunt feedback active mixer

Figure 3 illustrates the first generation of what has become known as the series-shunt feedback active mixer[6,7]. Resistors $R_{5A\text{-}D}$ serve as both IF feedback signal combiners and as the shunt feedback resistances for the IF feedback amplifier.

The combining takes place at their common junction with the bases of the driver transistors. Since each of these carries only half of the total IF feedback signal current, they are of the same value as would be derived for a single amplifier stage.

Resistors $R_{7A\text{-}C}$ perform the series feedback function for the IF feedback amplifier as well as establishing the quiescent bias currents in lieu of using dedicated current sources, and with a value of 100Ω each they produce an R_E of 33Ω, giving the mixer an IF signal amplification factor of around –2.16, or 6.7dB All capacitors shown are 10nF.

Resistors $R_{4A\text{-}D}$ provide collector biasing pull-ups. The resistors $R_{4\text{-}7}$ are in the form of SIP networks in order that the physical design be somewhat elegant and providing for the practical purpose of tracking over temperature.

Transformers $T_{1\text{-}3}$ are identical, being of a 1:1:1 ratio, in which case the centre tap of T_3 is not used. A commercially available transformer such as Mini-Circuits' T4-1 may be used here, or you can make one by twisting three wires together on a suitable ferrite core.

It is best to use transistors in the form of an array so that parameter matching doesn't become an issue. Intersil's (née Harris, née RCA) CA3054 and CA3102 arrays come to mind, the former being less costly and more plentiful.

The 3102 should be used where higher frequency performance is needed. Other transistor arrays, such as the series of dual transistors made by Panasonic and NEC, are equally suitable.

Performance in practice

We built an example circuit using the CA3054 for the transistor array using the components values shown in **Fig. 3**. For T_1, T_2, and T_3, we used the Mini-Circuits T4-1 transformer.

With a supply of 12V, the resulting quiescent bias current was approximately 12mA for each of the driver transistors Tr_3 and Tr_6. In addition, we made a comparable Gilbert Cell mixer. Its circuit was essentially that of Fig. 3 with the shunt feedback resistors R_5 and the associated DC blocking capacitors removed.

For the test signals, a local-oscillator signal of 10.0MHz at a level of 0.0dBm was used, although the circuit performs equally well at levels as low as –10dBm. For intermodulation measurements, the IF input signals were set at 500kHz and 510kHz. For conversion gain measurements, the first of these was used.

Mini-Circuits' popular SBL-1 diode ring mixer was also tested for comparison. The signals used were as for the active mixers, except that the local oscillator signal level was set at +7.0dBm to ensure proper operation.

Test results, listed in **Table 1**, indicated that there is a marked improvement in the distortion characteristics over the comparable Gilbert Cell mixer. Input intermodulation intercept point, IIP_3, is increased by 4.0dB. The 1dB compression point, P_{1dB}, is also improved by 1dB.

Closing the series feedback loop, however, decreases the conversion gain by 5.5dB. Even so, this is a considerable improvement over the SBL-1 – particularly in regard to the P_{1dB} compression point.

Using resistors in the output signal combiner of the first generation series-shunt feedback mixer was a matter of convenience. It was necessary in order to create a circuit that was suitable for possible future MMIC implementation. Their use does not impair the linearisation qualities of the mixer, but they do result in a decrease in conversion gain and in the apparent IIP_3 and P_{1dB}.

2nd-generation series-shunt feedback active mixer

By applying a pair of hybrid transformers as a means of combining the four switching transistor collector currents, as shown in **Fig. 4**, the compromises of the first generation series-shunt feedback mixer are decisively dealt with.

The result is a series-shunt active mixer with markedly improved performance[7,8]. This represents the second method discussed earlier in the

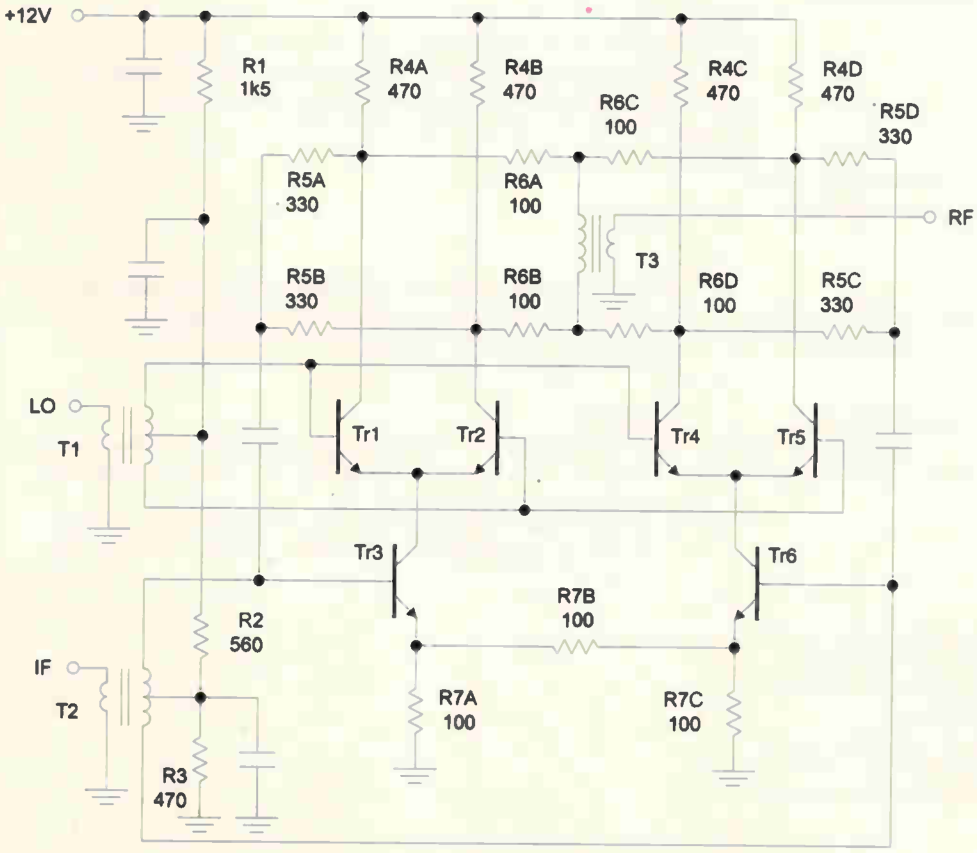

Fig. 3. First-generation series-shunt feedback mixer. Test results show that this circuit gives a marked improvement over the Gilbert cell.

combining and recovery of signal voltages.

Simply stated, the hybrid transformer combines the two pairs of collector currents in the following manner. Collector currents of switching transistors Tr_1 and Tr_2 are applied to the primary winding of hybrid transformer T_3. As a result, the common-mode components are added together at the centre-tap, while the odd-mode terms cancel.

Conversely, the odd-mode terms of are added together at the secondary winding, while the even-mode terms are cancelled. At the same time, the collector currents of switching transistors Tr_4 and Tr_5 are processed by hybrid transformer T_4 in a similar way.

Clearly, the LO signal appears at the secondary windings of T_3 and T_4. But the windings are respectively 180° out of phase, and therefore cancel, as was discussed earlier.

With the minor exception of bulk and induced losses in the transformer windings, the process of combining and recovering the signals is virtually lossless. This is a highly desirable circumstance.

The components shown in Fig. 4 follow very much with those of Fig. 3. However, there is now a single feedback resistor for each half of the mixer, these being resistors R_4 and R_5.

Hybrid transformers T_3 and T_4 are the same Mini-Circuits T4-1 transformers as used for T_1 and T_2 earlier. And as before, all capacitors are 10nF. All testing conditions remain as before.

Referring again to Table 1, the performance of the second-generation series-shunt feedback mixer greatly exceeds that of the earlier version. By replacing the lossy resistive combiner network with a pair of hybrid transformers, the conversion gain has been improved by 4.0dB, the IIP_3 by 8dB, and the P_{1dB} compression point by 5.0dB.

Although the open-loop Gilbert Cell mixer still has the advantage of slightly higher conversion gain, the second-generation series-shunt feedback mixer excels in all other respects. And it is a substantial improvement over the SBL-1.

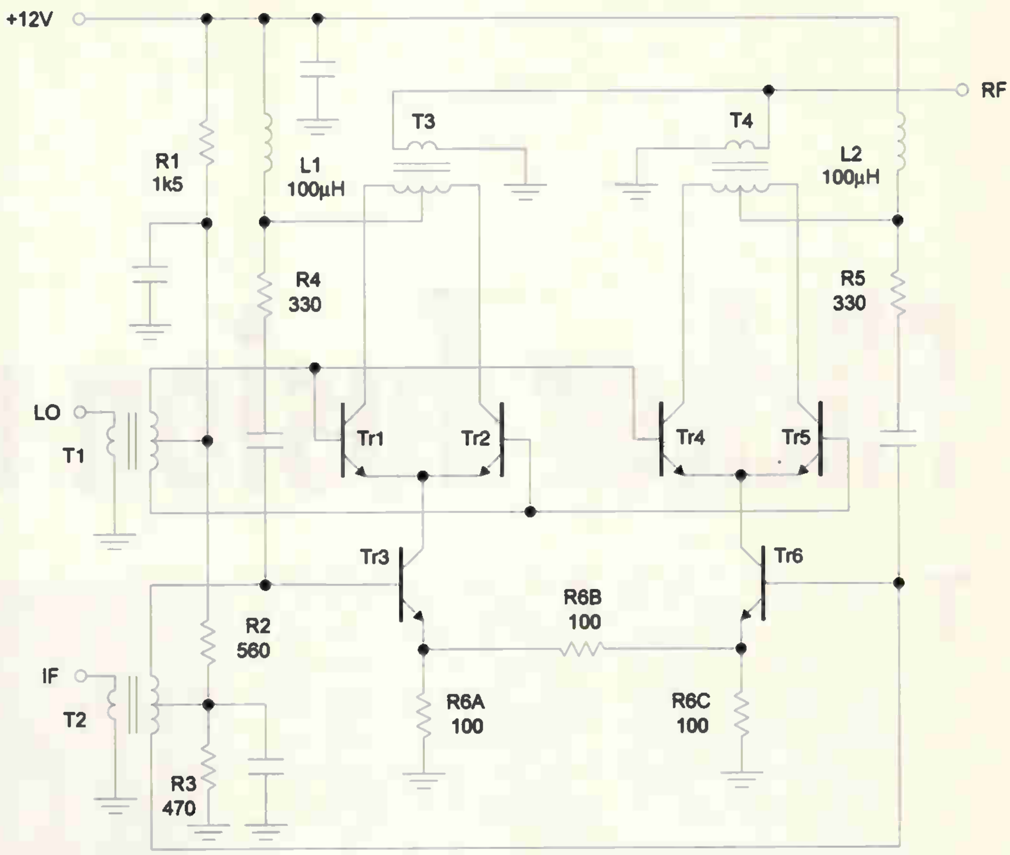

Fig. 4. Second-generation series-shunt feedback mixer. Adding a pair of hybrid transformers to combine the four collector currents eliminates the compromises of this circuit's predecessor.

In summary

The series-shunt feedback active mixer offers definite advantages over both the common Gilbert Cell active mixer and diode ring mixers.

With lower local-oscillator power requirements, low distortion, and higher saturable output power, the series-shunt feedback mixer is highly suitable for low-power high-performance communications systems.

Using resistors in the output signal combiner is straightforward and convenient, rendering the series-shunt feedback mixer suitable for MMIC implementation. With hybrid transformers used in lieu of the resistors, the results are a mixer of incomparable performance.

Noise figure – another important factor in mixers – was not addressed in these designs. This was because the mechanism that causes current-commutating mixers such as these to be noisy is still present[9].

It is entirely feasible that other topologies can be employed that will allow this characteristic to be improved, while at the same time retaining the desirable linear characteristics shown here.

For now though, the series-shunt feedback mixers presented here represent a method by which the dynamic range of this important element in radio design can be greatly improved, without excessive local-oscillator power generation or DC power consumption increases. ■

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Flicker fusion meter

The photoreceptors of the eye and the neurons of the visual system are relatively sluggish transducers. This is particularly true for the cones – the receptors that are colour sensitive, and respond in bright light. Rods on the other hand have faster responses. They signal intensity only – not colour – and are principally active in dim light.

The cones are largely clustered in the centre region of the retina, while the rods predominate around the periphery. The rods' faster responses and peripheral location explain why many people are able to detect the flicker of fluorescent lighting out of the corners of their eyes.

To the slower acting cones, this flicker is imperceptible. But there is of course a frequency below which flashing can be detected – even by the cones.

The flash rate at which the illumination just appears to be continuous is termed the critical flicker fusion frequency, which is frequently shortened to CFF. For many people the critical frequency is in the region of 20 to 40Hz.

Factors affecting fusion frequency

A number of factors affect CFF. These include, as indicated above, whether the light is viewed directly or with the eye slightly averted, so that the image is perceived peripherally. Brightness has an impact, since the visual system responds more quickly in bright light.

The receptors are 'photon gatherers'. Once they have captured sufficient photons a neural impulse is generated. Since bright light has a higher photon rate, the neural response is more rapid. In poor light there is an appreciable integration time, which is recognised in the game of cricket, when bad light is allowed to stop play. A fast moving cricket ball could hit a player, before its location had been recognised.

The panel entitled, 'A dim view is a slow view' describes a particularly compelling demonstration of the relatively slower processing of dim light.

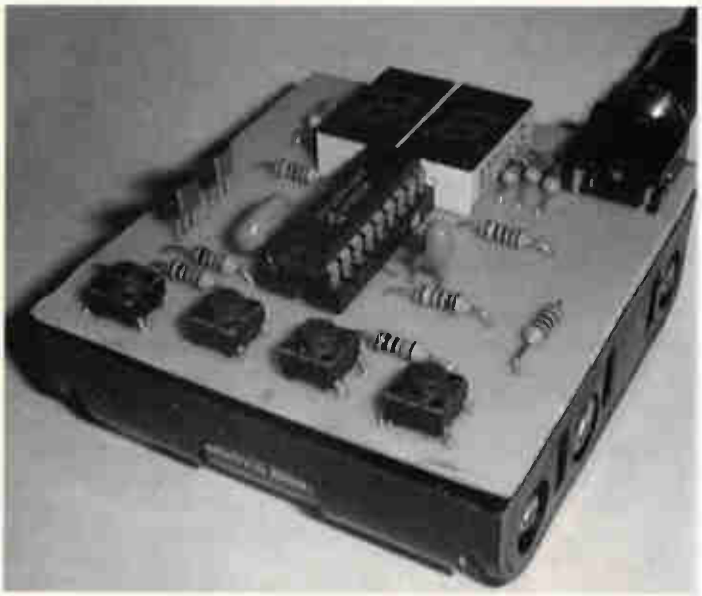
Not only do characteristics of the light stimulus modify CFF; so also do factors relating to the viewer. There is, for example, a clear impact of age. Older viewers exhibit a lower fusion frequency. In other words, their visual system is unable to resolve such high rate flashes as can younger people[1].

The stimulating effects of coffee and tea raise CFF[2], but there is some indication that this effect is influenced by personality. Extroverts are said to have lower base-line neural reactivity than introverts[3]. This is claimed to explain why, with their greater 'head-room', extroverts enjoy raising their activation by engaging in exciting activities. Consistent with this description, it is extroverts rather than introverts who appear to show the CFF speed-up effect of caffeine[4].

Sensitive to fatigue

Of particular interest, CFF has been shown to be sensitive to fatigue, and can serve as a measure of workers' general 'sharpness' levels. It has been used in this way to research the effects of shift working, for example with air traffic controllers[5].

As you would expect, a fatigued operator has a lower fusion frequency. To detect the change, portable light-flashing devices, sometimes referred to as fatigue meters, have



The maximum rate of flicker that a person can perceive relates to how tired that person is. This rate is affected by age and such things as whether you're introvert or extrovert. It is even affected by whether or not you've been drinking coffee. Peter Naish describes a meter that measures the 'critical flicker fusion frequency', and describes how to get the most from it.

Dr Peter L. N. Naish, B.Sc., D.Phil., C.Psychol. is with the Department of Psychology ot the The Open University in Milton Keynes.

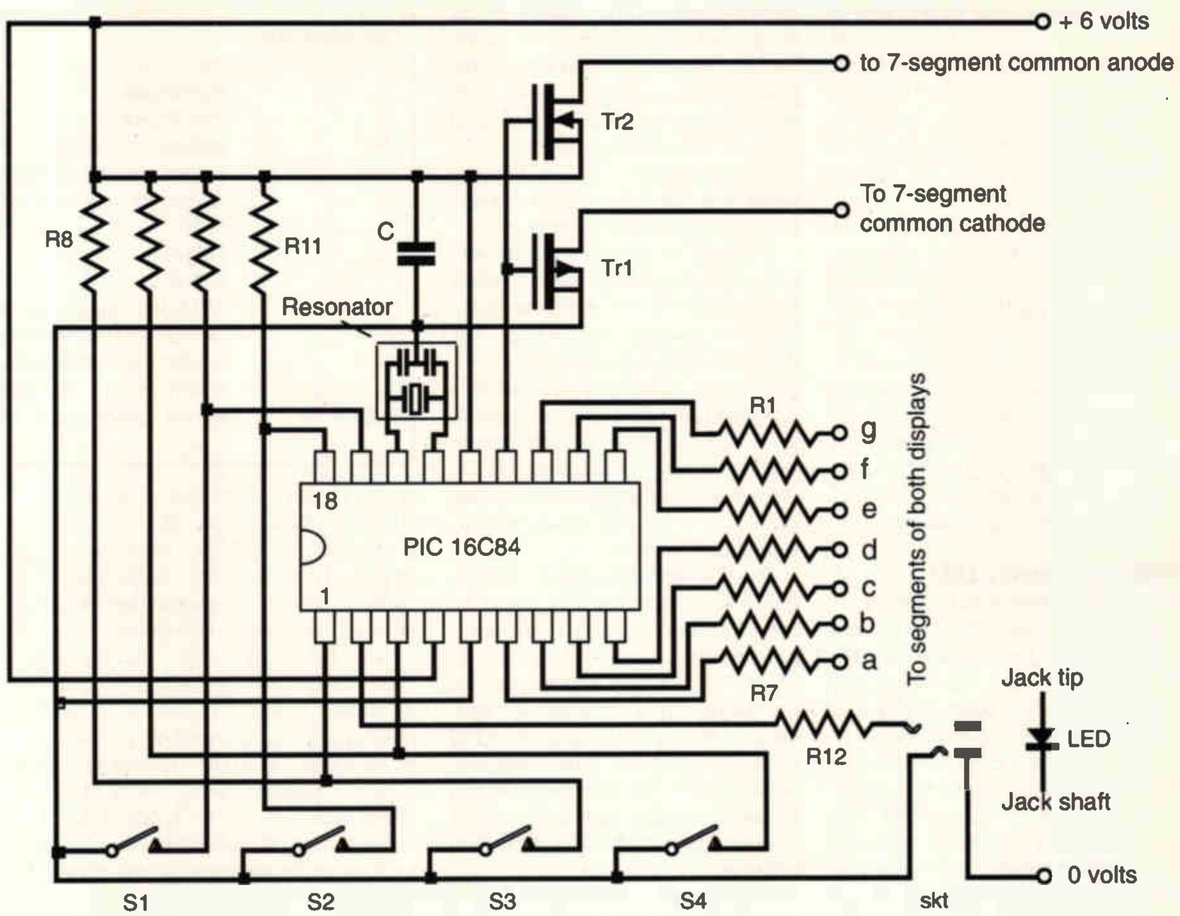


Fig. 1. Complete circuit of the flicker fusion meter, apart from the display. All 13 i/o lines of the PIC16C84 are used. The controller can be set to work with a crystal, resonator or RC network, depending on the accuracy you need. In this application, a ceramic resonator is the best compromise.

been designed to measure CFF in the workplace[6].

This measurement is particularly sensitive to fatigue in the visual system. A comparison of manual workers and vdu operators showed that it was only the latter whose CFF dropped significantly over a working day[7].

If a worker's fusion frequency does fall off appreciably over the day, it is likely that he/she will be feeling weary, will not be working as effectively, and may be more accident prone. The situation can be ameliorated if the worker takes adequate breaks or changes the task from time to time.

A number of different types of flicker-fusion meter have been used; the one to be described here, like that in reference 6, employs a flashing LED, but has more flexible control of the flash rate than the referenced device. It is based on a PIC microcontroller.

Circuit details

Shown in **Fig. 1**, the circuit uses a PIC 16C84, although the newer 16F84 could equally have been used. The device has thirteen i/o lines, all of which are used in this design. It also has an on-chip timer, which is used in setting the flash rate.

Frequency is displayed on two seven-segment LED devices, which, to conserve the battery, are normally turned off. Four push buttons are used. Respectively, these permit stepwise increases or decreases in frequency, cause a steady ramping (up or down) of flash rate, or 'freeze' the LED in the on state (without visible flashing).

Uses of these functions will be described later.

Two ports represent the PIC's i/o lines internally; Port A addresses five lines while Port B addresses the remaining eight. Full details of the PIC can be found on the data sheet, which may be downloaded in pdf format from the Arizona Microchip website, at http://www.microchip.com.

Input/output lines can be configured individually in software to act as inputs or outputs. When used as inputs, the pins associated with Port B can also be programmed to have internal weak pull-up resistors.

Using the internal pull-ups can lead to a useful reduction in external component count. It was simpler not to take advantage of it in this design though, as all the Port B pins are configured as outputs, to drive two 7-segment displays.

Inside the PIC, this port is represented as a single eight-bit register, so writing an appropriate bit pattern to this register drives corresponding pins high or low, turning on and off the various segments of the displays.

Bits 0 through 6 are used for segment control. Bit 7, on pin 13, permits a multiplexing action, so that the seven pins can drive fourteen segments in total, via current limiting resistors $R_{1\text{-}7}$. This is achieved by using one common anode display, D_1 (for the 'tens') and one common cathode, D_2 ('units').

The common pins are respectively taken to positive and ground via n- and p-channel field effect transistors T_1 and T_2. These transistors are driven by the eighth pin of Port B, which when high turns on one, and when low turns on the other.

By alternating rapidly between the two displays, at a rate greater than twice the flicker fusion frequency, the tens and units of the frequency display appear to be on simultaneously and steadily.

The bit patterns required to turn on the appropriate combinations of display segments are stored in a look-up table in the PIC. So, to display the digit 8, which has all seven segments on, the bit pattern is such as to switch every pin high, i.e. it is seven 1s, or 7F in the hexadecimal notation of the PIC, i.e. $7F_{16}$.

To produce an 8, the pattern sent to the port is 01111111, which you will notice leaves bit 7 off, i.e. at zero. This is the multiplexing bit. When the corresponding pin is low, as in this example, T_2 is turned on, which in this case means that 8 would be displayed in the units.

If the same bit pattern is inverted before sending to the port the result will be 801_{16}, or 10000000_2. With the multiplex bit high this turns on T_1. By taking every cathode of D_1 low – i.e. with bits 0 through 6 all zeros – all its segments are lit. Again this displays an 8, this time in the 'tens' position.

In this way, the same bit pattern can be used for a given digit. Whether or not it is inverted determines whether the tens or units display shows it.

Fusion frequency LED

With the Port B pins fully utilised for the frequency read out, one pin from Port A has to be used to drive the flashing LED that establishes the fusion frequency. The remaining four pins of this port are configured as inputs. But since the port does not have an internal pull-up facility, external resistors $R_{8\text{-}11}$ are required. The pins therefore normally sit high, and are pulled low when a corresponding single pole, normally open, push button is depressed.

By polling this port, button presses are detected and acted on. Momentary closure of S_1 or S_3 respectively increments or decrements the flash frequency by 1Hz.

Holding S_4 down causes the frequency gradually to ramp up or down. This is a useful facility, when trying quickly to get to a testee's fusion frequency. Once he/she says that the light seems continuous the one-step up and down buttons can be used to make finer adjustments.

For a final judgement, it is useful to be able to make an A/B comparison. For this purpose, freeze switch S_2 has been added. Holding this down stops visible flashing of the LED, while releasing S_2 returns it to the former rate. If the viewer can see the difference between these, the CFF point has not been reached.

Invisible flashing

Notice that the freeze option stops *visible* flashing; in fact this is achieved by turning on and off at approximately 1kHz, which is far above the CFF. The reason for adopting this approach is that the normal flashing is via a square wave, with 50% duty cycle. If the 'frozen' LED were in fact permanently on – i.e. 100% of the time – it would appear twice as bright. This would confuse the A/B comparison process.

Switch S_2 has a second role of turning on the frequency display; while the LED is in freeze mode the seven-segment displays are active. The default ramping direction is up – i.e. the rate gets faster while the button is depressed. But the direction depends upon which of the single-step buttons was previously depressed. If S_3 (down) is used before ramping, then the direction will be down, until S_1 (up) is next pressed.

The flashing LED, driven via current-limiting resistor, R_{12}, could be mounted directly on the board carrying the other components. This is effectively the approach adopted for the fatigue meter described in reference 6. However, leaving the LED exposed in this way leaves measurements open to the effects of ambient lighting, viewing angle and viewing distance; reproducibility is difficult to maintain.

I recommend that the LED is mounted at the base of a tube, about 25cm long and fitted with an eyepiece at the other end **Fig. 2**. This can take the form of a plug, with a small, say 5mm, axial hole, through which the LED is observed.

The viewing end of the tube should be fitted with an eye cup. This cup serves to limit interference from extraneous light and can be cut from foam rubber.

Paint the inside of the tube matt black to reduce reflections which would tend to enter the eye more obliquely and stimulate the rods.

In the prototype, the tube-mounted LED is connected to a flexible lead, so that the viewer can hold it at a comfortable distance from the control board. The lead is terminated with a 3.5mm jack plug; on the board a corresponding stereo socket is used. By using a stereo component the plug can act as a power switch, by shorting the 'spare' contact to ground.

The PIC can be configured to be externally clocked, or to use a crystal or *RC* oscillator. In this application the crystal option was chosen for greater stability, although rather than a crystal, I opted for a 4MHz ceramic resonator. This particular device has built-in capacitors, eliminating the need for additional components.

Four AA cells power the unit, capacitor *C* reducing transients from the rapidly switched T_1 and T_2. In the prototype, the components are mounted on a pcb, cut to the same dimensions as a four-cell battery holder. The holder attaches to the track side of the pcb and, as explained above, power is switched on by inserting the jack plug of the flashing LED.

Power consumption is low, being less than 5mA during flashing, and 14mA while displaying the frequency.

Programming the PIC

List 1 is a fully-commented program listing so only the salient details need be given here.

Within the PIC are configuration bits. These are set as required during programming. This aspect is not shown in the listing, since the compiling-programming software available from Arizona Microchip – and freely downloaded from their web site – offers the user the opportunity to select required parameters at the time of 'blowing' the chip.

Two parameters are of interest here: the clock oscillator and the watchdog timer. As already stated, the oscillator is configured to be crystal controlled. The watchdog timer is based around a separate, internal *RC* oscillator, which increments an 8-bit register. When the register overflows the device is reset.

To prevent the program from resetting it is necessary to include com-

Components

Microcontroller	PIC 16C84
T_1	ZVN 2106A
T_2	ZVP 2106A
LED	Standard red device
D_1	7.6mm common anode display, e.g. 5082-7731
D_2	7.6mm common cathode display, e.g. 5082-7740
$R_{1\text{-}7,12}$	560Ω
$R_{8\text{-}11}$	100kΩ
C	680nF
Resonator	4MHz with integral capacitors, CST4.00MGW
$S_{1\text{-}4}$	SPNO push button switches, Farnell code: 535-916
Skt	3.5mm stereo jack socket, RS code: 476-328
Jack plug	3.5mm mono, or stereo with poles shorted
Battery box	4 x AA, pcb mounting, Maplin code: CL19V

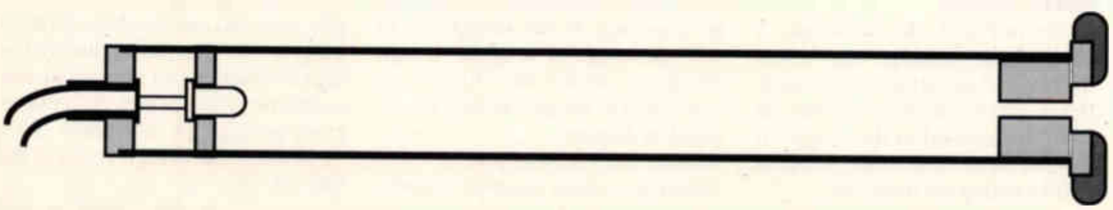


Fig. 2. Mounting the flicker LED in a tube in this way helps eliminate reading errors due to ambient light and movement.

A dim view is a slow view

The separation between the two eyes affords them slightly different views of the same scene. The size of this binocular disparity depends on the distance of a perceived object, and is used by the brain to make distance judgements.

In effect, the eyes are used to triangulate. Each reveals the direction in which an object lies; where the two direction lines intersect the object is to be found.

If an object is moving, its perceived location will be slightly outdated, due to the finite time taken by the visual system to respond. The delay is greater when the light is less bright.

This delay results in mis-triangulation, when one eye receives a dimmer view than the other – an effect that can be demonstrated by viewing a swinging pendulum through half a pair of sunglasses.

Make a long pendulum, for example by pinning a thread to the ceiling and tying on an object near to the floor. Set the pendulum swinging, then stand back some distance, say 2m.

The direction of swing should be across the field of view, not towards and away from the viewer; be sure that the swing is in a straight line. While observing with both eyes, cover one eye with a sunglass lens.

The pendulum weight should no longer seem to swing in a straight line, but in an ellipse, closer to the viewer when moving in one direction, then further away on the return swing.

This phenomenon's mechanism is illustrated in the diagram, which assumes that the sunglass filter is placed over the right eye.

When the pendulum weight moves from left to right, diagram i, the 'bright' left eye perceives it at point **a**. But the slow-responding right eye represents the weight as being further back along the path, at position **b**.

The two direction lines cross at **c**, which is where the brain decides the object actually is; the position is in front of the true line of swing. When movement is in the opposite direction, diagram ii, the corresponding positions become **a'**, **b'** and **c'**, so that the swinging weight is perceived to be behind the actual position.

When the pendulum is moving fastest, i.e. at the mid-point of its swing, the discrepancy is greatest. At either extreme the pendulum momentarily comes to rest, the slow eye catches up, and the weight is perceived in its true position. Consequently, the entire cycle appears to trace an ellipse, which is widest at the pendulum's mid-point.

It is interesting to note that this phenomenon was first predicted by a man who never witnessed it, as he was blind in one eye. In his honour it bears his name: the Pulfrich pendulum.

Estimating the eye's speed

It is possible to estimate the relative slowness of the darker eye.

Stand an object under the pendulum weight, high enough that the pendulum only just clears it at the lowest point of the swing. This object acts as a pointer to indicate the pendulum position; it will facilitate adjustments if it is attached to a long stick that can be reached from the viewing position.

With the pendulum seeming to make its elliptical path, move the pointer to and fro, until it seems to be immediately below the pendulum when at its furthest point – either in front or behind – from the true line.

Measure the magnitude of this displacement – shown as d on the diagram. It is also necessary to measure the viewing distance from the pendulum, *D*, and the separation of the eyes, s. By similar triangles, you will find that the distance of point **b** behind **a** – or **b'** behind **a'** – is given by,

$$s \times \frac{d}{D} \qquad (1)$$

Strictly, *D* should be replaced by *D–d*, for the case where the displacement is in front. But if the viewing distance is large compared with the size of displacement, then the approximation is reasonable.

Calculating temporal lag

From the distance of **b** behind **a** it is possible to calculate the temporal lag, provided the velocity of the pendulum weight is known, at the centre of its swing.

To calculate the velocity it is necessary to measure the length, *L*, of the pendulum thread, and the half-amplitude, *A*, of its swing. Length *A* is the distance from the rest position of the pendulum to the widest point of its swing, making sure that this was the width of swing used when the viewing measurements were taken.

From the equations of simple harmonic motion, to a close approximation the velocity is,

$$A \times \sqrt{\frac{g}{L}} \qquad (2)$$

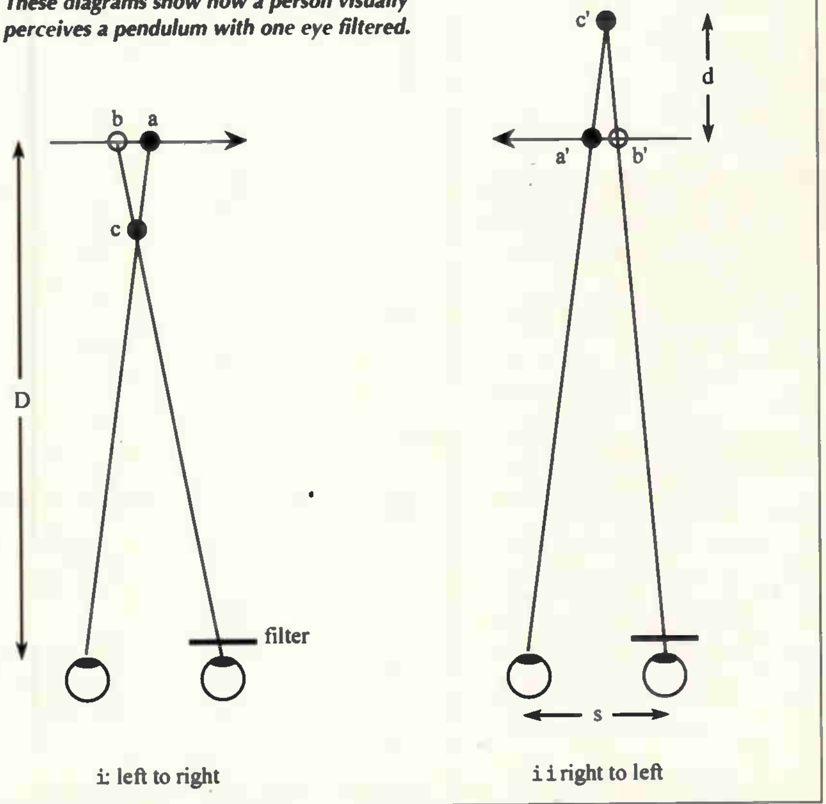
Here, g is the acceleration due to gravity. If the measurements have been made in centimetres, then the appropriate value for g is 981 cm s^{-2}.

Using the familiar expression distance = speed x time, equations 1 and 2 can be combined to provide an estimate of the difference in response times, Δt, between the eyes.

$$\Delta t = \frac{s}{A} \times \frac{d}{D} \times \sqrt{\frac{L}{g}} \qquad (3)$$

The value found is likely to be of the order of 10ms.

These diagrams show how a person visually perceives a pendulum with one eye filtered.



mands in the code that regularly clear the watchdog timer. The watchdog facility can be a useful means of automatically resetting the device, should it become stuck in a loop. It is not used in this application though, so the configuration is set to disable the watchdog timer.

Initialisation of the PIC takes place in the first part of the program, setting i/o pins to the desired input or output. It also assigns a prescaler to the watchdog timer. This assignment is concerned with the flash rate timing, which uses an on-chip timer.

The flash-rate timer is another 8-bit counter that can be incremented by an external clock, or by the internal system clock. In this design the system clock is used.

The clock runs at a quarter of the oscillator rate, so with the 4MHz resonator this design clocks at 1MHz. For some applications, such a rapid tick rate may be too fast for the timer, so the PIC offers the option of prescaling the tick rate by any power of 2, up to 2^8.

Alternatively, this same prescaler can be assigned to the watchdog timer, so that it does not have to be cleared as frequently: the prescaler has to be assigned to one or other timer.

For timing the flicker rate, this program uses the 1μs tick length, so to avoid lengthening this period the prescaler is assigned to the watchdog timer. Although the watchdog is disabled in the design, it is legitimate to make this assignment.

The flashing LED turns on and off with a 50% duty cycle, the duration of each half being determined by waiting an appropriate number of timer cycles. How many timer cycles are required is determined by the selected flicker frequency, which is proportional to the reciprocal of the frequency.

Rather than implement a software calculation to determine the necessary timer delay, I used a look-up table. The subroutine RATEFIX incorporates this table. Instead of the usual, single return-from-subroutine instruction at the end, this routine has multiple return instructions, in fact as many as there are elements in the table.

The 'returns' are also a little more complex than normal; the mnemonic RETLW represents 'return with literal (i.e. a number) in the W (working) register'. Each RETLW instruction is followed by a number, in hexadecimal. It is this value that is contained in the W register on returning from the routine.

In its second line, the routine adds a value, *n* say, representing the

List 1. Program listing for the flicker fusion fatigue meter.

```
; Hexadecimal values for variable names
PortA        equu    0x05          ;five bits wide - LED and control switches
PortB        equ     0x06          ;eight bits for mux 7-seg F readout
Rate         equ     0x0C          ;Register for frequency
timeCount    equ     0x0D          ;reg for no. of cycles in delay
HiDig        equ     0x0E          ;reg for bit pattern for freq readout - hi
LoDig        equ     0x0F          ;reg for lo digit 7-segment bit pattern
DecRate      equ     0x10          ;reg for bcd of rate
GPReg        equ     0x11          ;general purpose register
TenReg       equ     0x12          ;reg used in hex to dec conversion
SpeedReg     equ     0x13          ;reg for delay between ramp-up/down steps
SpeeDirect   equ     0x14          ;reg for ramp direction
RampStart    equ     0x15          ;reg for start value for ramping
TRISA        equ     0x05          ;Page 1, reg 5 is data direction reg ...
TRISB        equ     0x06          ;and same for port B
TIMER        equ     0x01          ;Timer0 address
Opt          equ     0x01          ;Option reg (in page 1)
PCL          equ     0x02          ;program counter (lower bits)
Status       equ     0x03          ;Status reg
Carry        equ     0x00          ;carry bit is bit 0
Z            equ     0x02          ;zero flag
pBit         equ     0x05          ;Page select bit in status reg
Intcon       equ     0x0B          ;Interrupt control reg
GIE          equ     0x07          ;Global interrupt enable bit
TF           equ     0x02          ;Timer overflow flag
; First have to set up IO ports etc, for when initially powered up
             ORG     0              ;Start address
SETUP        BSF     Status,pBit   ;Get to page 1, for setting port masks
             MOVLW   0x00          ;IO mask for port B.  It's 0000 0000 ...
             MOVWF   TRISB         ;all bits output for 7-segments
             MOVLW   0x17          ;0001 0111
             MOVWF   TRISA         ;One output for LED, rest input switches
             CLRF    Opt            ;Zero OPTION, then set desired bits
             BSF     Opt,7         ;Setting bit disables pull-ups
             BSF     Opt,3         ;prescaler to WDT, so no dividing
             BCF     Status,pBit   ;Back to page 0
             BCF     Intcon,GIE    ;Disable global interrupts
             MOVLW   0xD0          ;208 dec ...
             MOVWF   SpeedReg      ;... so takes 48 steps before ramping
             MOVWF   RampStart     ;Put value here too
             MOVLW   0x01          ;+1 to add from ...
             MOVWF   SpeeDirect    ;... up/down direction reg
             MOVLW   0x0A          ;10 down the list of 10 - 60 ...
             MOVWF   Rate           ;... to start at 20 Hz
             CLRF    PortB          ;turn off 7-segment displays at start
             CALL    BUTTONS       ;button not pressed, but need rate etc
; Now everything in place, go into main monitoring loop
MAIN         MOVLW   0x07          ;0111 to mask off main button bits
             ANDWF   PortA,0       ;get button state
             XORLW   0x07          ;should leave zero if no button down
             BTFSS   Status,Z      ;if zero get on with displaying
             CALL    BUTTONS       ;non-zero;see what's pressed
             MOVF    timeCount,0   ;get current delay setting
             MOVWF   GPReg         ;put in general counter
             MOVLW   0x08          ;1000 bit pattern
             XORWF   PortA         ;toggle bit 3 for flashing LED
             BTFSS   PortA,4       ;will be clear if speed-up button depressed
             CALL    SPEEDUP       ;go and deal with speed-up
TIMING       MOVLW   0x3E          ;62 dec, which is ~ 190 off a roll-over
             MOVWF   TIMER         ;start timer at preset value
             BCF     Intcon,TF     ;clear timeout flag
WAIT         BTFSS   Intcon,TF     ;time out yet?
             GOTO    WAIT          ;keep showing frequency
DOWNTIM      DECFSZ  GPReg         ;go through the delay steps
             GOTO    TIMING        ;off for next step
             GOTO    MAIN          ;all steps done: go toggle LED and repeat
; Next section used when ramp button depressed
SPEEDUP      BCF     Status,Z      ;make sure zero flag is clear
             MOVF    SpeeDirect,0  ;get either +1 or -1
             ADDWF   SpeedReg      ;increment or decrement
             BTFSS   Status,Z      ;result a zero?
             RETURN                 ;if not just return
             ADDWF   Rate          ;add ±1 to inc/dec rate
             MOVF    RampStart,0   ;start value for ramp; don't wait 256 steps
             MOVWF   SpeedReg      ;to give 48 delay steps in either direction
             CALL    OVERRUN       ;make sure rate is in usable range
             CALL    RATEFIX       ;get required time delay for rate ...
             MOVWF   timeCount     ;... and store it
             RETURN                 ;now flash at new rate
; Following handles button presses
```

Continued over page...

desired flicker frequency, to the PIC's program counter. The effect is to make the program jump forward by *n* instructions, to encounter one of the RETLWs. The returned number is the timer delay appropriate to the frequency.

A similar principle is used in the subroutine PATTERN, which uses a value between 0 and 9 to return with a corresponding 7-segment bit pattern, to send to Port B.

Implementing the design

Layout is not critical, but at the breadboard stage the capacitor *C* was found to be necessary, to avoid spurious frequency jumps and unwanted device resets.

As I explained earlier, it is preferable to mount the flashing LED in a viewing tube, but adequate results are likely to be obtained if the LED is mounted directly on the main circuit board. Notice, though, that the frequency readout would then also be in view of the person being tested.

It is likely that knowledge of their CFF would influence some people's judgments – attempting to 'beat' their last score for example. For this reason, I recommend that the read-out be covered during use, if a board-mounted LED is used. In this configuration, without the jack socket, an on-off power switch would be required.

The jack socket, switches and battery holder stipulated in the component list fit my pcb design.

Evaluating and using the meter

On connecting power, it will be immediately apparent whether the device is functioning, with LED flashing and frequency displayed when the freeze switch is closed.

The quickest way of making a measurement is to use the ramp-up switch, stopping when the flicker is no longer perceptible. A better result will be obtained by finding a mean; after ramping up, take the flicker rate well above the CFF, then ramp down until the flicker is just perceptible.

An average of the two scores will be a reasonable estimate of the CFF. A more accurate result may be achieved by following ramping with finer adjustments, by means of the up and down switches.

Even this technique can be influenced by the testee's judgement criteria; for example a tendency to say 'Yes' to the slightest glimmer of a flicker will lead to a higher CFF than a strategy of saying 'No' to anything but the clearest blinking.

Eliminating false readings

If a truly criterion-free measure is required, and speed of measurement-taking is not important, then the two-interval forced-choice, or 2IFC, method should be used. To carry out the 2IFC procedure A/B comparisons are conducted, using the freeze button; it is important that the testee cannot see whether the button is depressed or not.

Pairs of presentations are given, one 'frozen' the other flashing. The testee should be asked to look away from the LED during the transition between switch positions; he/she looks only when the flash/no-flash conditions are established, and with a gap of a few seconds between the two.

In this case, the testee is *required* to state – even if unsure – which of the two intervals was flashing. The operator should arrange that sometimes it is the first of the pair that flashes, sometimes the second. If the viewer's hit rate is only 50%, then he/she must be guessing, so

```
BUTTONS     BTFSC   PortA,0       ;up button - will use pull-ups
            GOTO    TRY2          ;if up, try other button
            INCF    Rate          ;go faster
            MOVLW   0x01          ;+1
            MOVWF   SpeeDirect    ;for adding when speed ramping
            MOVLW   0xD0          ;208 ...
            MOVWF   RampStart     ;... 48 off roll-over
TRY2        BTFSC   PortA,2       ;down button
            GOTO    GETON         ;if not down get on with setting params
            DECF    Rate          ;go slower
            MOVLW   0xFF          ; equiv to -1
            MOVWF   SpeeDirect    ;for adding when ramping speed down
            MOVLW   0x30          ;48
            MOVWF   RampStart     ;decrement from 48 when ramping down
GETON       CALL    OVERRUN       ;check rate still in range
            CALL    DECIM         ;go and decimalise rate
            CALL    SEGPAT        ;use dec values to get 7-seg patterns
            CALL    RATEFIX       ;go and get delay for current frequency
            MOVWF   timeCount     ;... and put in counter
            BTFSS   PortA,1       ;freeze button
            CALL    FREEZE        ;make LED look steady
DEBOUNCE    MOVLW   0x07          ;0111 to mask off button bits
            ANDWF   PortA,0       ;is button still down?
            XORLW   0x07          ;should leave zero if no button down
            BTFSS   Status,Z      ;if zero get on with displaying
            GOTO    DEBOUNCE      ;wait until button up
            RETURN                 ;all done
; Next bit flashes v. fast, to look permanently on
FREEZE      MOVLW   0x08          ;1000
            XORWF   PortA         ;toggle LED
            MOVF    LoDig,0       ;get units value of frequency
            MOVWF   PortB         ;display the value
            BCF     Intcon,TF     ;make sure timer not overflowed
            CLRF    TIMER         ;zero timer, ready to wait ...
UnitCycle   BTFSS   Intcon,TF     ;... 256 µs
            GOTO    UnitCycle     ;keep waiting
            COMF    HiDig,0       ;get 10s frequency value, invert for readout
            MOVWF   PortB         ;display the value
            BCF     Intcon,TF     ;clear it again
            CLRF    TIMER         ;start another ...
TenCycle    BTFSS   Intcon,TF     ;... 256 µs
            GOTO    TenCycle      ;keep waiting
            BTFSS   PortA,1       ;freeze button still down?
            GOTO    FREEZE        ;yes - repeat
            CLRF    PortB         ;turn off 7-segs before leaving
            RETURN                 ;all done
; Routine to stop rate value going beyond length of look-up table
OVERRUN     MOVLW   0x32          ;max value Rate should contain
            SUBWF   Rate,0        ;take from actual value
            BTFSC   Status,Carry  ;will be clear if Rate has not gone over max
            CLRF    Rate          ;if over, wrap round to zero
            RETURN
;           Following is table with Nos. of delay steps to get required rate
RATEFIX     MOVF    Rate,0        ;get the rate (0=10 Hz, 50=60 Hz)
            ADDWF   PCL           ;jump forward in table
            RETLW   0xFF          ;255 dec for 10 Hz
            RETLW   0xE8
            RETLW   0xD5
            RETLW   0xC4
            RETLW   0xB6
            RETLW   0xAA
            RETLW   0x9F
            RETLW   0x96
            RETLW   0x8E
            RETLW   0x86
            RETLW   0x80
            RETLW   0x79
            RETLW   0x74
            RETLW   0x6F
            RETLW   0x6A
            RETLW   0x66
            RETLW   0x62
            RETLW   0x5E
            RETLW   0x5B
            RETLW   0x58
```

the flash rate is above the CFF.

Start with the rate a little below fusion, then gradually increment it, giving several pairs of presentations at each stage. Continue, until the 50% criterion is just reached; this will be the unbiased critical frequency.

Variation in CFF is most easily tested by finding the fusion frequency first without, then while wearing sunglasses. When the testee is wearing sunglasses, the frequency should be lower.

```
              RETLW      0x55
              RETLW      0x52
              RETLW      0x50
              RETLW      0x4D
              RETLW      0x4B
              RETLW      0x49
              RETLW      0x47
              RETLW      0x45
              RETLW      0x43
              RETLW      0x41
              RETLW      0x40
              RETLW      0x3E
              RETLW      0x3D
              RETLW      0x3B
              RETLW      0x3A
              RETLW      0x39
              RETLW      0x37
              RETLW      0x36
              RETLW      0x35
              RETLW      0x34
              RETLW      0x33
              RETLW      0x32
              RETLW      0x31
              RETLW      0x30
              RETLW      0x2F
              RETLW      0x2E
              RETLW      0x2E
              RETLW      0x2D
              RETLW      0x2C
              RETLW      0x2B
              RETLW      0x2A;       60Hz
; Following changes to decimal format - nibbles = tens & units
DECIM         CLRF       DecRate     ;zero current value
              MOVLW      0x0A        ;10 as frequency starts at 10 Hz
              ADDWF      Rate,0      ;get actual frequency ...
              MOVWF      GPReg       ;... and save it
              MOVLW      0x0A        ;a counter for ten operations
              MOVWF      TenReg      ;store in register
DECUP         INCF       DecRate     ;start incrementing the register
              DECFSZ     TenReg      ;do up to ten additions
              GOTO       MAINDEC     ;if not ten yet, nothing to correct
              MOVLW      0x0A        ;if ten, reset counter
              MOVWF      TenReg      ;to count next ten
              MOVLW      0x06        ;add another 6 ...
              ADDWF      DecRate     ;... to rollover to tens nibble
MAINDEC       DECFSZ     GPReg       ;working through entire No
              GOTO       DECUP       ;repeat
              RETURN                  ;all done
;             following gets correct 7-segment patterns from digits
SEGPAT        MOVF       DecRate,0   ;get the BCD value
              ANDLW      0x0F        ;mask off lower nibble
              CALL       PATTERN     ;get the pattern
              MOVWF      LoDig       ;and put in units store
              SWAPF      DecRate,0   ;get other nibble at lower end
              ANDLW      0x0F        ;mask again
              CALL       PATTERN     ;find its 7-seg pattern
              MOVWF      HiDig       ;and store it
              RETURN                  ;all done
PATTERN       ADDWF      PCL         ;jump ahead
              RETLW      0x3F        ;pattern for zero
              RETLW      0x06        ;1
              RETLW      0x5B
              RETLW      0x4F
              RETLW      0x66
              RETLW      0x6D
              RETLW      0x7D
              RETLW      0x07
              RETLW      0x7F
              RETLW      0x6F        ;9
              RETLW      0x40        ;minus
              END
```

The difference in inter-flash intervals represented by the frequency difference may be compared with the timing difference calculated by the method detailed in the panel entitled, 'A dim view is a slow view'. The two are unlikely to be identical, since each depends on brightness levels; however, the values should be of the same order of magnitude.

Testing a range of people should show a tendency for CFF to fall with age and with tiredness. In women, some variation may be found over the menstrual cycle.

In the workplace, the device can be used as a fatigue meter, testing at the start and end of the working day. If a vdu operator shows an appreciable drop in CFF of, say, 10% over the day, then it would be appropriate to examine the working practice and ergonomical layout of the workstation.

Standard health and safety practices should be employed, checking that there is no glare or reflection from monitor screens, that the material displayed is easily legible and that the worker wears appropriate glasses if required. Ambient lighting levels must be adequate.

There is some evidence[8] that the flickering of standard fluorescent lighting can lead to an experience of stress, particularly for those with high CFF thresholds; driving the tubes at high frequency leads to greater comfort.

As a general rule, workers should take regular breaks, which not only relieve the demands of the task, but also give a period when the eyes can take in scenes of different brightness and at different distances. ■

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Optoelectronics spots cancer cells in seconds

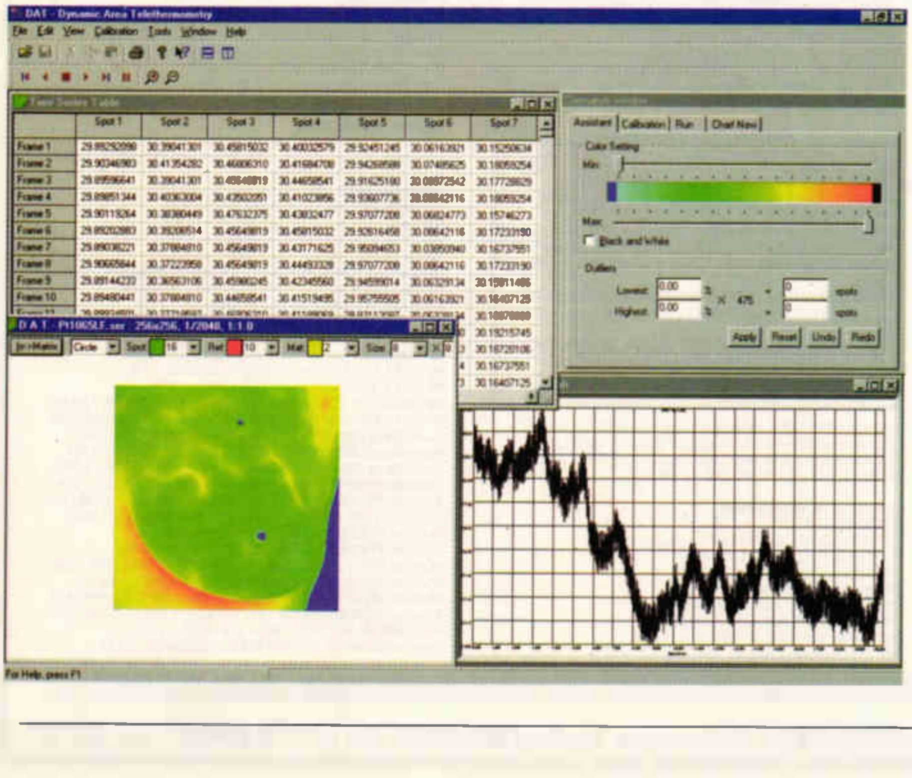
A new technique called 'light-scattering spectroscopy' helps detect pre-cancerous cells in a fraction of a second using just an endoscope, light beam and some DSP. Pete Mitchell reports.

Cancers are much more curable if spotted early, before malignant cells have spread elsewhere. So doctors are always looking for new ways of detecting early-stage tumours or "dysplasias", preferably before they even become visible.

But it's not easy: the usual method of extracting a tissue sample by biopsy and then examining it microscopically is time-consuming, invasive, and sometimes painful.

Now, groups of US researchers have developed optolectronic methods that need just an endoscope and a light beam – no needles or aspirators – and produce results in a fraction of a second.

Both methods are based on a technique called light-scattering spectroscopy. LSS has been used for some years in chemistry to study the size and shape of small spheres such as droplets. More recently, the MIT Laser Biomedical Research Centre, run by physics professor Michael Feld, has been applying it to measure the spheroid nuclei of body cells. The key fact is that cancer precursor cells are different from normal ones – they are closely packed, with unusually large nuclei crammed full of DNA.



In the LSS method, the clinician shines light through the probe onto the patient's tissue, and the probe collects the light that bounces back. The spectrum of this scattered light is slightly different from that of the original beam, in a way that depends on the size and refraction properties of the nuclei. Examining this characteristic 'signature' of the tissue can reveal a dysplasia.

The probe is designed for use on epithelial tissue – the material that lines cavities of the body such as the mouth, bladder wall and colon. The epithelium is often the body's first line of defence against cancer-causing substances, and many tumours originate there – including lung, stomach, breast and cervical cancer. Such cancers are virtually invisible at the early stages, even through an endoscope.

Gastroenterologist Jacques Van Dam is working with Feld to test the LSS probe clinically. "Being able to see the changes while actually there with the patient has not been possible before," said Van Dam. "If these very subtle changes are detected before they become cancerous, we can prevent cancer from forming."

The lab has also developed a more advanced LSS imaging device that scans areas of tissue several centimetres across. This instrument produces a series of light beams formed from white light with coloured filters and a polarizer. An electronic camera records a pair of images at each wavelength of the reflected light in two separate polarizations. The two images are then subtracted, which cuts out scattered background light and leaves behind only images relating to the cell nuclei.

Using digital signal processing, this analysis can be done in a fraction of a second and the results displayed in a way that is easy for the doctor to interpret. "By analyzing the intensity variations in this back-scattered component from colour to colour, the nuclear size and density can be mapped," says Feld.

He predicts that, within two years, these new devices will lead to a new class of endoscopes and other diagnostic instruments that allow physicians to obtain high-resolution images. These easy-to-read images will map out normal, pre-cancerous and cancerous tissue the way a contour map highlights elevations in reds, yellows and greens.

In clinical tests, the probe accurately distinguished dysplasias in the bowel and oesophagus (gullet), where simple endoscopy would not have worked. For the oesophagus, even biopsies are very hard to interpret, and experienced pathologists often disagree on biopsy results.

The stakes are high: if a dysplasia is present, the usual treatment is surgical removal of the whole gullet. Feld believes the LSS method would be a great improvement.

Other people are exploring infra-red technology, created by the military and aerospace electronics sectors, to detect more advanced tumours non-invasively. One such instrument is BioScan, developed by the New York company OmniCorder.

BioScan senses and records heat patterns radiated by the human body. Body heat is closely associated with blood flow (technically called perfusion), and tumours are notoriously good at keeping themselves well supplied with blood.

So mapping 'hot spots' on, say, a breast, can often reveal where a solid tumour is forming. Traditional mammograms (breast X-rays) often give uncertain results because the transparency of breast tissue to X-rays varies sharply between women, depending on their age and physiological state. Mammograms are also very uncomfortable and pose the usual risk from ionizing radiation.

But the heat differences are so small that only the most sensitive of IR detectors will do. Omnicorder uses QWIP (quantum well infra-red photodetection) technology developed by NASA and Lockheed Martin, to which it has obtained the biomedical rights. The technology relies on hybrid circuits cooled to very low temperatures to reduce thermal noise.

In the BioScan, a digital infrared camera containing the QWIP sensor measures the very small changes in heat energy caused by perfusion changes. The camera is sensitive to temperature changes of less than 0.015°C and has a speed of more than 200 frames per second.

This data is analysed using a powerful workstation and a proprietary technique called dynamic area telethermometry, invented by Dr Michael Anbar, the founding scientist of OmniCorder.

This generates an image of the target area and points out the presence and size of a tumour.

As well as screening, this can also check the effectiveness of radiotherapy or chemotherapy on a known tumour. Some new therapies also attempt to interfere with the flow of blood to tumours, so BioScan can give feedback on how well this is working.

The BioScan underwent extensive testing at the Dana-Farber cancer institute in Boston, and was licensed for sale in the US in December. OmniCorder says it is "inundated" with orders for the device.

The company has now contracted AEG Infrarot-Module to manufacture its QWIP cameras modules in volume, using a technology developed by the German company in conjunction with the Fraunhofer Institute in Freiburg. ■

Evaluate capacitors for SMPS designs

When choosing a capacitor for a linear power supply, you don't need to know much more than capacitance value, working voltage and ripple rating. For modern switch-mode power supplies though, high frequencies, complex waveforms and the demand for ever increasing efficiency make the choice significantly more difficult. Cyril Bateman explains what the problems are, and how to sort them out.

The ever-increasing demand for small, lightweight, efficient equipment has resulted in an explosion in the number and variety of switched-mode power supply integrated circuits. Many of the latest designs are available only in minute surface mount packages, encouraging designers to use physically small capacitors and inductors.

Increase in switching frequency can reduce the theoretical capacitance value needed – hence the physical size of the component. Since almost all practical capacitors differ considerably from these theoretical ideals, trading off size for frequency introduces many pitfalls for the unwary designer.

For a given capacitance and working voltage, the smallest physical size usually results from using either high-K ceramic chips, or tantalum or aluminium electrolytic capacitors. These possess three undesirable attributes[1].

- As frequency increases, their apparent, measurable capacitance reduces. It can become much smaller than the marked, low frequency, nominal value.
- For a given *CV* product, case size reduction invariably increases the ESR of the capacitor.
- Measurable self-inductance.

At high frequencies, the equivalent series resistance of an electrolytic capacitor can exceed its capacitive reactance. When this happens, the capacitor's measured phase angles become small – a few degrees only. The measured impedance curve then appears flat-bottomed over a wide frequency band.

At some frequency, the capacitive and inductive reactances are equal and cancel. Measured impedance is then equal to the capacitor's ESR. Above resonance, the capacitor's measured impedance increases with frequency, **Fig. 1**.

Some manufacturers provide nominal impedance or ESR values for their ranges, usually at 100kHz and at room temperature. High frequency capacitance and inductance values are rarely stated though.

Using a suitable, variable-frequency *LCR* meter, these parameter changes with frequency can be accurately measured. Since even a used meter can be extremely expensive, one may not be available.

These measurements are also possible using simple methods and low-cost laboratory instruments. Taken with with care, such measurements can provide useful accuracy.

Impedance is traditionally measured by passing a known AC current through the capacitor, and measuring the resulting voltage drop. Impedance is equal to voltage drop divided by the through current[2].

Measurement basics

At switching power supply frequencies, the capacitor becomes a very low impedance. Four-wire test lead connections are essential. For consistency and accuracy the capacitor should be mounted in a suitable test jig.

When an AC current is passed through the ideal, or perfect capacitor, having neither inductance nor resistance, the voltage waveform lags that of the current by 90°. The capacitor produces an impedance with a phase angle of –90°.

At any one frequency, a practical capacitor can be represented by a series combination of inductance, capacitance and resistance. These combine to produce an impedance with a much reduced phase angle. Depending on frequency, phase can be either positive or negative.

An *LCR* meter converts this measured impedance and phase angle into two components only,

$$|Z|\angle\theta = R \pm jX$$

representing ESR (resistance) and reactance.

If the measured phase angle is negative, the meter calculates a capacitance value, if positive, an inductance value. Because only a single frequency has been measured, the meter cannot segregate this reactance into its inductive and capacitive components.

$$|Z|\angle\theta = R \pm jX = \sqrt{R^2 + X^2}$$

Here, *R* is capacitor ESR at the measured frequency while *X* is reactance at that frequency.

With practical capacitors, this net reactance has both a capacitive and inductive component. In principle, provided impedance and phase angle are measured at no fewer than two frequencies, it is feasible to extract these components from the values of *X* measured. There's more on this in the panel entitled, 'Three-component modelling'.

Impedance

Generating a known alternating current at various frequencies can be difficult.

One common method is to apply a constant voltage via a known current-limiting resistor. This resistance's value must be much greater than the impedance of the capacitor being measured, so that any change in current due to change in the capacitor's impedance can be ignored.

Measurements of electrolytic capacitors can be made using the internal source resistance of a signal generator to limit current. Accuracy of this method can be poor though, since the current can vary with frequency. Also, the voltage drop across the capacitor may be extremely small – at best a few millivolts.

Measuring impedance: Method 1

Accurate impedance measurements can be made, based on the techniques used to measure the insertion loss, or attenuation, of EMC filters. Such a procedure was originally defined in MIL-STD-220.

Using a spectrum analyser and tracking generator, a picture of the capacitor's performance can be obtained – even to very high frequencies. This allows very quick comparisons between capacitor types.

Actual impedance values of the capacitor by frequency can be calculated from these attenuation measurements, as detailed in the panel entitled, 'Implementing Method 1'.

With the capacitor mounted in a suitable test jig, such measurements can be carried out using low-cost laboratory equipment. You need only a 50Ω low distortion signal generator, a high input impedance RF millivoltmeter, a 10dB isolating attenuator and a 50Ω

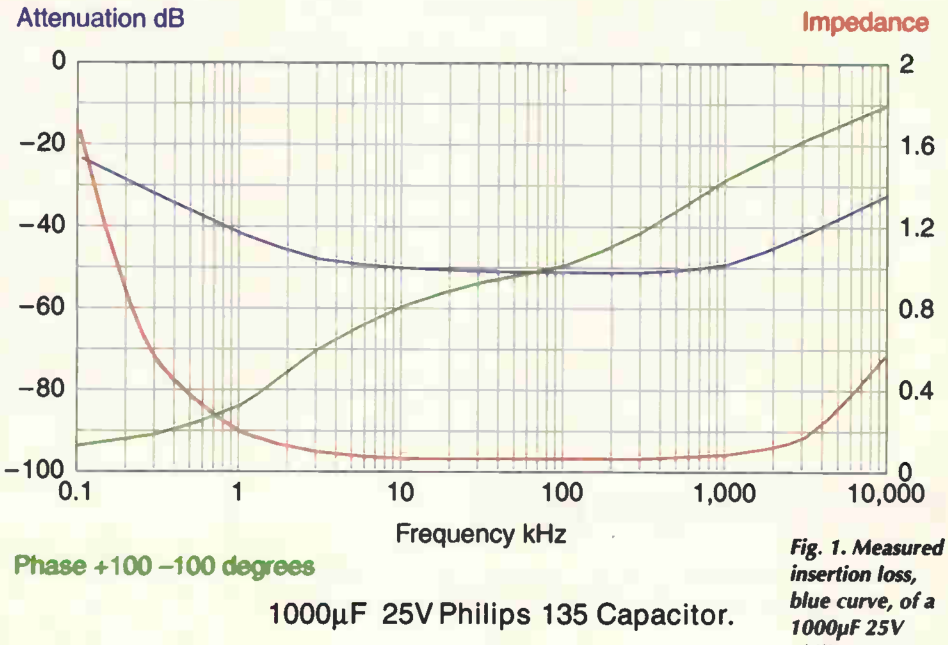


Fig. 1. Measured insertion loss, blue curve, of a 1000µF 25V Philips 135 capacitor using test jig 1, Method 1. Impedance, in red, is calculated using,

$$Z = \frac{25 \times A}{1 - A}$$

where,

$$A = 10^{[illegible]}$$

Self inductance was calculated as 8.75nH, impedance at 100kHz measured 0.06Ω.

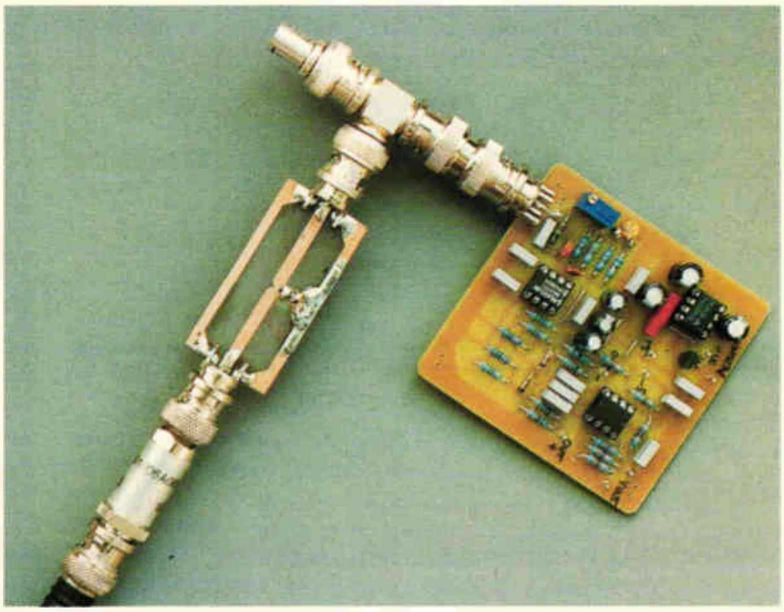


Fig. 2. Impedance measurement method, using test jig 1, Method 1, with a high input-impedance meter. The capacitor lead wires were soldered to the stripline and ground plane. The 10dB attenuator isolating the signal generator from line reflections is connected to one jig connector, the terminating 50Ω load to the other.

non-inductive termination, **Fig. 2**.

The test jig shown, 'test jig 1', has other uses. It was used with a function generator and oscilloscope to provide the measurements shown in Figs 6a), 6b), and 6c) in the June 2000 issue[3].

A true four-terminal measurement method, it can provide extremely accurate measurements of impedances ranging from 5Ω to 500Ω. Accuracy depends on the signal generator source, the terminating load's actual impedances and your ability to measure small voltages.

Lower impedances require very small voltages be measured. Ideally a spectrum analyser or a tuned narrow bandwidth receiver would be used here, to minimise contributions from noise and distortion.

Although the voltage being measured is in fact complex – i.e. a vector having both magnitude and phase – to calculate impedance, you need only measure voltage magnitude using a conventional voltmeter or spectrum analyser.

This method can provide accurate measurements of impedance and phase angle. However using only common laboratory equipment, the small capacitor voltages complicate phase angle measurements. An accurate phase angle reference must also be established, Fig. 1.

Capacitor performance in practice

Capacitance and inductance values, calculated from measured $R \pm jX$ for typical 220μF/10V capacitors. One random sample only of each style was measured.

Unit/parameter	10kHz	30kHz	100kHz	300kHz	1MHz
220μF/10V Philips 037	Aluminium				
IZI Ω	0.824	0.785	0.738	0.706	0.671
ESRΩ	0.818	0.783	0.74	0.71	0.67
Capacitance	129μF	68.4μF	27.1μF	9.1μF	2.0μF
220μF/10V Rubycon YXF	Aluminium				
IZIΩ	0.404	0.372	0.341	0.322	0.308
ESRΩ	0.394	0.372	0.34	0.32	0.32
Capacitance	153μF	94μF	44μF	20.7μF	9.2μF
220μF/10V Rubycon ZL	Aluminium				
IZIΩ	0.122	0.091	0.082	0.076	0.073
ESRΩ	0.10	0.09	0.09	0.08	0.08
Capacitance	187.5μF	169.5μF	122.3μF	104.5μF	1.21nH
220μF/10V Rubycon ZA	Aluminium				
IZIΩ	0.087	0.041	0.030	0.028	0.035
ESRΩ	0.04	0.04	0.03	0.03	0.03
Capacitance	189.8μF	185.4μF	200.9μF	0.69nH	2.39nH
220μF/10V Elna RSH	Aluminium				
IZIΩ	0.313	0.290	0.270	0.254	0.242
ESRΩ	0.31	0.29	0.27	0.26	0.25
Capacitance	171.7μF	119.7μF	55.0μF	27.6μF	19.33μF
220μF/10V AVX TPS	Tantalum				
IZIΩ	0.143	0.089	0.06	0.045	0.040
ESRΩ	0.10	0.08	0.06	0.05	0.05
Capacitance	149.2μF	100.9μF	62.3μF	45.7μF	0.83nH
220μF/10V Sanyo Oscon	Oscon				
IZIΩ	0.077	0.027	0.01	0.01	0.029
ESRΩ	0.02	0.02	0.01	0.01	0.01
Capacitance	201.3μF	192.0μF	230.9μF	2.13nH	4.23nH

These values were derived using my dedicated meter, switched to Method 3 for impedance then Method 2 for ESR and capacitance. Many other suitable capacitors are available, from other stockists. ESR cannot exceed $|Z|$. Where this occurs in the table, it is caused by insufficient resolution in my phase angle measurement.

While the rapid increase in capacitance of the 220μF/10V Rubycon ZA and the 220μF/10V Sanyo Oscon may look odd, this is simply a reflection of the effect self-inductance has on apparent capacitance when approaching series resonance. A notable impedance null is frequently found at resonance with low-loss capacitors. With electrolytics, series resistance usually dominates, so the effect is not there.

Three-component modelling

The results table was calculated using the two-component capacitor model. For best accuracy, a three-component evaluation is needed, as indicated by,

$$|Z| = \sqrt{ESR^2 + (X_C - X_L)^2}$$

Here X_C is capacitive reactance and X_L the inductive reactance at the measured frequency.

Solving for three unknowns requires a minimum of two measurements at differing frequencies. Ideally, a swept measurement at several frequencies is used.

Some recent swept-frequency component analysers, such as the HP4194 and HP4195, are provided with internal software routines, which automatically calculate the three-component model[7].

These evaluate parameters at frequencies where the measured impedance is a factor of √2 smaller and larger than the maximum and minimum values measured[8].

This works well for many stable components, but not for electrolytic capacitors, having ESR and capacitance values that change with frequency. Self inductance for these however *is* relatively constant with frequency.

I prefer to estimate this inductive component by taking a series of impedance measurements at frequencies well above resonance. But why bother?

Taking the Oscon capacitor which resonated at 190kHz as an example, I measured impedance at 1MHz intervals up to 10MHz. Above 2MHz its apparent measured inductance stabilised close to 5nH, its impedance then increasing linearly with frequency.

A capacitor that measures as an inductive impedance can still act as a capacitor to decouple noise or store and discharge energy. It has an inductive behaviour simply because at that frequency its inductive reactance exceeds its capacitive reactance. At high switching frequencies, it will exhibit an inductive overshoot as in Fig. 6a) in the June issue[3].

Having a reasonable estimate for self-inductance, using the above three-component equation can produce a better estimate of the true capacitance value at any frequency.

Even at 100kHz though, it can be difficult to guarantee similar path lengths and phase delays between the reference and measurement channels.

Simplified measurements: Method 2

In Method 1, when measuring low-impedance electrolytic capacitors, there's a wide ratio between the 50Ω source and loads used as references, and the capacitor's impedance. This requires voltage measurements over a wide dynamic range.

A lower-value reference resistor can dramatically reduce these voltage differences, simplifying the measurement and improving accuracy using low-cost instruments.

In this method, the test capacitor is mounted in a 50Ω microstrip line, on a double sided board, in series with the signal generator. The reference resistor forms the ground return.

Complex voltages, $V(1)$ and $V(2)$, at levels suitable for impedance and phase angle measurements, are found on each of the capacitor lead wires, **Fig. 3**.

By these means, a low-cost RF millivoltmeter[4] can be used with a phase meter[5] to characterise the test capacitor. While a number of calculations are needed to convert the measured voltages and phase angle into the required impedance, ESR and capacitance-inductance values, the method is quick and easy to apply – and cheap to carry out. To facilitate these calculations, I use a small program written for my programmable calculator.

In practice, the main difficulty is providing a known value, non-inductive resistor. I needed to measure impedances at 100kHz from 0.01 to 1.0Ω, increasing to 2.0Ω at lower frequencies. A reference resistor around 0.5 to 1.0Ω would be suitable.

Making a non-inductive resistor

Conventional 1.0Ω 1% resistors are readily available, but these usually have a spiral 'cut', used to trim to final value. Combined with the resistor's physical length, this results in sufficient self-inductance that the resistor's impedance is measurably increased at 100kHz – degrading measurements. For accuracy, a spiralled reference resistor should not be used.

Surface-mounted chip resistors offer less inductance for two reasons. Firstly, a straight 'L' cut is often used to trim to value and their physical lengths can be shorter. Even better, certain types are available that are wider than they are long. Sometimes, these are effectively three 1206 resistors in parallel.

A typical 1206 chip has some 1 to 1.5nH inductance, so with three in parallel, this construction provides mini-

Implementing Method 1

For consistent measurements, a jig that maintains a good 50Ω impedance up to and beyond the insertion point of the capacitor being tested is essential. Capacitors should be soldered to this jig using the same lead lengths as you will use in your application, Fig. 2.

I use a 3mm-wide stripline on double-sided FR4 circuit board. The capacitor is connected in shunt with this line, to the ground plane. This ground plane covers the reverse side of this jig, except for a circular area where the capacitor mounts.

The jig is suitable for leaded and surface-mount components. Both sides of the printed board are linked together at the capacitor ground connection using vias and a soldered copper foil wrap.

As with all 50Ω measurement systems, reflections caused by mounting a capacitor across the line must be minimised. A 10dB attenuator should be inserted in the signal-generator cable, as close as possible to the test jig.

If a high-impedance measuring instrument is used, a 50Ω through-terminating resistor, or a terminating 50Ω load and 'T' piece, should be placed at the jig output, Fig. 2.

If a coaxial cable is necessary between the test jig and the measuring instrument, a second 10dB attenuator should be connected to the test jig output. The coaxial cable should be terminated in 50Ω at the measuring instrument.

While each 10dB attenuator reduces the signal level by 10dB, reflected signals are twice attenuated, so are reduced by 20dB.

A measurement of attenuation, either as a voltage ratio or decibels, is taken for each measurement frequency. At each frequency, the 'jig out' voltage, with the measurement test jig replaced by an empty jig, should be noted.

With no other change in set-up, the measurement test jig with capacitor, should then be inserted into position, and the 'jig in' voltage measured, Fig. 1.

$$\text{Attenuation ratio} = \frac{V_{jig(in)}}{V_{jig(out)}} = A$$

$$\text{Attenuation in dB} = 20\log\frac{V_{jig(in)}}{V_{jig(out)}}$$

Convert any decibel readings to attenuation ratio,

$$A = 10^{\frac{dB}{20}}$$

$$\text{Impedance} = \frac{25A}{1-A}$$

This equation holds good provided both source and load impedances are 50Ω. In the above, the figure '25' represents the Thévenin equivalent of the source/load impedances.

This method can provide very accurate results, but because of the wide dynamic range voltage measurements needed, it is

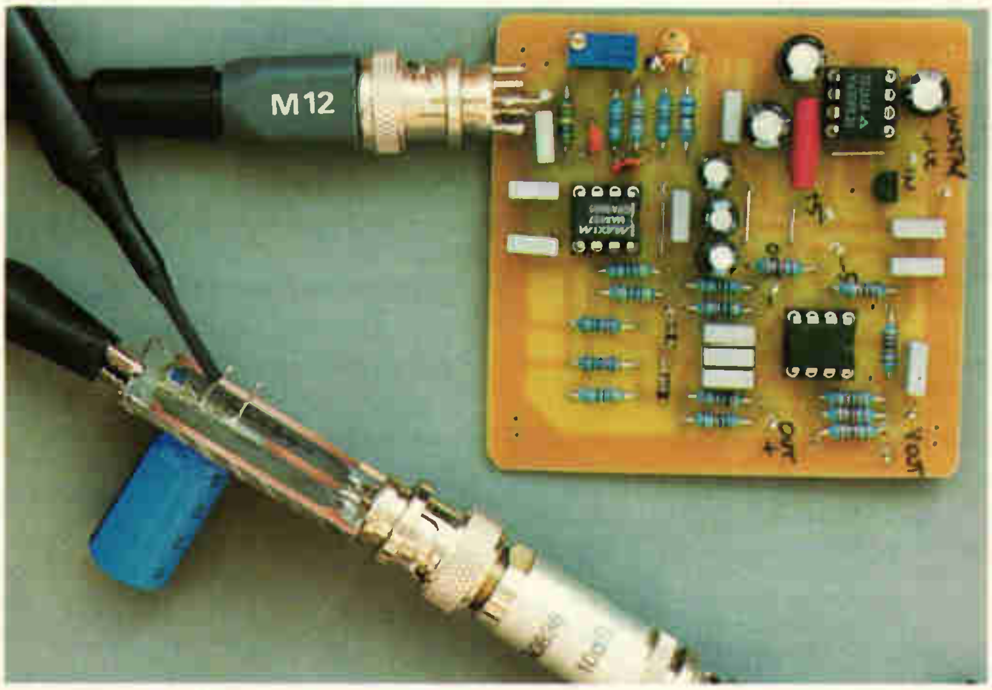


Fig. 3. Using test jig 2, Method 2, to measure impedance, with a high input-impedance meter. My stack of three 1218 surface mount chip resistors, total resistance 0.4989Ω, are visible near the test-probe ground clip. Compared with Fig. 1, the test voltages, VM(1) and VM(2) have larger magnitudes. This facilitates both voltage and phase measurement.

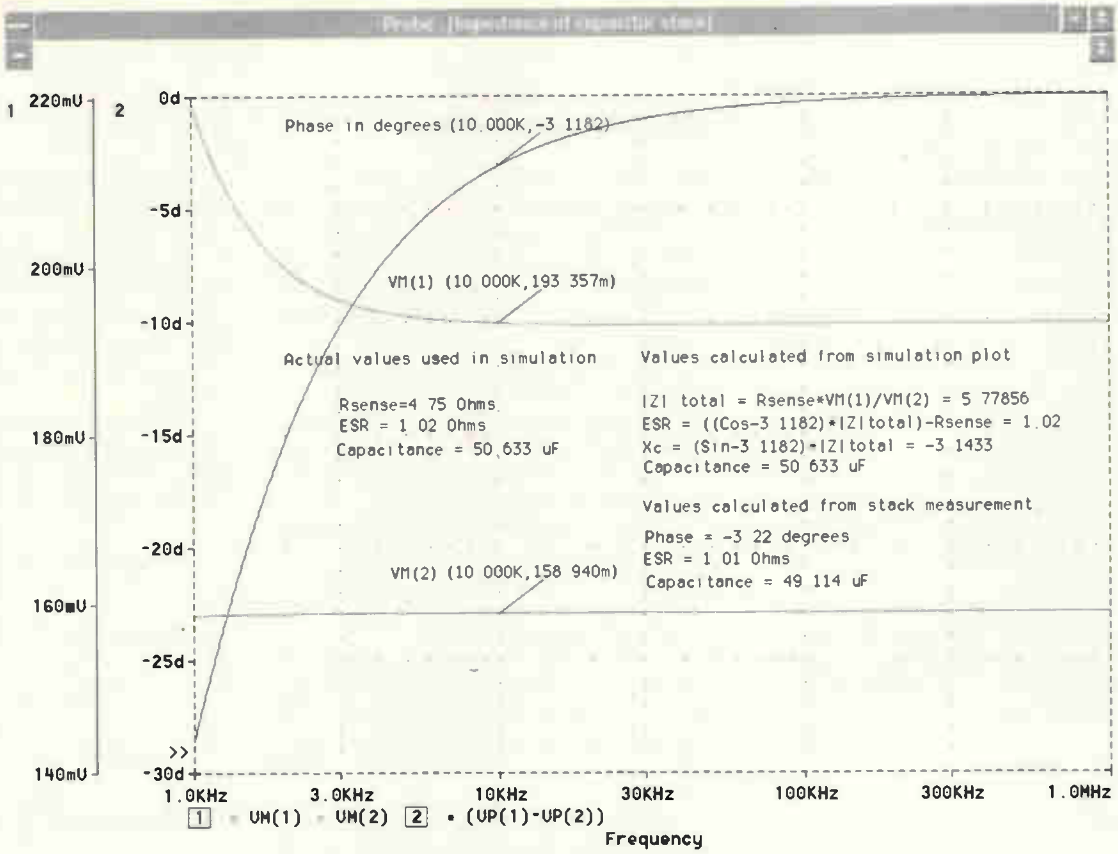


Fig. 4. Simulation used to verify Method 2. A stack of 10µF plastic-film capacitors, each accurately pre-measured at 10kHz, was assembled and measured. These pre-measured values were simulated. ESR and capacitance was calculated from simulated results and confirm the calculation method. Both sets of measured values agreed, with less than 5% error.

mal self inductance[6].

I bought a number of 1.5Ω resistors, Philips type PRC201. These are 1218 size, comprising three 1206 resistors in parallel. Measuring voltage drop while passing a 100mA DC, I was able to select a number of identical sets of three, effectively nine 1206 resistors in parallel. Each set makes a non-inductive, near 0.5Ω value, for my test jigs.

The selected group of three 1218 chip resistors were stacked together then soldered in place on my jig. This assembly can be seen in the photograph, Fig. 3.

To validate my reference resistor, I mounted a conventional 1% 1Ω resistor rated at 0.6W on this jig. At 100mA DC, it measured 0.996Ω. At 1kHz it measured 0.998Ω while at 5MHz it measured 1.007Ω – equivalent to some 5nH inductance. This confirmed that my reference resistors have minimal inductance.

As with all self-built measurement systems, the remaining problem was to test its measurement accuracy when measuring capacitors.

At 10kHz and above, the largest capacitance I can accurately measure on my 0.1% bridge is 11µF. I assembled a stack of five 10µF metallised PET capacitors, which I first carefully measured at 10kHz.

To increase this stack's ESR to represent a typical electrolytic capacitor, I added a 1Ω, 1% series resistor. Using a 1%, 4.7Ω resistor as the reference resistor and Method 2, I measured this stack, then ran a PSpice simulation.

Implementing Method 2

By comparing the voltages at the unknown capacitor, now placed in series with the test signal, with those on a low value, non-inductive, current sensing resistor, the voltage range that needs to be measured, is minimised.

This smaller jig again uses a 3mm wide stripline on double sided FR4 circuit board. The current sense resistor, connected between one capacitor lead and ground, terminates this jig.

The value of the current sensing resistor should approximate the mean impedances to be measured. It must be non-inductive and its true resistance accurately known.

To minimise line reflections to the signal generator, a 10dB attenuator should be inserted in the signal coaxial cable, immediately adjacent to the test jig.

The basis of this measurement is clear from Fig. 3. Two ground-referred voltage measurements at each frequency are needed. These are easily made using a high input-impedance millivoltmeter with a conventional high-impedance oscilloscope probe, contacting each capacitor lead wire in turn, Fig. 3.

Voltage measured at the capacitor lead nearest the signal generator is *V*(1). The voltage measured at the capacitor lead near the current sensing resistor is *V*(2).

Both voltage magnitudes will be similar, simplifying voltage and phase angle measurements.

A measurement of phase angle difference across the capacitor, is also required. A high input-impedance phase meter with two identical oscilloscope probes can be used.

The reference probe connects to the *V*(1) capacitor lead wire while the measurement channel probe connects to the V(2) lead. Both probe earth leads are grounded.

Using normal laboratory instruments, this method can measure impedance and phase angle of a capacitor with good accuracy. This is because relatively high and similar voltages are measured.

Using '$|Z|_{total}$' to describe the combined impedance of the test capacitor and sense resistor,

$$|Z|_{total} = R_{sense}\frac{VM(1)}{VM(2)}$$

where *VM*(1) and *VM*(2) are voltage magnitudes measured at *V*(1) and *V*(2) using a normal voltmeter.

$$ESR = \left(\cos\ phase\ angle \times |Z|_{total}\right) - R_{sense}$$

Here, *phase angle* is *VP*(1)–*VP*(2).

$$X_C = \sin\ phase\ angle \times |Z|_{total}$$

See reference 5 for more on the above equation. If *phase angle* is negative,

$$C = \frac{1}{2\pi f X_C}$$

Or if *phase angle* is positive,

$$L = \frac{X_C}{2\pi f}$$

$$|Z|_{capacitor} = \sqrt{ESR^2 + X_C^2}$$

Voltages *V*(1) and *V*(2) are complex, having both magnitude and phase. But for this method you only need to measure their voltage magnitudes, Fig. 4.

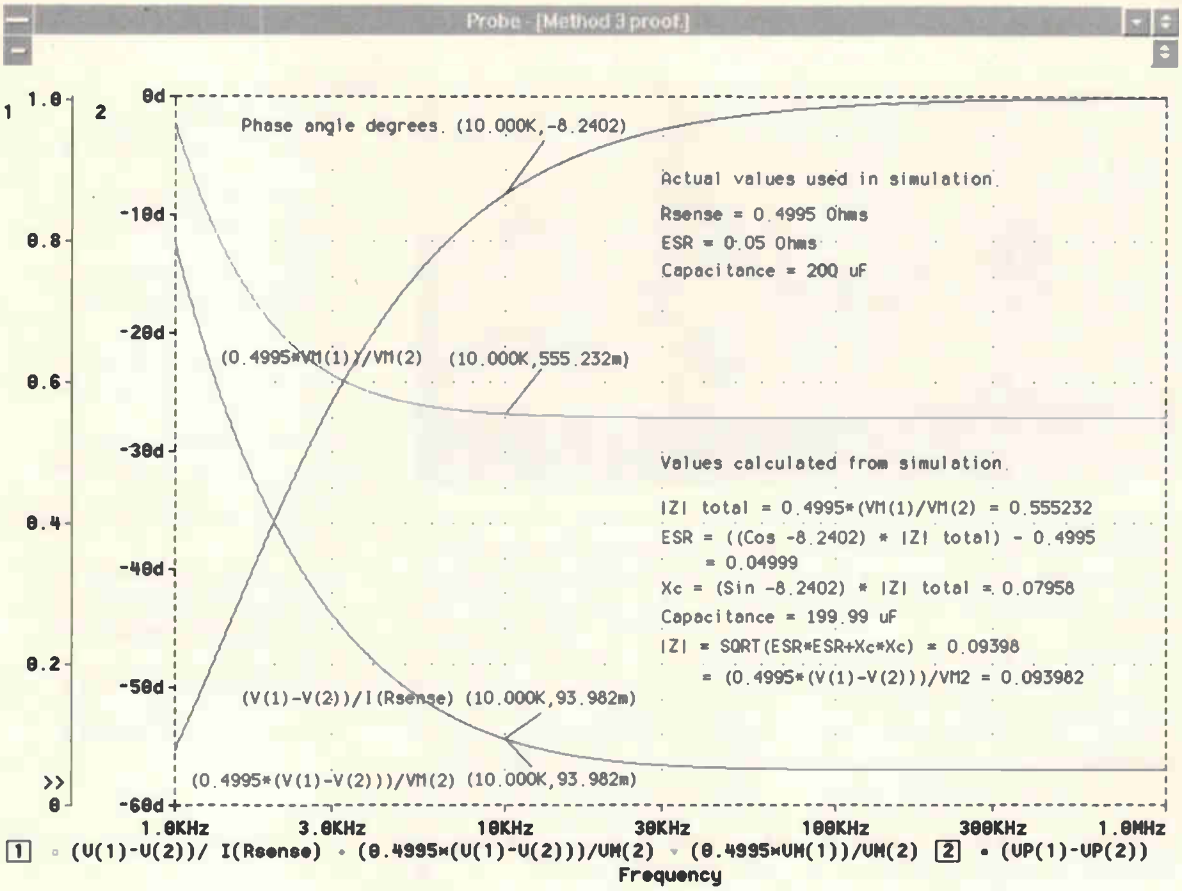


Fig. 5. Simulation used to verify Method 3. Impedance was plotted in two ways. Both curves overlay exactly in the plot. The Method 2 ratio needed to calculate ESR and capacitance is also plotted.

Measured voltage and phase angle agreed closely with the simulation. The ESR and capacitance values obtained from my measurements were within 5% of the capacitance and ESR for the assembly, **Fig. 4**.

This accuracy is more than sufficient for my electrolytic capacitor measurement needs. There's more on this method in the panel entitled, 'Implementing method 2'.

Other methods

At the self-resonant frequency of the capacitor, its measured impedance exactly equals its ESR. This frequency is determined by the capacitor's effective capacitance and self inductance.

Using this resonance technique and inserting additional inductance external to the capacitor, ESR can be measured at lower frequencies. This inductance can be provided either by extending the capacitor leads or by adding a low loss inductor.

A small additional inductance – even that from 1cm length leads – can significantly reduce the resonance frequency, allowing ESR measurements over a range of frequencies.

Method 3 – impedance meter

In method 2, both voltage measurements are ground referred, allowing easy measurement of voltage magnitude using conventional instruments. Using these complex voltages, you can derive a measurement method, giving a direct digital readout of impedance magnitude and phase of the unknown.

Suppose you have an unknown resistor in series with a known resistor to ground. Because they each pass the same current, the unknown resistor value can be calculated by measuring the voltage drop across the unknown, and the voltage drop across the reference resistor.

$$R_{unknown} = \frac{R_{ref} \times V_{drop(unknown)}}{V_{drop(ref)}}$$

If you look at the PSpice plots in **Fig.** 5, you will see that V_1 and V_2 have both magnitude and phase. They are complex voltages. Using the vector difference of these two voltages, you can calculate the impedance of the capacitor, Fig. 5.

$$|Z| = R_{sense}\frac{V(1)-V(2)}{VM(2)}$$

At low frequency, you could 'float' a conventional battery powered multimeter to measure this difference vector. For higher frequencies, a high input

Implementing Method 3

Using the jig outlined in Method 2, both voltages $V(1)$ and $V(2)$ are complex.

If you take the vector difference of these two and divide by the current passing through the reference sensing resistor, you have a direct measurement of the capacitor's impedance, Fig. 5.

$$|Z| = \frac{V(1)-V(2)}{R_{sense(current)}}$$

Voltages $V(1)$ and $V(2)$ are complex.

$$R_{sense(current)} = \frac{VM(2)}{R_{sense}}$$

where $VM(2)$ is the R_{sense} voltage drop and,

$$|Z| = R_{sense}\frac{V(1)-V(2)}{VM(2)}$$

At very low frequencies, the easiest way to measure this vector difference is by using a conventional battery powered multimeter, connected between $V(1)$ and $V(2)$.

At higher frequencies, a true differential voltmeter reading is needed. This can be provided by using a high input-impedance, instrumentation amplifier having high common-mode rejection to the highest frequencies measured.

Noted that this differential voltage can be very small, compared with the common-mode voltages at $V(1)$ and $V(2)$, so the instrumentation amplifier must also have a small output offset voltage.

The differential voltage is easily divided by $VM(2)$ using a modified PM128 meter, exactly as used for my tanδ meter design.

This method provides a direct reading impedance meter, usable over a wide frequency range. It is a ratio method. Since both ratio-ed voltages are measured concurrently, accuracy does not depend on signal source impedance or amplitude. It can even be used with a signal generator whose output changes with load and frequency.

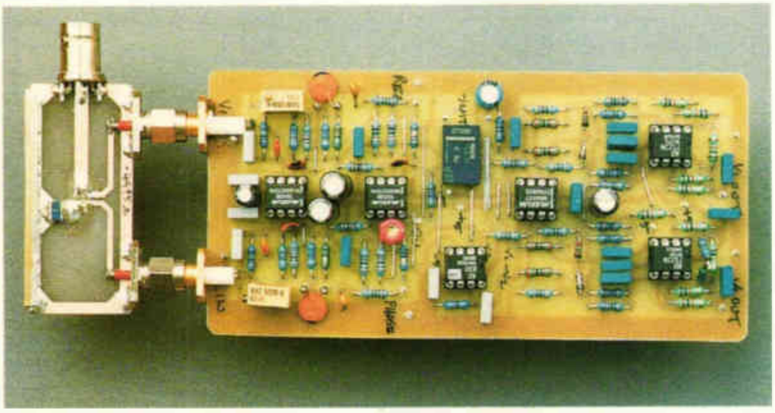


Fig. 6. My prototype test meter with special jig, developed for the Method 3 differential voltage measurements. Both are usable from 10kHz to 10MHz. This meter outputs a direct reading of capacitor impedance, from 0.001 to 1.999Ω, on a modified PM128 meter. A switch provides a direct reading of the 'R_{sense}×VM(1)/VM(2)' ratio, simplifying calculations needed for ESR and capacitance. A buffered, phase-equalised output, suitable for a phase meter is provided.

impedance, differential instrumentation amplifier with a flat response and good common-mode rejection up to the highest measurement frequency is required.

I decided to investigate this approach using two identical high-impedance input channels, similar to my RF millivoltmeter[4]. These would be followed by two identical rectifying stages, as used in this meter. The rectified DC outputs were divided using a PM128 meter modified to ratio mode, as used for my tanδ meter, in the January 2000 issue.

A relay could be used to select between measuring the two input channels or the differential measurement, as required. This new meter could then directly display either measured impedance, as the vector result of $(V1-V2)/VM2$. Alternatively it could be switched to display $VM(1)/VM(2)$ as needed for method 2. See the panel entitled, 'Implementing Method 3'.

I have designed and implemented a suitable dedicated meter **Fig. 6.** Its design, construction and use form the subject of my next article. ■

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Dynamic art

Douglas Clarkson argues that electronics engineers and artists should work together to produce dynamic works of art, and makes suggestions for simple artistic lighting effects using opto-mechanics.

Just as an artist can call up a wide selection of hues and colours to complete a creative commission, so the electronics engineer has available a wide range of products and devices that can be used to develop systems for 'creative' visual effects.

In stage effects, computers interfaced to complex production systems have allowed new heights of technical achievement. Also, once a complex lighting sequence has been set up, it can be easily replicated on successive performances.

Most electronic systems are designed with functionality in mind. But there is increasing interest not just in constructing electronic systems that perform a necessary task but in exploring the creative potential of a broad spectrum of physical systems. The aim is to basically make life more varied and stimulating.

Moving art

For a variety of reasons, most works of art are static. These predominantly consist of images frozen in two/three dimensions. It may be that artists would want to use time-varying effects, but they do not have the technical ability to implement them. Likewise it may be that those competent in electronics have the ability to bring dynamism into art, but have little interest in doing so.

When we perceive a conventional image on display, we subconsciously move in relation to it to change our perspective. In doing so, we make sense of the image's lines, colours and forms. This tells us in part how we naturally observe objects.

We discriminate boundaries between areas. We register the specific details of a particular focus of the visual field. Our brain scans for moving images. In the final product, the brain integrates all these processes seamlessly together. The process of visual perception is complex and so successful that we appear to observe effortlessly.

There is something, however, more attention seeking about images that change their appearance in time. In the supermarket aisle, attention seeking devices attract the eye of the shopper with a single pulse of a red LED every few seconds. This indicates that we have the potential to discriminate a relatively small change out of a large and complex static visual field.

If a visual display has elements that change over a specific time frame, it will receive more attention than one that is static. Perhaps one of the reasons why television is so compulsive is that it presents a visual field that is always changing. It is latching onto those receptors in the visual field that would instinctively direct us to movement.

Design objectives

This discussion revolves around using electronics to introduce cyclic variation into a discrete number of light sources. One option for the suggested scheme is to use the light sources as directly-visible elements in a display. Alternatively, they can be used in association with a piece of art to produce reflective or refractive enhancements.

While this approach introduces electronics into art, it does not restrict the input of the artist. He or she still has the task of setting up the actual nature of the 'time dependent' profile and the medium in which the lights are used. The end product can be considered as a sub-system for incorporating into appropriate works of art.

One example of how the lights can be applied involves a series of coloured panels. Each panel can be back lit by a lamp, which is part of such a controlled sequence. This is an approach that can be cascaded, adding complexity, and the output power of the light sources described can be modified.

For the time being though, I will outline a system that runs essentially on a 9V DC supply.

Ways and means

Considering the full range of analogue and digital circuits at your disposal, there is a bewildering number of ways that such a process can be implemented. Computers make the number of options boundless. This article though outlines an effective artistic light controller, made simple by using an opto-mechanical system.

Consider *n* light sources as having a function, f_1, f_2, f_3... f_n. Each function varies independently with time in a cyclic manner and with uniquely defined amplitude, phase relationship and frequency.

To make such a system interesting, around twelve or more separate lighting channels are needed. Implementing this would require 12 distinct oscillator and drive systems. It is likely that this would lead to a lack of spatial co-ordination between the elements of the display. It is also unlikely that there would any co-ordinated relationships between the components.

A less complex option would be to have a smaller number of oscillators, say four. These oscillators, designated

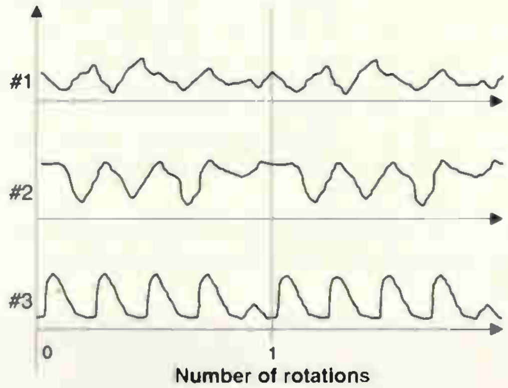


Fig. 1. Notional variation of sensor parameter as a function of degree of rotation and with the cycle repeating itself every revolution.

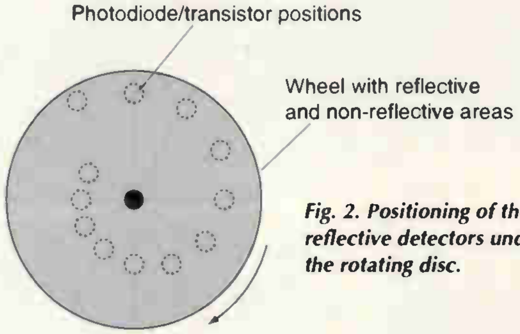


Fig. 2. Positioning of the reflective detectors under the rotating disc.

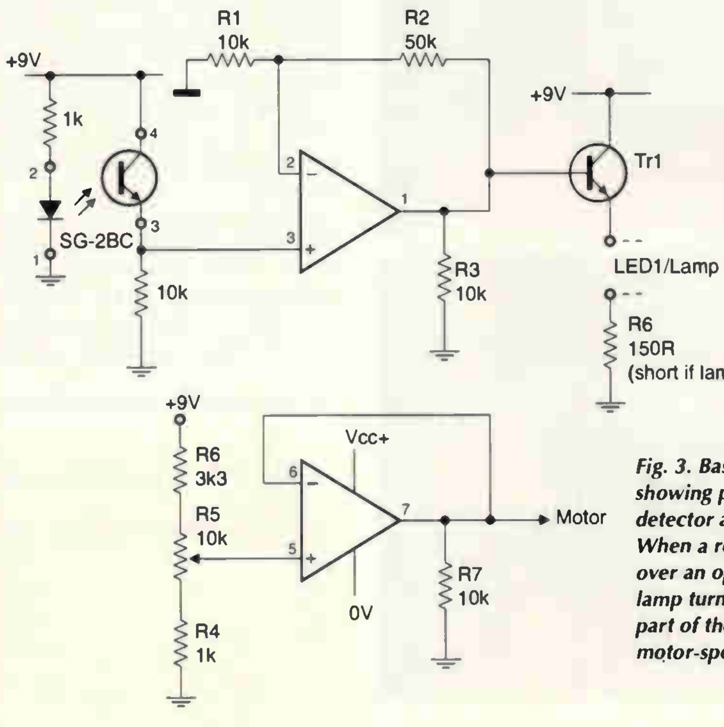
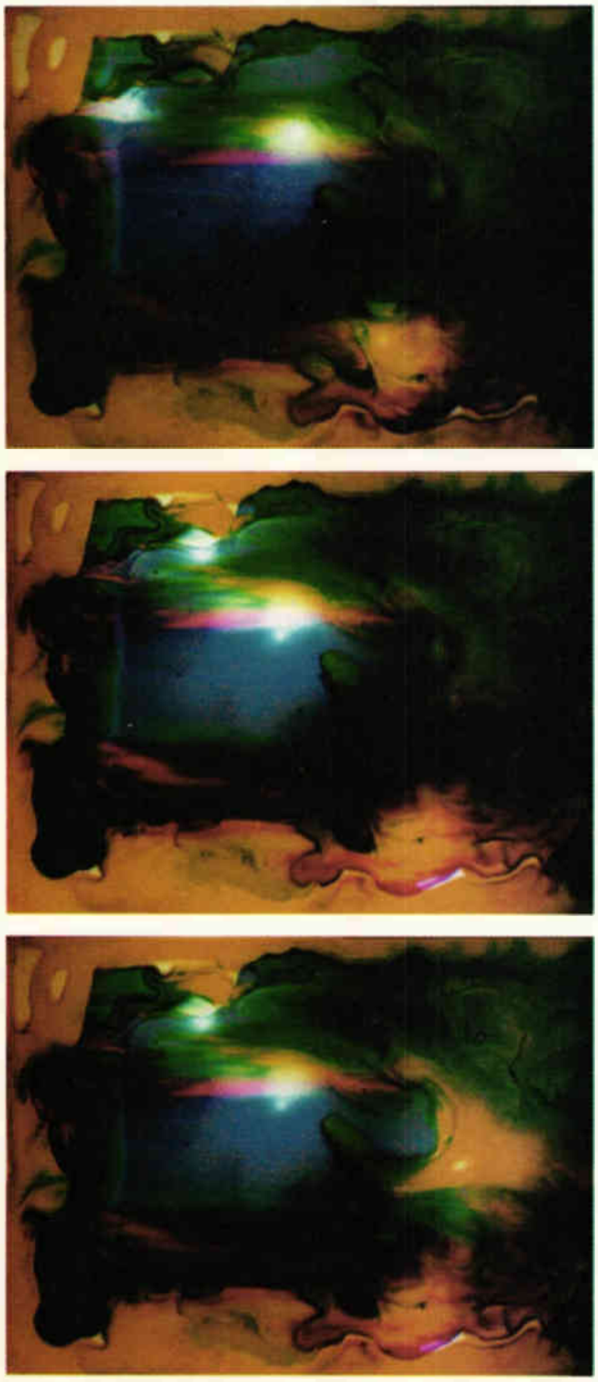


Fig. 3. Basic single channel system showing photodiode/phototransistor detector and output drive system. When a reflective area of the disc is over an opto device, its associated lamp turns on gradually. In the lower part of the diagram is the simple motor-speed adjustment buffer.

A, B, C and D, could be arranged to produce sum and difference effects, A–B, A–C, A–D, B–C, C–D, A+B, A+C, A+D, B+C and B+D. This would again tend to produce a lack of spatial co-ordination between the elements though, and we are not really in control of the combinations.

Fig. 6. A series of photographs taken using a cyclic pattern change on painted glass and six channels of variable light output.



Also, these options would require more than a few analogue/logic chips to implement effectively.

Towards coherence

It is possible to have a PC provide the ultimate solution. A computer with 12 a-to-d channels could output appropriate analogue voltages under software control. As a design problem, though, this is beginning to look daunting and expensive. Are there more practicable ways of achieving the same end point?

A more attractive option is to maintain the same 'cyclic period' for all elements but to introduce independent phase and amplitude modulation for all elements. **Figure 1** shows some notional variations of sensed parameter as a function of degree of rotation. The cycle essentially repeats itself after each cycle.

A novel way to establish this phase/amplitude relationship across say twelve discrete signals is to encode optical reflectivity/transmission parameters onto a circularly rotating disc and use phototransistors/photodiodes. This is done in association with variation of optical properties of the disc surface to provide the phase and amplitude relationships. A complete cycle of signals will thus be replicated in one revolution of the encoded disc, but each channel can have its own identity.

Some element for synchronising the variation in levels of the illumination channels is needed. But remember that the goal is for visual effect only – not for precise control and processing.

Figure 2 shows a disc for such a system, involving twelve photodiode/phototransistor elements in reflective mode.

In my design, the small optical units are actually positioned flush with the circuit board. The rotating disc – reflective side face down – blocks out stray light. As the disc rotates, it passes under units of reflective detector pairs.

As an alternative, it would be possible to use the system I am proposing in 'transmission' mode using ambient light to modulate the detector array.

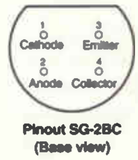


Fig. 4. Pin out of photoreflective device used in the author's prototype (SG-2BC - Farnell code 441-566). Device diameter is 4mm.

The optical modulator

Each detector pair comprises an integral phototransistor and photodiode. In this way, a voltage is produced that is proportional to the degree of reflection of the surface. Thus the sequence of areas to be illuminated can be carefully selected to take advantage of the known sequence derived from the rotation of the encoded wheel.

It may also be possible to introduce an element of randomness into the display. This could be done by building some eccentricity into the circular disc, which rotates under the phototransistor/photodiode heads.

This module is intended to be self contained. After being powered up it will operate as an integral part of an item of art with a minimum of controls. A different 'template' of the disc produces a different series of voltage signals from each of the channels.

My prototype consumes very little power. Its motor only consumes around 50mA at 6V. It would also be possible to use RGB LEDs so that the variation of sensed voltage was also translated to colour changes.

The concept of a relatively slow rotating disc maps well into the visual changes that may be expected. Thus a typical repeat interval of 15 seconds requires low levels of rotation of around 4rev/min. This requires either a DC motor with very high gearing or a stepper motor with small step intervals.

Table 1. Table of rotational speed of iron core motor as a function of drive voltage.

Rotational speed (rev/min)	Drive voltage
2.9	2.8
4.8	4.0
5.7	5.0
8.4	7.5

A stepper motor with control of input frequency from 1 to 50Hz would be appropriate. But a stepper motor with 48 steps per revolution is a little too coarse relative to the true continuous movement of the DC motor. A 1.8° stepper would provide adequate resolution, but they tend to require more complex drive systems and consequently more power consumption.

In my prototype, I used a small geared iron-core motor. With variation in voltage control, this gives a range of rotational speed as indicated in **Table 1**. The motor is essentially operating in unloaded mode.

Elements of circuit design

The elements of detection/amplification of a specific channel are indicated in **Fig. 3**. Nominally, the system is adjusted so that full reflection outputs 6V and zero reflection, 1V.

Power transistors are included in the output drive elements to allow either LEDs or conventional filament bulbs to be used. Resistors in the output circuit can be included as needed. The maximum current per channel is around 100mA. **Figure 4** shows the pin-out of the photoreflective device I used.

A simple potentiometer controls the motor's rotational speed. The motor can be switched to rotate either clockwise or anticlockwise. A conventional quad operational amp carries out the simple electronic signal processing.

Enter the artist

After establishing the core 12-channel unit, outlined in **Fig. 5**, there is then the creative challenge to create the 'reflection template' to control the series of changes with rotation. This is where the artist changes place with the electronics designer in order to communicate an idea or effect to the audience.

Why not cover the underside of the array with 4mm diameter ball bearings and use the physics of light reflection to provide a highly complex level of variation? Bearings are only one of many such 'translational devices' for transforming a change in some physical parameter to a more visible set of parameters – which is part of a separate aspect of artistic expression.

Figure 6 indicates a series of photographs taken of an 'installation' where cyclic pattern change on painted glass operates on six channels of variable light output.

Final thoughts

In my view, the step from an electronic device with an everyday function to one that is promoted as a phenomenon of artistic expression could double the end value of the system.

Surely there is scope for serious and committed 'wired' artists, who could produce devices and constructions that would be gladly accepted with pride in a 21st-century habitation, organisation or institute?

Such devices must of course be constructed carefully, to good standards of safety, design and functionality. They would give satisfaction to their designer. But they would also satisfy the customer that there is evidence of careful thought tempered with rational design and construction. This would produce a practicable contrast to the splendidly vacant collections of contemporary art.

Are there artists who would welcome assistance in translating effects into reality using electronic techniques? Conversely, are there are electronics engineers who could work with 'artists'.

Every self respecting electronics engineer should ask himself/herself the question, "If I am so creative with electronic components and circuits, are there areas of creative endeavour that might I turn my hand to?" ■

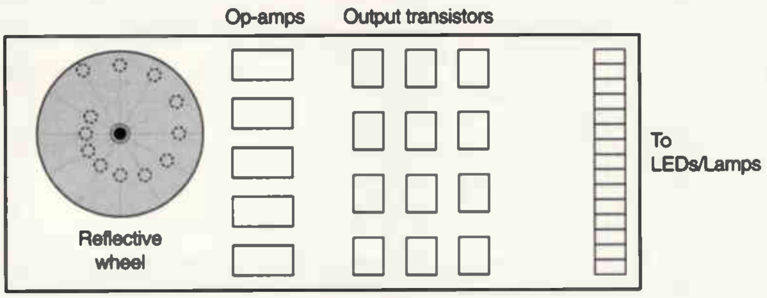
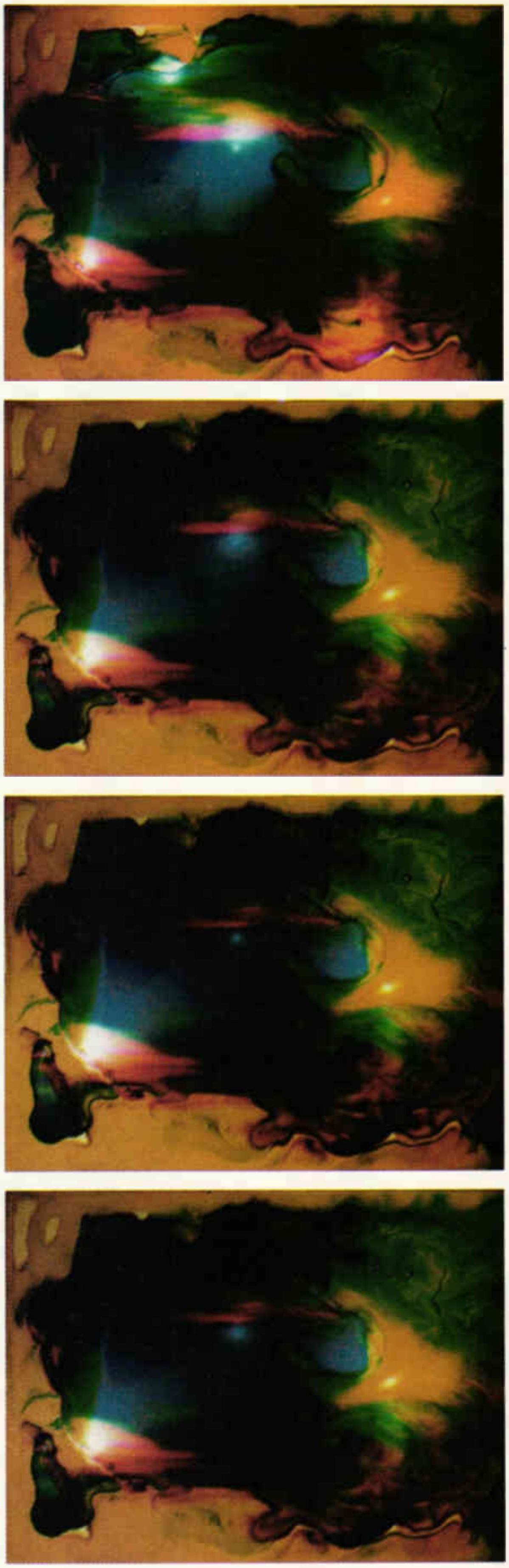


Fig. 5. Outline layout of the author's twelve-channel system.



NEW PRODUCTS

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Data acquisition software

Version 5.5 of Adept's Dasylab data acquisition software lets users monitor test stands, development rigs and processes on screen. It also alerts users to predefined events remotely, via networks, the web, email or SMS-enabled mobile phones. Its point-and-click interface lets the user set up, see and document data acquisition applications without programming or command-line instructions. The software is for applications that link sensors and instrumentation to a PC. On-screen displays and controls, triggering and signal processing are set by dragging icons into place and connecting them with a mouse click.

Adept Scientific
Tel: 01462 480055

MOSFET for 42V automotive

Intersil's latest power MOSFET is designed for the new automotive 42V battery systems currently being developed for the next generation of vehicles.
The HUFA7510S3S UltraFET is a 75V, 75A MOSFET with an on-resistance of 10mΩ, which the company says gives one of the lowest resistances per area of silicon available. In addition the FET features a 75V drain-source breakdown voltage, giving it the ruggedness required for higher-voltage automotive systems. The device is engineered to withstand high energy in avalanche mode and features a combination of low gate charge, fast recovery time, and a lower reverse body-diode charge (QRR) for applications that require fast switching speeds and improved power efficiency. Intersil's UltraFET process technology achieves the lowest possible on-resistance per silicon area, lower gate charge and improved dynamic switching characteristics. UltraFET devices withstand high peak currents and energy in the avalanche mode, important when switching inductive loads. The HUFA7510S3S is immediately available for volume production in the TO263 package.

Intersil
Tel: 01344 350250

P-channel HEXFET power MOSFETs

International Rectifier has introduced four 12 and 20V p-channel HEXFET power MOSFETs for portable applications. They come in TSSOP-8 packages. The 12V IRF7701 has 11mΩ $RDS_{(on)}$ at 4.5V. Applications include cell phones, digital cameras and notebook computers. The other three products are the IRF7702, IRF7750 and IRF7700.

International Rectifier
Tel: 020 8645 8001

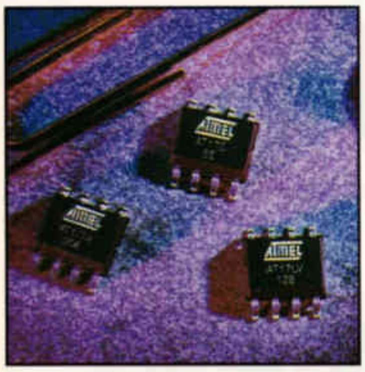
Development environment

Netsilicon has announced Net+OS – an out-of-the-box product that integrates an embedded development environment, real-time operating system and the firm's Net+Works architecture. It contains the development tools, networking software and hardware needed to begin application development using system-on-chip technology to design internet and Ethernet connected devices. It includes Green Hills Software's Multi2000 IDE for Windows, ThreadX RTOS from Express Logic and Net+Works drivers, protocols and services for web, email, FTP, flash and SNMP applications. It comes with a Net+Arm development board including an integral 10/100 Ethernet, two serial ports, two parallel ports, RAM, ROM and a JTag port. The package also includes a debug tool that connects the JTag port to the parallel port of the development PC and is used for code downloading.

Netsilicon
Tel: 00 1 781 398 4548

Reprogrammable FPGA memories

Atmel has introduced the AT17 reprogrammable FPGA configuration memories in eight-pin narrow-bodied SOIC packages



and 65k, 128k and 256k densities. These nonvolatile memories can be used to load program data into any SRAM based FPGA during power-up.
Atmel
Tel: 00 1 408 441 0311

Quad 10-bit transceivers

Vitesse has introduced the VSC7182 and VSC7186 quad 10-bit Gigabit Ethernet and Fibre Channel transceivers for switches, routers, hubs, host adapters, backplane connectors and test equipment. These 3.3V ICs allow full duplex operation at 1.05 to 1.36Gbit/s. Each of the four transmitters serialises 10-bit input (8B and 10B encoded) for transmission, while the four receivers perform the reverse on incoming serial data. They have integrated JTag access ports. Each device dissipates 2.2W typical and they come in 208-pin, 23mm BGA packages.
Vitesse Semiconductor
Tel: 01634 683393

3V transceiver

Fairchild has introduced the 74LVTH16500 18-bit universal bus transceiver for 3V systems. It has tri-state outputs and operates at 3.7ns maximum at 3.3V V_{cc}. There is input and output interface capability to 5V V_{cc} systems. For off-board driving applications, such as backplanes, memory arrays, telecoms switches and networking, the device has outputs rated at +64 to –32mA. Latchup performance exceeds 500mA. It is fabricated using BiCMOS.
Fairchild Semiconductor
Tel: 01793 856856

Optical encoders

Optical encoders are available from Tecan through a bespoke service for the supply of discs and similar components in control, positioning and measurement applications. Different manufacturing techniques are used. Photochemical machining, for example, will produce burr and

stress-free results typically with accuracies of ±10 per cent of the thickness of the metal used - from 1.5 to 0.01mm. For more accuracy, electroforming techniques are used, a typical tolerance being ±8μm and, in some cases, ±2μm can be achieved.
Tecan Components
Tel: 01305 765432

Six-channel audio d-to-a converter

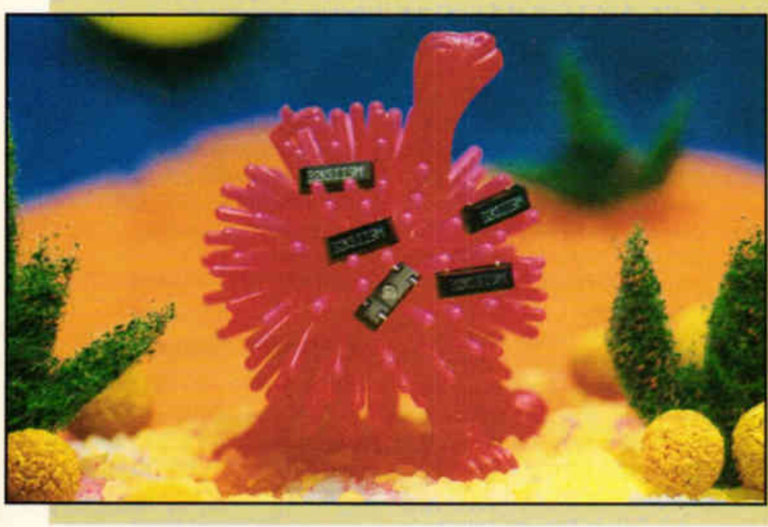
Burr-Brown's PCM1604 is a six-channel audio d-to-a converter with 24-bit resolution and 192kHz sampling for multi-channel audio applications such as DVD. It contains six 24-bit 192kHz converters on a monolithic IC and can be used in AV or HDTV receivers, car audio, multi-channel home theatre, surround-sound processors and digital mixing consoles. The device uses multi-level delta-sigma modulation to improve audio dynamic performance and reduce jitter sensitivity. Dynamic range is 105dB and it has –95dB THD+N on each channel. The internal digital filter operates at 8 times oversampling for a 96kHz sampling rate or 4 times for 192kHz, with selectable sharp or slow roll-off. The filter has -82dB stopband attenuation and ±0.002dB passband ripple. Functions include digital attenuation, mute and zero flag for each channel. The device accepts standard audio data formats of 16, 18, 20 and 24-bit and can be used with 128, 192, 256, 384, 512 or 768 f_s system clocks. It has single-ended analogue outputs and operates on dual +3.3 and +5V power supplies.
Burr Brown
Tel: 01923 233837

Programmable buzzer

Dau has introduced an audible buzzer with a choice of 16 outputs. The output can be set to be constant, swept tone or one of 16 options. It can be reset. For use in vehicles or buildings as an alarm, it has a sound output of 90dB at 1m. It can be hard wired or connected with spade terminals to the power source. The unit can be run from 12 to 35V DC or 12 to 24V AC. The AC can be applied directly to the unit and is a standard in the licensed trade. The DC option is standard for vehicles. The choice of tone can be set by a code switch, which the user can change. The device measures 61 by 43mm and is 28mm high.
Dau Components
Tel: 01243 553031

BDM tools for Motorola

Comsol has been appointed European distributor of background debug mode (BDM) tools from P&E Microcomputer Systems. DOS versions are used as the development environment in Motorola evaluation board packages. The Windows

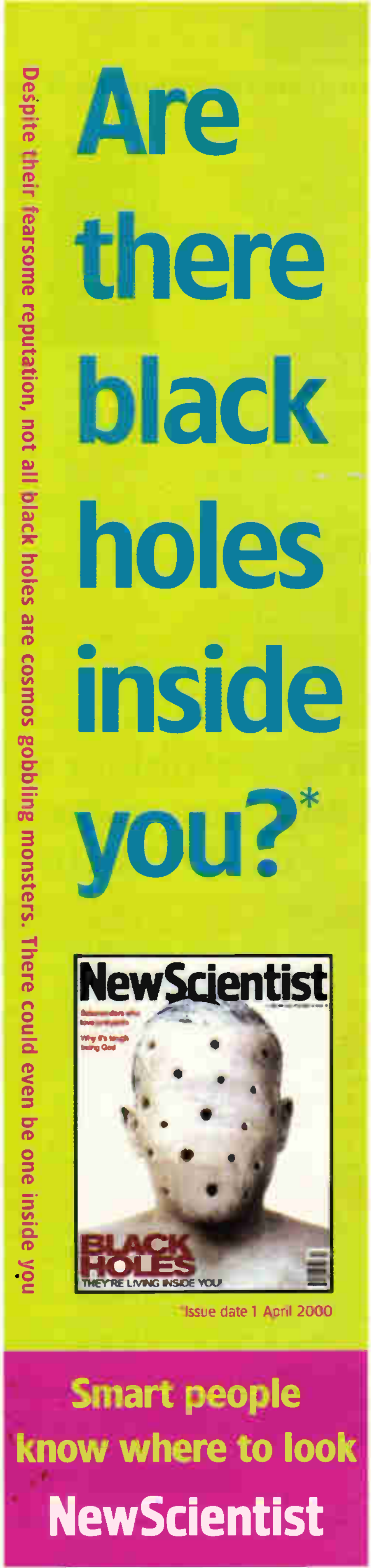


Tuning fork oscillator

Flint has introduced a surface-mount tuning-fork crystal from Raltron for time of day clock applications in portable communication equipment and other compact appliances. The RSE comes in an SMD package and 16mm tape packaging, suitable for automatic and HD surface mounting. It is available in A, B, C and D options, with A and B having a footprint of 10.41 by 4.06mm and C and D 8.7 by 3.7mm. A and C come with single I/O lead connections and B and D with dual connections. Available with 6 or 12.5pF load capacitance, it has a nominal frequency of 32.768kHz and a frequency tolerance of ±20ppm maximum. Operating range is –40 to +85°C, and the device can withstand up to 230°C for 20s in solder reflow conditions.
Flint
Tel: 01530 510333

CIRCLE NO.113 ON REPLY CARD

CIRCLE NO.114 ON REPLY CARD



CIRCLE NO.115 ON REPLY CARD

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versions include in-circuit debugger, BDM programmer, register editor, direct BDM control libraries and BDM cables.
Comsol
Tel: 01932 829460

4Mbit SRAMS

Memotech has announced 4Mbit SRAMs from Brilliance Semiconductor for battery powered applications such as digital set-top boxes. Available in 256k by 16 or 512k by 8 configurations, they are made on 0.25μm CMOS. Based on a six-transistor memory cell design, they have a data retention voltage of 1.5V with an operating current of 30mA for the 512k by 8 device and a standby current of 3μA. They come in Jedec standard TSOP, TSOP2, STSOP and CSP packages.
Memotech
Tel: 01223 370060

DC power supplies

The PQ range of current-fed DC power supplies from Kingshill includes 44 models supplying 10 to 625V DC and 5 to 600A DC, with powers up to 10kW. The units function as voltage or current sources. In voltage-source mode, if the load increases above the current command setting, automatic crossover to current mode occurs. Differential amplifiers isolate programming lines from DC output, enabling remote programming at any distance from the load. The units are programmable by resistance, voltage, current or optional IEEE488 and RS232. Diagnostic functions are integral to the control loop, while circuits identify the control function. There are three levels of overvoltage and overcurrent protection. They have start and stop push-buttons.
Kingshill
Tel: 01634 821200

Optical spectrum analyser

Agilent has introduced an optical spectrum analyser for characterising WDM optical components and systems. The 86140B provides a wavelength accuracy of 10pm from 1480 to 1570nm. It includes a built-in WDM application for measuring and recording channel power, wavelength and optical signal-to-noise ratio (OSNR). Applications include testing WDM passive components such as filters, multiplexers and Bragg gratings.
Agilent Technologies
Tel: 07004 666 666

Transistor array in 0605 packaging

Rohm has introduced surface mount small-signal bipolar and digital transistor arrays in 0605 equivalent packaging for mobile phones, camcorders and personal stereo equipment. More than 30 transistor devices

GPIB and Ethernet Interface for PCI

National Instruments has introduced the GPIB-ENET/100 Ethernet-to-GPIB controller and the PCI-8212 combination GPIB and Ethernet interface for PCI. The interface is for connecting, sharing or controlling GPIB instruments on Ethernet networks. Both are HS488 compliant. With the controller, users can access remote test equipment from anywhere in the world via TCP/IP protocols on 10baseT and 100baseT networks. It is shipped with NI-488.2 and NI-Visa API software for Windows 2000, NT and 9x. Any program previously written in either API runs unmodified on the controller. NI488.2 and NI-Visa can be integrated with Labview and Measurement Studio. The controller can be configured with DHCP or with a configuration utility. No dip switches or jumpers are required and it has installation options including rack mounting, Din rail mounting, wall mounting and stackable standalone.
National Instruments
Tel: 01635 572400

Please quote *Electronics World* when seeking further information

are available in five and six-pin EMT packages including dual small-signal bipolar and dual digital transistor arrays. The latter have built-in bias resistors and optional diodes to simplify circuit design and reduce the need for external components. Suitable for operating voltages up to 50V, the transistors are rated for currents up to 150mA.
Rohm Electronics
Tel: 01908 282666

One-chip RS232 transmit/receive

Intersil has introduced single-chip RS232 transmitters and receivers for PCs, peripherals and portable, battery operated appliances. The ICL3241 and ICL3243 have a transmitter circuit that converts TTL and CMOS signal levels to RS232 levels at minimum data rates of 250kbit/s. Applications include laptops, palmtops, printers, terminals, barcode readers, scanners and high-speed modems. These 3V products require 0.3mA supply current during normal operation. A manual and automatic power-down features increase battery life by reducing standby supply current to a 1μA trickle. A shutdown mode conserves energy in battery powered applications. The receivers remain active during shutdown. The ICL3243 detects when the receiver inputs go to ground from, say, a disconnected RS232 cable, and automatically places itself into shutdown mode.
Intersil
Tel: 01344 350250

Humidity sensor

Panametrics makes a general purpose thin film polymer capacitive type RH sensor for OEMs. The Minicap 2 sensor has a TO-18 configuration. The dielectric constant of the polymer thin film changes with the atmospheric RH, resulting in linear capacitance changes as a function of relative humidity. It is

unaffected by water condensate, is immune to most reagent vapours and can handle temperatures up to 180°C. It can measure RH from 5 to 95 per cent with a linearity of ±1 per cent RH and a typical response time of less than 60s for 90 per cent of total range, or faster for small step changes. It requires a +1V excitation and returns a typical signal between 170 and 230pF from 0 to 100 per cent RH.
Panametrics
Tel: 020 8643 5150

CompactPCI processor

Teknor Applicom has introduced the Raptor RAP-C810 CompactPCI processor in a 3U footprint. The board is available from Wordsworth. Celeron-powered with up to 500MHz CPU speed, 66MHz front side bus and 128kbyte level two cache, the board is for test and measurement, in-vehicle transportable and automation applications. Features include integrated 2D and 3D graphics and up to 128Mbyte of synchronous SDRAM on a 144-pin SODIMM socket.
Wordsworth Technology
Tel: 01732 861000

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Epson has developed the SG-W oscillators with output frequencies up to 135MHz. They use PLL technology to reduce long-term jitter levels. They are for networking applications, PCs and peripherals. The SG-531xxW and SG-615xxW each come in three versions with output frequencies of 66.7 to 135, 55 to 135 and 26 to 135MHz. The SG-636xxW comes in 41 to 135 and 32 to 135MHz, and the SG-710xxW in 67 to 135 and 80 to 135MHz versions. Frequency stability is ±100 or ±50ppm. Supply voltage is 3.3 ±0.3V or 5.0 ±0.5V.
Epson Electronics
Tel: 00 49 89 14005 363

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Hyundai is launching Risc and DSP microcontrollers. The GMS30C2216/32 embedded microprocessors are for applications that handle digital and analogue signals such as DVDs, flash cards, digital cameras, PDAs and multimedia electronic systems. Based on an architecture from Hyperstone, the El-32x core combines a 32-bit Risc core with a 16 to 32-bit fixed point DSP using an integrated instruction set to provide a unified programming model. The 32-bit wide 96-way register set supports parallel operation of the ALU, DSP and load-store units, delivering 2.4Gop/W. The 8kbyte of on-chip DRAM runs at CPU clock speed. Operating at 108MHz, power consumption is 180mW at 3.3V. On-chip peripheral support includes DRAM controller, software programmable PLL, 32-bit timer, three serial I/O lines, interrupt controller and a 16 to 32-bit bus interface addressing 4Gbyte of memory. It has power-down and sleep modes. Power-down halts instruction execution, while DRAM refresh and the internal timer are maintained. Sleep stops everything, drawing 30μA until a wake-up signal is received.
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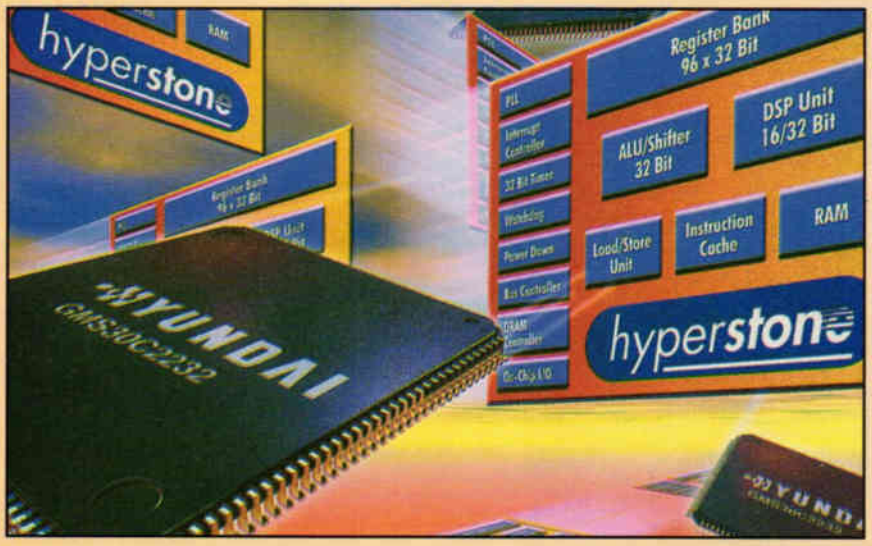

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CIRCLE NO.118 ON REPLY CARD

distributed power applications in computing and communications. The single-output units have output voltages of 3.3, 2.5, 2, 1.8 and 1.5V, with a nominal input of 48V and an input range of 36 to 75V. Users can draw up to 60A from the BXF300 and 80A from the BXF400. Both include short-circuit, overvoltage and over-temperature protection, remote sensing and remote on-off.

Artesyn Technologies
Tel: 00 353 24 25572

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Matrox Imaging
Tel: 01753 665500

Inductors with shields

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twelve sizes with different height to size to current ratios. The 3622's package is 5mm in diameter, with up to 1mH inductance. Lower values carry

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Meggit Electronic Component
Tel: 01793 487301

Development kit for TCP/IP

Rabbit Semiconductor has released the Rabbit 2000 TCP/IP development kit containing the hardware and software to design a microprocessor-based application that networks via Ethernet and uses internet protocols. The kit contains a TCP/IP development board with an 8-bit microprocessor. Dynamic C software development system, power supply and PC serial cable for real-time debugging. The software includes integrated editor, compiler and debugger, so no in-circuit emulator is required. Sample demonstration programs include HTTP Web server and SMTP mail client. Hardware reference schematics help with development. Executable code can be downloaded into flash memory or optional battery-backed RAM. Two communication ports are available – an RS232 port and a factory configurable port for either RS485 or RS232. Features include four high-current outputs, four digital inputs, seven timers, real time battery-backable clock and 10baseT Ethernet interface. TCP/IP source code is provided.
Rabbit Semiconductor
Tel: 00 1 530 757 8400

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Devlin Electronics has developed the ML family of keytops to match the Cherry ML key switches. The keytops are

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Devlin Electronics
Tel: 01256 467367

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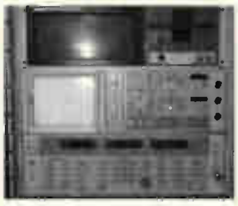

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Loudspeaker crossover networks: constant voltage or constant power? John Watkinson's answer might surprise you.

SPEAKERS' CORNER

In the last few issues of this magazine there has been a significant number of constant-voltage crossover circuits published, not to mention a certain amount of debate. It is heartening to see such a resurgence of interest in this subject, but there does appear to be a some confusion over the question of constant voltage or constant power operation.

In an ideal loudspeaker, the use of more than one drive unit and the resulting crossover would be transparent to the listener. The internal operation of the loudspeaker is the designer's problem. But the sound it produces should be a function of the audio waveform and by definition should not reveal anything about that construction.

At this point, the conventional speaker designer will state that theory is all very well but it can't be achieved in practice and so there is no point in trying. The only reply I can make is to point out that active loudspeakers with inaudible crossovers are commercially available. And they really do sound more realistic.

Clearly the ideal is a practical possibility. But how is it done? Quite simply the answer is to do what acoustics allows, with the degree of precision the ear demands. This means that the crossover design, the crossover frequency and the directivity and positioning of the drive units all have to meet certain criteria. If all are not met, the result is unsatisfactory.

For example, an ideal crossover topology used with an inappropriate mix of drive units or an inappropriate crossover frequency won't sound much better than a conventional crossover. This would lead to the false conclusion that the ideal crossover is unnecessary.

Figure 1a) shows one of the criteria we are trying to meet. This is that the radiation from the two drive units should sum to the original waveform at the listener. However, this isn't the only issue. Figure **1b)** shows that another requirement is for this condition to be met off-axis as well. Then the reverberant field will also carry no evidence of the use of two drive units.

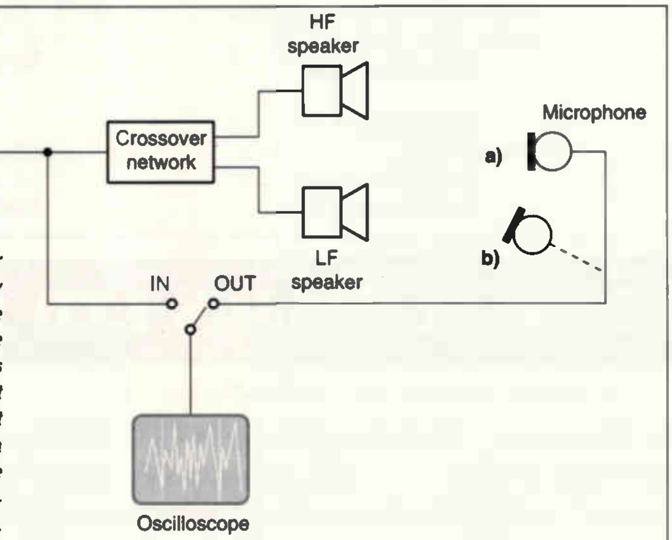

Fig. 1. The criterion for accuracy in a two-way speaker is that the waveform created in the air at microphone a) is identical to the input waveform. This is difficult enough, but it is also a requirement to achieve the same result with off-axis microphone at b).

If we meet only the first criterion, we could get a wonderful on-axis frequency and time response. But the power response – i.e. the power radiated in all directions – could be unsatisfactory in the vicinity of the crossover. This symptom is only too obvious on most two-way loudspeakers.

The only solution is that the directivity functions of the two drive units must be similar in the crossover region. In traditional speaker design this criterion is seldom met.

Figure 2 shows that to obtain an inaudible crossover, the two drive units must be acoustically close at the crossover frequency. This means that the distance between them must be smaller than the crossover wavelength.

If this requirement is met, then it doesn't matter which drive unit contributes to the sound, so that the two contributions can add up in front of the two drivers. In practice, this requires a relatively low crossover frequency assuming typical woofer dimensions.

Once the drivers are acoustically close, something wonderful happens. Once in a while, amidst the struggle which is high quality audio, physics gives the designer a break. **Figure 3** shows that if a constant voltage crossover is used, in which the high and low frequency waveforms add back to the original, we can get both a flat voltage response and a flat acoustic power response through the crossover frequency.

This defies conventional logic. Figure **3a)** shows that both outputs of a constant voltage crossover are – by definition – 6dB down, i.e. half the midband voltage, at the crossover frequency.

As power ordinarily goes as the square of the voltage, one would conclude that each driver is receiving one quarter of its midband power, and so it is. However, it is completely incorrect to assume that the radiated power at the crossover frequency is then half the midband power.

What happens at the crossover frequency is that there are two acoustically close drive units radiating the same signal. These drive units are in one another's close field. So what happens is that the radiation from one doubles the radiation impedance seen by the other and *vice versa*, Fig. **3b)**.

The result of the doubling of the acoustic impedance is that the output of each driver is only 3dB down (half power) at the crossover so the sum of the two contributions is 0dB down, or flat, as in Fig. **3c)**.

I must stress that to make this work requires suitable drive units. Don't think that a blameless constant-voltage crossover will transform the performance of the average two-way bookshelf travesty because the dome tweeters won't work at a low enough frequency. If the transducers aren't up to it and the crossover frequency has to be raised, the technique doesn't work.

Once the two drive units are no longer acoustically close they don't augment one another's impedance at the crossover and a constant-voltage crossover doesn't give the required results. Instead a constant power crossover is needed.

In general, the spacing between the drive units will be a large part of a wavelength and the directivity function when both drivers are working at the crossover will be seriously lobed and audible as a coloured reverberation.

All that can be done is to narrow the crossover region by using steep filters. The trouble is that they almost invariably sound pretty grim because they impair the time or step response.

As far as I can see, achieving an inaudible crossover to a dome tweeter isn't possible theoretically, nor have I heard it done. The poor directivity function of domes makes them fundamentally narrow bandwidth drivers and pushes crossover designers into corners that they wouldn't be in with wider band transducers.

The result is the characteristic 'dome sound' which results from a combination of the poor directivity characteristic of the driver itself along with the poor phase characteristics of the necessarily steep crossovers.

I personally prefer the 'original waveform sound' where the loudspeaker hasn't put hoofprints in the acoustic output. ■

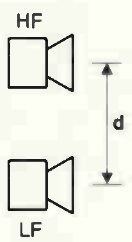

Fig. 2. Distance between speakers d must be small compared to wavelength of sound d at crossover frequency. This suggests closest possible mechanical coupling and low crossover frequency.

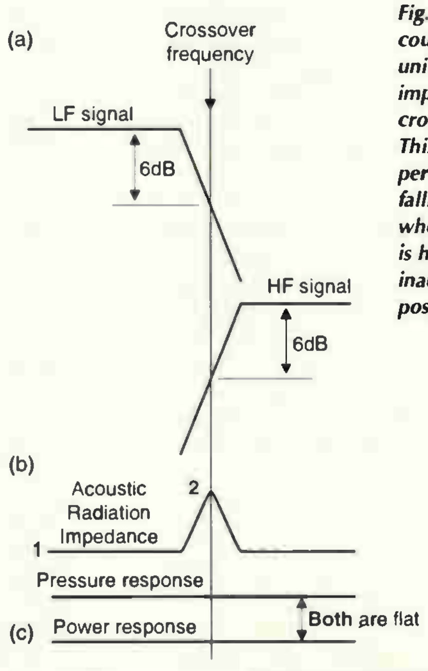

Fig. 3. Acoustic coupling between drive units doubles the impedance at the crossover frequency. This makes up perfectly for the power falling to one quarter when the drive voltage is halved. Thus an inaudible crossover is a possibility.

SPECIAL RELATIVITY right or wrong?

In telecomms, synchronising clocks around the Earth is an important issue. Al Kelly believes that the correction applied to such clocks is not explained properly by existing theories because they rely on the notion that light has a constant velocity.

Light travels around the Earth faster eastward than westward. Does not the Special Theory of Relativity claim that the speed of light is a constant?

The standard answer to this conundrum is that the Special Relativity applies solely to uniform straight line motion. It is claimed that, no matter how big the circle, motion along its periphery cannot be said to approach straight line motion. This is said to be so even if the best measuring instruments devised cannot pick up the divergence from straight line motion over the portion of circumference being used.

In his paper launching the theory in 1905, having applied his theory to straight line motion, Einstein then applied it to a closed curve of any shape. This rather undermines the popular explanation!

The circuit does not have to be as large as the cross section of the Earth to detect this effect. A Frenchman, Sagnac, found in 1914 that light signals go around a disc of 1 m diameter faster against the spin of the disc than in the direction of the spin, **Fig. 1**.

By a measurement made solely upon the spinning disc, he recorded the difference in the time of the signals sent in the opposing directions. As shown in the diagram, the time for the signal to traverse from the light source at A via C-D-E-F-C is less than the time in the opposite direction A-C-F-E-D-C.

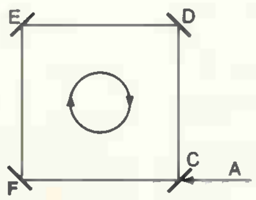

Fig. 1. Sagnac test. Michaelson and Gale used such a test to show that the speed of light is different depending on whether it is travelling eastward or westward.

The light source was fixed to the spinning disc; the measurement of the time difference was at an interferometer at C also fixed to the spinning disc. Sagnac produced a formula that exactly matches the difference in the times taken in opposing directions.

This formula can be derived, by assuming that the light travels in relation to the fixed laboratory. But, the measurement of the time difference is done solely aboard the disc. What can this mean? The only explanation possible is that the time aboard the spinning disc and in the fixed laboratory is the very same. This is not in accord with Special Relativity.

Another defence of Relativity theory is the claim that the light path upon the disc is longer in one direction than the other. But, the circumference of the disc, as measured by someone upon it, is surely the very same in both directions.

In a test in which signals are sent around the Earth from a fixed position, the light signal is emitted upon the spinning Earth, and the record of the time difference taken by the opposing light paths is solely upon the Earth. To claim, in this case, that the circumferential distances east and west are different is bizarre.

A test done in 1926, by Michelson & Gale, first showed that the speed of light was not the same eastward and westward around the Earth. They constructed a rectangular circuit of over a mile in periphery. This was a Sagnac test on a disc of diameter 9500000m diameter – the diameter of the Earth at that latitude.

In the case of the Earth at the equator, the difference between the times taken in opposing directions is 414.8ns. This result is enormous when considered against the accuracy – one million times better than that – required to-day of standard clock-stations. The difference between the times going northward and southward around the globe is zero.

The International Telecommunications Union (ITU) sets the rules for synchronising clock stations. A signal sent eastward around the globe has to allow for the fact that it travels at the speed of light c plus the rotational speed of the Earth at that latitude v, giving $c+v$.

A signal sent westward has to allow for the fact that the speed of the signal is $c-v$. According to Special Relativity theory, the speed of light is a constant. Not only that, but the direction is not supposed to matter to Special Relativity; going east, west or north should have the very same speed.

As shown in the **Fig. 2**, a ground clock station at A is to be synchronised with a ground station at B, via a satellite S. The signal sent from A to B travelling in the same direction as the spin of the Earth takes more time than in the reverse direction.

A third defence that is used is that the $c+v$ and $c-v$ are only average figures and that the instantaneous velocity of the light signal is always equal to c. On a perfectly circular circuit the $c+v$ in one direction is the velocity that would be measured at a million spots on the circumference; how then can the average become c?

Take v as 250000km/s. In a million measurements the speed of the signal is 550000km/s while the claim is that the instantaneous speed is 300000km/s. Bunkum.

The ITU apply the necessary correction and call it 'a relativistic correction, for the rotation of the Earth'. But it is not a relativistic correction. A person at a fixed position sends signals eastward and westward around the globe.

There is no relative motion concerned. How then is it that the signals arrive back at different times? There is only one sensible explanation. The signals are travelling at different speeds around the globe. Taking the speeds to be $c \pm v$ in the opposing directions agrees exactly with the experimental result.

A test was done in 1976 in which an atomic clock was transported on an aeroplane from Washington (USA) to Tokyo. Also, a signal was sent between the two clock-stations. A correction had to be applied to the signal exactly as described above, while the transported clock needed no correction.

Despite this, the ITU claims that a correction of 207.4 nanoseconds has to be applied to the time on a clock brought around the Earth at zero height and very slowly; these stipulations ensure that there can be no correction due to General Relativity (height over sea level) or to Special Relativity (speed). This correction is a nonsense.

The President of the organisation in Paris which oversees these rules wrote to the author 'you are right stating that the Sagnac effect is not relativistic'. That is an honest answer.

For the sending of signals from one site to another, their rules work fine. Saying the correction is 'relativistic' is a misnomer. The rules are wrong in the case of physical transportation of a clock from one site to another, but that is very rarely done, and can conveniently be overlooked.

There is also a correction applied to the clocks that ride on a satellite, to keep them in synchronisation with clocks on the ground. This is also supposed to be a relativistic correction. But, there is virtually no relative motion between the satellite clock and the ground station; the only relative motion is caused by the slight variation in the orbit of the satellite from an ideal orbit.

The correction applied is huge; it can amount to as much as 7500ns per day in a typical case; the clock is preset to alter by this amount each day, so that it will keep the same time as the clock fixed to the ground. It is calculated from the *absolute* velocity of the satellite compared with the *absolute* velocity of the ground station clock, in relation to the centre of the Earth, as it orbits around the Sun.

This correction is due to this absolute velocity, and not to the relative velocities of the satellite clock and the ground clock.

A very simple assumption would fit all of the these necessary corrections. If we assume that the signals travel with the Earth, on its orbit around the Sun, but do not adapt to the daily spin of the Earth, this fits the facts. In this case, the absolute velocity of the satellite versus the ground station accounts for the correction applied to the satellite

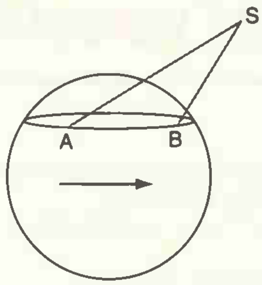

Fig. 2. Synchronising clocks on Earth. A and B are ground stations, while S is a satellite.

clock; the c±*v* of the signals sent around the Earth is explained simply by the spin of the Earth affecting the speed of the signal.

What could cause the signals to behave in this fashion? If light and gravity went together, on the Earth's orbit around the Sun, then the result would be fully explained.

We must then take it that the Special Theory of Relativity is not correct and that time and space are absolute, not relative. The speed of light is no longer sacrosanct.

But, what about the many many experiments that fit Relativity theory? A thousand things may fit a theory but, if one fact does not, the theory fails. In the words of Huxley 'the great tragedy of science – the slaying of a beautiful hypothesis by an ugly fact'. ■

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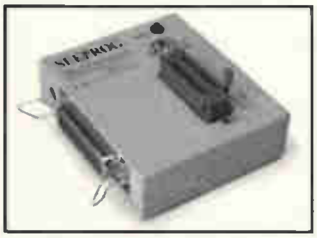

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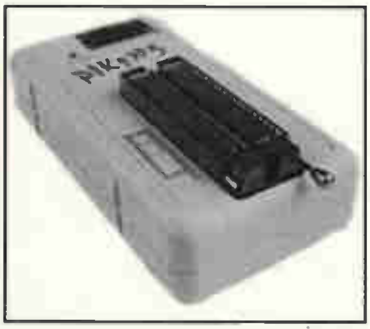

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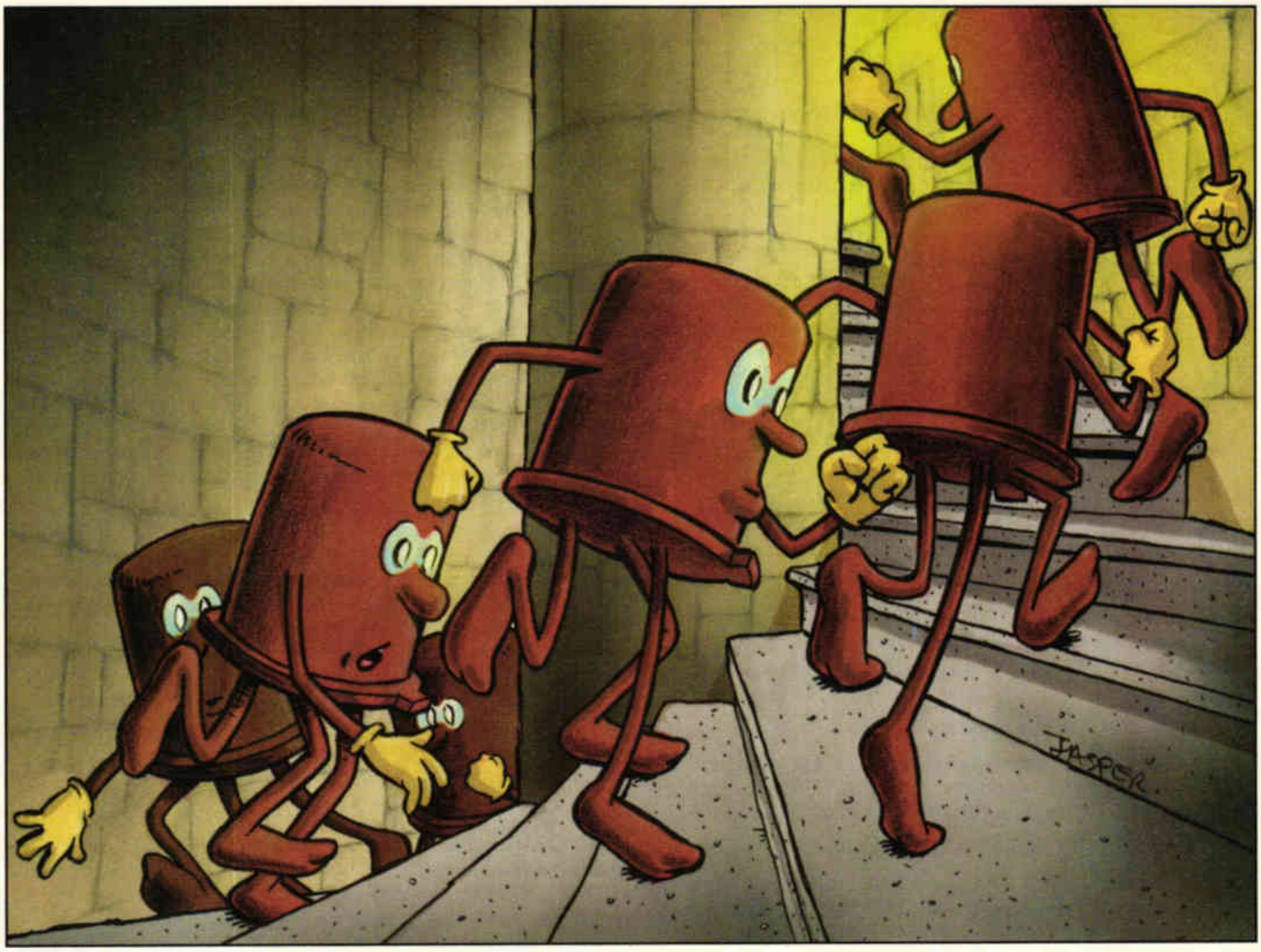

BETTER BUFFERS

Having been unsuccessful in looking for information to help him design complementary compound emitter followers, Dave Kimber set about developing his own guidelines. These include easy-to-use approximations for distortion components. This first article is an introduction to the subject and a discussion of how to predict Class-A followers at currents up to audio power-amp levels.

The complementary compound emitter follower is a useful building block. When compared with the ordinary emitter follower, it can have higher input impedance, lower output impedance and lower distortion, so it makes a superior buffer. In addition, there is a variant that can provide some voltage gain.

Although the textbooks mention this circuit block, not much guidance is given on suitable component values. I discovered this when designing an FM tuner. I wanted to use a complementary compound emitter follower – CCEF from now on – as a buffer for the birdy filter between the FM discriminator and the stereo decoder. I found that I didn't know how to pick component values; then I found that none of the books I had was much help either.

I was also prompted to investigate CCEF when reading Self's book on audio power amplifier design[1]. He found that a complementary pair of CCEF made an excellent Class-B output stage, but with a quiescent current set at what seemed to me a very low figure in the region of 7–15 mA.

Everybody 'knows' that quiescent currents in Class B output stages are somewhere between 30 and 100mA and relatively non-critical. It turns out that while this may be the case for a Darlington-pair output it is certainly not true for a CCEF output. Self also found that the quiescent current set-

David P. Kimber B.Sc.

ting was quite critical if crossover distortion is to be minimised.

If you are not familiar with the emitter follower, there's more background in the panel below.

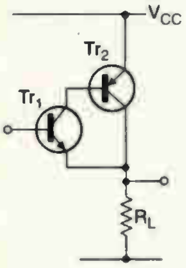

Fig. 1. The simplest complementary compound emitter follower is a basic emitter follower with a transistor added. Input impedance is much higher, but output impedance and distortion remain the same.

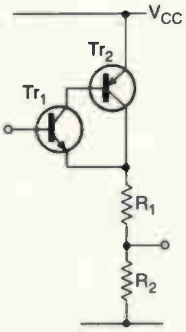

Fig. 2. This variant of the complementary compound emitter follower provides voltage and current gain.

Simple complementary compound emitter follower

The simplest CCEF just adds a second transistor, **Fig. 1**. Total current gain is then roughly $\beta_1 \times \beta_2$, which provides a much higher input impedance.

However, other things being equal, the simple CCEF does not reduce output impedance or distortion. This is because the input transistor, Tr_1, runs at a reduced current, approximately I_{out}/β_2, so has much lower g_m. This counteracts the current gain provided by the output transistor Tr_2.

However, by running at a higher current, the output impedance can be reduced while still leaving the input impedance above what it would be for a plain emitter follower.

A variant of the simple CCEF is the simple complementary feedback pair, **Fig. 2**, which provides voltage gain.

The emitter follower

The basic emitter follower, **Fig. A**, can be analysed as follows. This analysis assumes that the transistor follows the Ebers-Moll model, which relates collector current I_c to base-emitter voltage V_{be}, and can be written as:

$$I_c = \exp\left(40\left(V_{be} - V_o\right)\right) \quad \text{(mA)} \tag{1}$$

where V_o is the base-emitter voltage which produces a collector current of 1mA, and is normally somewhere around 0.6V. Collector current I_c is in milliamps. Differentiating this gives the transconductance, or g_m:

$$g_m = 40\exp\left(40\left(V_{be} - V_o\right)\right) = 40 \times I_c \quad \text{(mA/V)} \tag{2}$$

Another way of looking at this is that a small change in V_{be} produces a change in I_c which, is equal to the current change that voltage change would cause across a small resistor r with a value given by:

$$r(\Omega) = \frac{25}{I_c} \tag{3}$$

Here, I_c is in milliamps again.

For small signals you can treat a transistor as if it had an infinite g_m but with this small resistor r in the emitter circuit, **Fig. B**. Now the voltage gain of an emitter follower can be found by treating r and R_l as a voltage divider network:

$$gain = \frac{R_l}{R_l + r} \approx 1 - \frac{r}{R_l} \tag{4}$$

For maximum output voltage swing, you would normally arrange the quiescent conditions such that the output sits at about half the supply voltage, i.e.,

$$R_l(\mathrm{k}\Omega) = 0.5\frac{V_{cc}}{I_c} \tag{5}$$

Substituting equations (3) and (5) into (4), you will find that the gain is given by

$$gain \approx 1 - \frac{0.05}{V_{cc}} \tag{6}$$

This will be reduced if, as is usually the case, there is an additional AC load in parallel with the DC load. Output impedance is approximately equal to r, if the input signal source impedance is low.

So far we have ignored the base current I_b. For small signal transistors this is usually proportional to I_c,

$$I_b = \frac{I_c}{\beta} \tag{7}$$

The input impedance is approximately given by,

$$R_{in} = \beta \times R_l \tag{8}$$

Finally, you can estimate the distortion. The proportion of the input signal that appears across the base-emitter junction is, from equation (6), $0.05/V_{cc}$. It can be shown that for smallish signals, the percentage of second-harmonic distortion generated by a base-emitter junction is equal to the peak signal amplitude in mV. See reference 2 for more information on this.

The dominant distortion for the emitter-follower will be second-harmonic, although higher harmonics are likely to be significant too. The distortion will be reduced by the 100% negative feedback, to give a final figure,

$$Distortion\ (2nd\%) \approx 1.414 \times V_{sig} \times \left(\frac{0.05}{V_{cc}}\right)^2 \approx \frac{V_{sig}}{280 \times V_{cc}^2} \tag{9}$$

where V_{sig} is the RMS signal voltage in millivolts and V_{cc} is the supply voltage in volts. For a 2V signal and a 10V supply, this comes to 0.07%.

However, this increases quadratically as the AC load increases. For example, an AC load in parallel with a DC load of the same value will give four times as much distortion. This is because the base-emitter signal voltage is increased and the available negative feedback is reduced.

To achieve low distortion the DC load must be low compared with the AC load, i.e. the quiescent current must be high.

This treatment of the emitter follower may seem laborious, but it is a good introduction to the techniques used for the CCEF. The emitter follower can be improved in other ways – for example by introducing an active load – but this is outside the scope of this article.

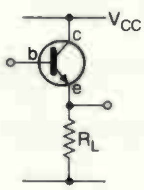

Fig. A. Basic emitter follower has no voltage gain and is widely used as a buffer, to prevent loading on the stage before it and increase its output current capability.

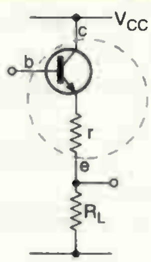

Fig. B. For small signals, it is possible to treat the transistor of an emitter follower as though it had infinite gain, provided you also assume that there's a low-value resistor, r, in series with the emitter.

Improving the CCEF

Adding a single resistor across the base and emitter of Tr_2, as in **Fig. 3**, brings much more flexibility. But this simple change makes analysis more complicated.

The main advantage is that Tr_1 runs increased current, so has a higher *gm* than for the simple CCEF. This reduces distortion, but at the expense of lower input impedance – although still higher than for the plain emitter follower.

The circuit can be considered to operate in three regions. In the low current region, the voltage drop across R is too low to turn on Tr_2 so effectively the CCEF degenerates into a plain emitter follower consisting of Tr_1 only.

In the medium current region, both transistors are active, but the current in Tr_2 is sufficiently low that its input impedance does not unduly load R. In other words, Tr_2 is voltage driven.

At high currents, Tr_2's input impedance now swamps R, i.e. Tr_2 is current driven. Resistor R still boosts the current in Tr_1, so increases g_m, but this effect becomes less important as the total current rises.

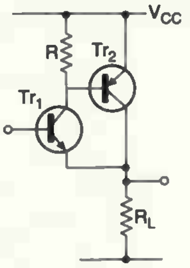

Fig. 3. Adding a resistor across the base and emitter of Tr_2 increases the operating current of Tr_1, and with it, the transistor's g_m.

A Class-A CCEF will operate in the medium current region or the lower end of the high current region.

A Class-B CCEF – to be considered in a third article on this topic – will operate in all three regions. In each case the critical feature is the effective transconductance, because this determines distortion and output impedance.

Low-current region. Tr_2 begins to turn on when its base-emitter voltage reaches somewhere around 600mV. Thus the low current region is the range 0 to $600/R$ mA. Transconductance, g_m, is given by equation (2) in the separate panel, and so increases linearly with current from 0 to approximately $24\,000/R$ mA/V.

Medium-current region. This region is bounded at the lower end by the low current region at $600/R$ mA. The transition to the high-current region can be considered to occur when the input impedance of Tr_2 is equal to R. This happens at,

$$I_{c2(m-h)} = 25 \times \frac{\beta_2}{R} \quad \text{(mA)} \qquad (10)$$

The total current is Ic_2 plus Ic_1, but Ic_1 will change only slightly from the current at the top of the low current region. This is because it only has to provide some base current for Tr_2 and allow for a small rise in the voltage across R. Thus the size of the medium current region, expressed as a ratio of the total current, is,

$$\frac{(25 \times \beta_2 + 600)}{600} = 1 + \left(\frac{\beta_2}{24}\right) \qquad (11)$$

The transconductance in this region has two contributions. One comes directly from Tr_1 and is approximately $24000/R$ mA/V, as for the top of the low current region. The second contribution is the transconductance of Tr_2 multiplied by the voltage gain of Tr_1 and R – with care over units!

$$g_{m(eff\,Tr_2)} = 40 \times I_{c2} \times \frac{24000}{R} \times \frac{R}{1000} \qquad (12)$$

$$g_{m(total)} = \frac{24000}{R} + 960 \times I_{c2} \quad \text{(mA/V)} \qquad (13)$$

At the upper end of the region Tr_2 begins to load R which reduces the gain. A better approximation is then,

$$g_{m(total)} = \frac{24000}{R} + \frac{960 \times I_{c2}}{\left(1 + \frac{I_{c2}}{I_{c2m-h}}\right)} \quad \text{(mA/V)} \qquad (14)$$

where $I_{c2(m-h)}$ is the current for the medium-high transition, given by equation (10).

Low-medium transition. It is instructive to compare the transconductance just below and just above the low-medium transition. Assume R equals 1kΩ, which might be a suitable value for a small signal CCEF. Then the transition occurs at 0.6mA. For $g_{m(0.5mA)}$, I_{c1}=0.5mA and I_{c2}=0mA, while for $g_{m(1mA)}$, I_{c1}=0.6mA and I_{c2}=0.4mA,

$$g_{m(0.5mA)} = 40 \times 0.5 = 20\text{mA/V} \qquad \text{using eqn (2)}$$

$$g_{m(1mA)} = \frac{24000}{1000} + 960 \times 0.4 = 408\text{mA/V} \qquad \text{using eqn (13)}$$

Thus a ratio of 2 in current has given rise to a ratio of 20 in g_m. This transition has a very sharp knee. I will be returning to this issue in the third article.

Note that when I_{c1} is 0.5mA, Ic_2 will actually be about 18μA. This adds 17.6mA/V to the 20mA/V calculated above, so the approximations are not too good just below the transition.

High-current region. Current in Tr_1 still has not increased very much at the lower end of this region. Transconductance in this region is dominated by Tr_2 current gain amplifying g_m of Tr_1. Transconductance is still given by equation (14), although it can be rewritten as,

$$g_m(\text{mA/V}) = \frac{2400}{R} \times \left(1 + \frac{\beta_2}{1 + \frac{I_{c2(m-h)}}{I_{c2}}}\right) \qquad (15)$$

This makes the mechanism more explicit. Equations (14) and (15) really are equivalent, although they look very different.

For higher currents it is necessary to add in Ib_2 to Ic_1:

$$g_m(\text{mA/V}) = 40 \times \left(\frac{600}{R} + \frac{I_{c2}}{\beta_2}\right) \times \left(1 + \frac{\beta_2}{1 + \frac{I_{c2(m-h)}}{I_{c2}}}\right) \qquad (16)$$

For very high currents, Ic_2 is much greater than $\beta_2 \times 600/R$. This can be reduced to the approximation,

$$g_m(\text{mA/V}) \approx 40 \times I_{tot}$$

which is the same as the equation for very small currents!

Thus the effect of R is to greatly boost transconductance in the middle region by increasing the current through Tr_1. But at very low and very high currents the simple formula for an emitter follower applies.

Medium-high transition. With the medium-high transition, as before, assume R is 1kΩ and also assume β_2 is 240. Then from equation (10), the medium-high transition occurs when I_{c2} is 6mA. Assume total currents of 5mA and 10mA.

$$g_{m(5mA)} = 24 + 960 \times \frac{4.4}{1 + \frac{4.4}{6}} = 2461\text{mA/V}$$

$$g_{m(10mA)} = 24 + 960 \times \frac{9.4}{1 + \frac{9.4}{6}} = 3540\text{mA/V}$$

In the first instance, I_{c1}=0.6mA and I_{c2}=4.4mA, while in the second, I_{c1}=0.6mA and I_{c2}=9.4mA.

This transition is much softer than the low-to-medium one. A ratio of 2 in current has given rise to a ratio of only 1.4 in transconductance.

Figure 4 shows the overall picture. There is a sharp knee into the medium current region, with steeply rising transconductance. This then gradually moves into the high current region where transconductance slowly rises until the simple emitter follower has caught up with it again.

Current gain of Tr_2 determines the size of the boost while R sets the position of the boost.

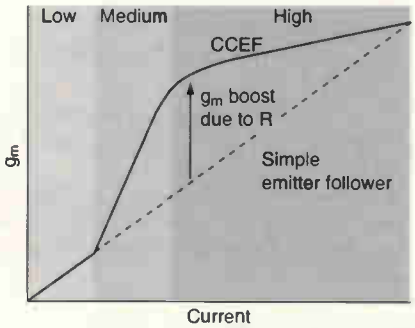

Fig. 4. There's a knee in the medium-current region, which then rises slowly into the high-current region, where the simple emitter follower catches up with the CCEF.

Distortion

Now we can consider signal distortion. For low distortion, the transconductance of the CCEF should be high when compared with the reciprocal of total load impedance, and should not vary much with current. So it is important to steer well clear of the low-medium transition and the lower part of the medium current region if low distortion is required. This is because in these areas g_m is low and strongly dependent on current.

As for the simple emitter follower, distortion will be mainly second harmonic but with higher harmonics not far behind. To estimate distortion you can use two tricks.

The first is Baxandall's 'reverse distortion' method[3]. If distortion is not too high, then the distortion produced by an amplifier fed with a distortionless signal is approximately equal – phase reversal aside – to the pre-distortion required in the input signal in order to generate a distortionless output.

This trick is useful when, as is the case here, we have a means of calculating the required input to an amplifier in order to achieve the desired output but the converse would become intractable.

The second trick is that the amount of second-harmonic distortion can be estimated if the gain of the amplifier at the signal peaks can be calculated. There's more on this in the panel entitled, 'Estimating second-harmonic distortion'.

I'll continue to use the same example CCEF, with R=1kΩ, β_2=240, and as before a supply of 10V with the quiescent output at half this voltage. For a current of 5mA, R_l is then 1kΩ. A 2VRMS signal would swing the current from 2.17mA to 7.83mA if no distortion was present.

You can find the transconductance for these two currents, then estimate the distortion,

Estimating second-harmonic distortion

Assume that the gain of an amplifier varies linearly with input voltage i.e.,

$$gain = gain_0 + a \times V_{in}$$

Then the amplifier will generate pure second harmonic distortion. Integrating the above gives,

$$V_{out} = gain_0 \times V_{in} + \frac{1}{2} \times a \times V_{in}^2$$

plus a constant, which can be ignored. Then if a sinusoidal signal with peak voltage V_{pk} is applied, the output positive and negative peak voltages will be,

$$V_{out+(pk)} = gain_0 \times V_{pk} + \frac{1}{2} \times a \times V_{pk}^2$$

$$V_{out-(pk)} = -gain_0 \times V_{pk} + \frac{1}{2} \times a \times V_{pk}^2$$

The first term in each case is the fundamental signal, amplified by the gain, and the second term contains equal amounts of the second harmonic distortion and a DC level shift.

$$V_{out(fund)} = gain_0 \times V_{pk}$$

$$V_{out(2nd)} = \frac{1}{2} \times \frac{1}{2} \times a \times V_{pk}^2$$

As a fraction:

$$distortion_{2nd} = \frac{1}{4} \times a \times \frac{V_{pk}^2}{gain_0 \times V_{pk}}$$

You can rewrite this in terms of peak gains by noting that,

$$gain_- = gain_0 - a \times V_{pk}$$

Then

$$a \times V_{pk} = \frac{1}{2} \times (gain_+ - gain_-)$$

$$gain_0 = \frac{1}{2} \times (gain_+ + gain_-)$$

So,

$$distortion_{2nd} = \frac{1}{4} \times \frac{(gain_+ - gain_-)}{(gain_+ + gain_-)}$$

This result is exact if gain varies linearly with input voltage. It provides a useful estimate if this is not the case, but second harmonic distortion is the dominant one. It breaks down completely if higher even harmonics are dominant. I am sure this result is not original, but I don't recall seeing it in print before.

If third harmonic distortion is dominant then an analogous result is,

$$distortion_{3rd} = \frac{1}{12} \times \frac{(gain_+ - gain_0)}{gain_0}$$

If both second and third harmonics are present, but higher ones are smaller, then these results can still be used but in a modified form,

$$distortion_{2nd} = \frac{1}{8} \times \frac{(gain_+ - gain_-)}{gain_0}$$

$$distortion_{3rd} = \frac{1}{12} \times \left(\frac{\left(\left(\frac{gain_+ + gain_-}{2}\right) - gain_0\right)}{gain_0} \right)$$

$$g_{m-(2.17mA)} = 24 + 960 \times \frac{1.57}{1 + \frac{1.57}{6}} = 1219\text{mA/V}$$

$$g_{m+(7.83mA)} = 24 + 960 \times \frac{7.23}{1 + \frac{7.23}{6}} = 3172\text{mA/V}$$

$$distortion_{(g_m)} = \frac{1}{4} \times \frac{g_{m+} - g_{m-}}{g_{m+} + g_{m-}} = 0.111$$

$$distortion_{(circuit)} = \frac{0.111}{g_{m0} \times R_1} = 0.005\%$$

This is over ten times better than the emitter follower. Most of the advantage comes from the higher g_m of the CCEF, but g_m is also changing more slowly with current.

Adding an AC load would have the same effect as for the emitter follower, i.e. the distortion would vary roughly as the square of the ratio of the DC load to the total load.

Circuit design

We now have the necessary knowledge to return to the original question. How can we choose suitable component values for a CCEF, given the environment – e.g. supply voltage – and requirements – e.g. signal level, load?

The first step is to work out the peak signal current and choose a Tr_2 quiescent current that is a little higher than this, so we keep clear of the low current region. At this stage it may be worth doing a quick calculation to see if an ordinary emitter follower can provide sufficient performance.

The value of the DC load resistor is found using Ohm's law. This leaves R to be chosen. The best value would seem to be the one that maximises the transconductance boost, and hence minimises distortion and output impedance. This means placing the operating point somewhere in the region of the medium-high current transition.

Unfortunately this depends on the current gain of Tr_2, which is a poorly controlled parameter, but fortunately this transition has a smooth knee so accuracy is not too important. So, we invert equation (10),

$$R = 25 \times \frac{\beta_2}{I_{c2}} \tag{17}$$

Here, R is in ohms and I_{c2} in milliamps.

Slightly higher values of R may reduce distortion because the operating point is moved into the low end of the high-current region, where transconductance changes only slowly with current. Slightly lower values of R may help high-frequency performance.

Given R, the performance of the circuit can be calculated. If not quite good enough, try an increase in quiescent current and recalculate. If the CCEF still does not meet the requirements, then at least you have pushed it as far as it will go.

My second article considers the complementary feedback pair, as a special case of the CCEF. A third article will look at using the CCEF in a Class B output stage. ■

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1. 'Audio Power Amplifier Design Handbook,' Douglas Self, Newnes 1996, Chapter 5.
2. Baxandall, Peter J, 'Audio power amplifier design,' *Wireless World*, February 1979, p. 72.
3. Baxandall, *op cit*, p. 71.

Beginners' corner

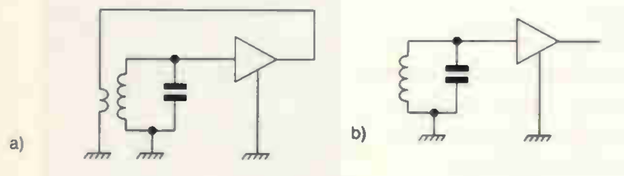
Design your own Colpitts oscillators

Radio-frequency oscillators play an important role in wireless communications. One is required in every transmitter, and there is also at least one in most receivers. Here, as an aid to giving students and beginners practical experience of building circuits, Ian Hickman describes one of the widely used types – the Colpitts oscillator.

White light contains a wide range of visible frequencies, but yellow light from a sodium lamp contains a much narrower range. A laser, on the other hand produces monochromatic light, containing just a single frequency component.

In a similar way, the earliest wireless transmitters, using a spark gap as the source of radio frequency energy, produced a broad band of frequencies. This was soon reduced to a rather narrower band using a resonant circuit, more analogous to the sodium lamp than white light.

Fig. 1. Two types of oscillator; a) using a maintaining amplifier, b) negative resistance oscillator.



As the art progressed, it was desirable to produce radio-frequency energy comprising just a single spectral line – analogous to the output of a laser. Subsequent modulation would then spread the energy over just the necessary bandwidth to communicate the message to be transmitted, rather than spraying it all over a wider range like a spark transmitter.

So instead of producing broadband energy, and then using a tuned circuit to select just a part of it, oscillators using valves were invented. Here, a continuous oscillation in a tuned circuit was maintained by an active device – the valve – which made up the inevitable losses due to dissipation in the tuned circuit.

If the valve were disabled, the oscillation would die away to nothing in a few cycles, or tens of cycles at most.

Two sorts of oscillator

One way to make an oscillator is to connect a suitable RF amplifier to the circuit, and then feed its output back into the tuned circuit, in the same phase. If the gain around the circuit, from the input to the amplifier, to its output, through the tuned circuit and back again to its input, exceeds unity, then rather than dying away, the oscillation will build up in amplitude. Eventually, the amplitude will reach a stable limit, as the energy the amplifier can supply is itself limited, by its supply voltage and its standing current.

This type of oscillator is illustrated in Fig. **1a)**. Depending on whether the amplifier is inverting or non-inverting, the coupling winding used to feed its output energy back into the tuned circuit will have to be connected one way round or the other, to ensure that the feedback is positive. Thus the amplifier's output is a replica of the oscillating voltage in the tuned circuit, and continually makes up the losses.

These losses are due to the finite Q – or quality factor – of the capacitor, and more particularly the inductor. The Q of the latter may be a few hundred at most, but probably less than a hundred, while that of the capacitor will usually be ten or more times better. Operating Q will also be affected by the loading of the active device's input and output impedances upon the circuit.

There is another sort of oscillator, as illustrated in Fig. **1b)**. At first glance, it looks as though the circuit can never

In this series

As explained in a preliminary article in the May 2000 issue, this series is intended to help students – and anyone interested in getting to grips with RF design – as a background in practical electronic circuitry and troubleshooting.

The series was originally developed in response to the government's RF Engineering Education Initiative. Below is a list of the three tutorials that have already appeared, together with my plans for future articles in the series – 'Beginners' corner'.

- Timer circuit using the 555, June issue.
- Audio oscillator – Wien bridge based, July issue.
- h_{fe} tester, August issue.
- Radio-frequency oscillator, Colpitts type.
- Audio filter and oscillator – state-variable based.
- Capacitance meter.
- Radio-frequency oscillator/receiver involving negative resistance.

work, as there is no way of coupling the output of the oscillator back into the tuned circuit. But it can work, if the amplifier is suitably designed.

The trick is to make the impedance, seen looking into the amplifier's input, a negative resistance. This then cancels output losses in the tuned circuit, resulting in a stable level of oscillation, of amplitude determined by the particular circuit design.

This type of oscillator can be very useful, and will be featured in a later article in this series. But for the present, let's look at a version of the Fig. 1a) variety.

Amplifier-maintained oscillation

The circuit of Fig 1a) uses a coupling winding to inject the 'make-up' energy from the amplifier, back into the tuned circuit.

Designers often prefer a circuit arrangement that dispenses with the separate winding, for a number of reasons. Taking technical considerations first, an RF oscillator does not usually use a closed flux path for the inductor. It may be purely air cored, or have a 'tuning slug' of ferrite usually, dust iron, or sometimes brass.

Either way though, the coupling between the two windings will be somewhat unpredictable. It is not simply determined by the turns ratio, as would be the case in a transformer with a closed flux path, such as a mains transformer. And from the point of view of practical economics, both component and labour costs are increased.

So an arrangement where both input and output of the maintaining amplifier are directly connected to the tuned circuit will often be preferred. Now the amplifier is a three-terminal device; input, common and output, so there must be three connection points on the tuned circuit. Consequently, a single capacitor and inductor, as in Fig. 1, will not suffice, and the arrangement of **Fig. 2** is employed.

With three connection points available on the tuned circuit, an active device, be it valve, transistor or FET, can be connected as shown. An interesting property of this circuit is that Z_2 and Z_3 must be impedances of the same type, either both capacitors or both inductors, while Z_1 must be the other type. Furthermore, the common terminal of the active device, cathode, emitter or source, must be connected to the junction of Z_2 and Z_3.

If Z_2 and Z_3 are inductors – or, more likely a single, tapped inductor – the circuit is known as a Hartley oscillator, while if Z_2 and Z_3 are capacitors, it is known as a Colpitts oscillator. The Colpitts arrangement is often preferred, as an untapped inductor is cheaper to make, test and assemble than a tapped component.

The Colpitts oscillator

Within the basic 'tapped capacitor' arrangement – actually two separate capacitors – several variants are possible. **Figure 3a)** shows one of these.

The base of the transistor is referenced to ground DC-wise, so the average emitter current is determined by the value of R_1 and the voltage of the negative rail. Effectively, the large decoupling capacitor C_3 connects the collector to the junction of L_1 and C_2, exactly as in Fig. 2, where Z_1 is the inductor L_1 and Z_2 the capacitor C_2.

The amplitude of the oscillation builds up, resulting in the transistor being cut off for much of each cycle, due to a bias voltage being built up across C_2. Collector current thus flows in narrow pulses, the transistor operating in 'class C'. The fundamental component of these pulses, at the resonant frequency of the tuned circuit, maintains the oscillation.

Making an oscillator useful

It is all very well making an oscillator, but to be useful, it must be possible to bleed off some of the signal, for use say as a local oscillator in a superhet receiver, or as the exciter in a transmitter.

The signal should be bled off in such a way as not to impose excessive further loading on the oscillator. If it does load the oscillator, the Q of the tuned circuit would be reduced, degrading the oscillator's stability and purity.

Various pick-off arrangements are possible. One possibility is to draw off a little current from the emitter, via a small capacitor of a few picofarads, or a single turn coupling winding on L_1 could be used. Another popular arrangement is to draw off the signal across a large capacitor in series with the earthy end of C_2. In the case of Fig. 3, where the oscillator runs at about 5MHz, its value might be a few nanofarads.

Figure 3b) shows another variation on the theme, this time requiring only a single supply rail. With the emitter dc referenced to ground via L_2, base current is supplied via R_1. Capacitor C_3 couples the tuned circuit to the base of the transistor, while preventing the current via R_1 being simply shunted to ground via L_1.

At the operating frequency, the reactance of L_2 is very high. So as far as the emitter circuit is concerned, it is, like R_1 in Fig. 3a), virtually an open circuit – just a convenient way of supplying the emitter current.

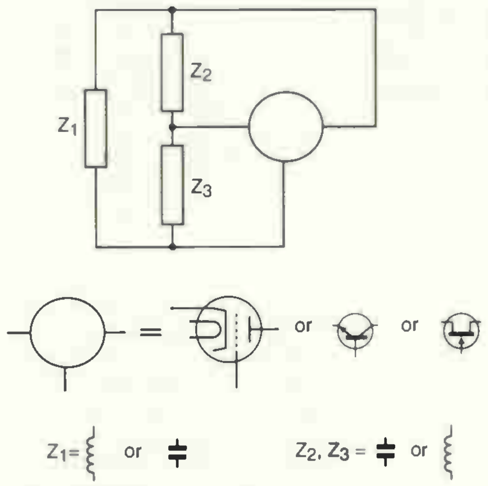


Fig. 2. Varieties of Fig. 1a) type oscillator based on a maintaining amplifier. No separate feedback winding is involved here.

Inductor L_2 can be a commercial RF choke, as can L_1 – at least for experimental purposes.

Magnitude of the emitter current is determined by R_1, the current gain of the transistor and the voltage of the positive rail. In fact, if R_1 is low enough and the Q high enough, the amplitude may increase to the point where on positive peaks, the base voltage actually rises above the positive rail. This is not a desirable condition, as severe damping is applied to the tuned circuit, to the detriment of the purity of the waveform and the stability of the frequency.

Given the considerable device-to-device variation in current gain h_{fe}, the circuit of Fig. 3b) is thus not as 'designable' as that in a). There, the average emitter current is determined principally by the value of R_1 and the voltage of the negative rail.

Emitter current can be set at a level such that the transistor does not bottom, removing one of the causes of close-in phase noise and resulting in a purer output waveform. The advantages of this were appreciated long ago[1].

Might it squeg?

Choosing the value of C_3 in Fig. 3b) also needs care. If it is too small, the circuit will not oscillate; too large and the circuit will 'squeg'.

Squegging is when the amplitude rises up so fast that a negative voltage builds up on the base end of C_3, cutting off the transistor completely. The oscillation across the tuned circuit dies away, but the base of Tr_1 is left at a negative voltage, charging only slowly towards +15V via R_1, on the time-constant $R_1 \times C_3$.

When the base reaches about +0.6V,

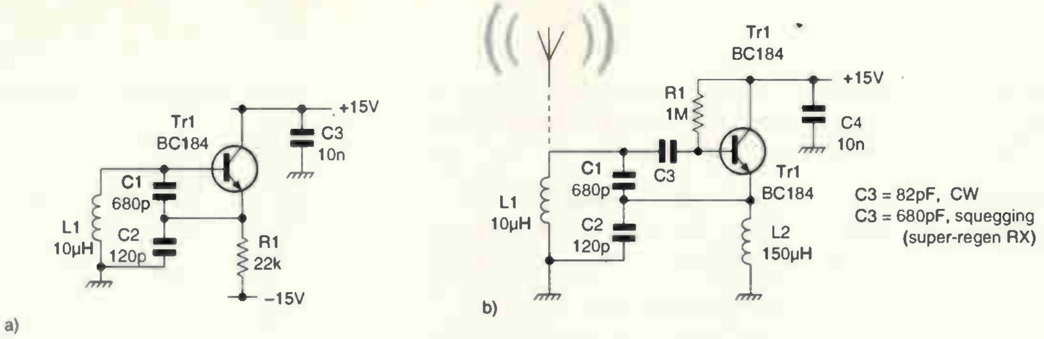


Fig. 3a). A Colpitts oscillator, and b), another variant, which may – or may not – squeg.

the transistor begins to conduct, the RF oscillation rapidly builds up again, resulting in the next cycle of the squeg frequency – a much lower frequency than the r.f. oscillation frequency.

A squegging oscillator forms the basis of a 'super-regenerative' receiver – a type of receiver that saw extensive military use in the Second World War. If a short whip antenna is connected to the tuned circuit, as indicated in Fig. 3b), then any energy incoming at the resonant frequency of the tuned circuit causes an increase in the squeg frequency. It also causes a corresponding small change in collector current.

This current change can be monitored by the drop across a resistor, connected in series with the +15V supply. Thus the received signal is demodulated and available at the collector of Tr_1 – you have a simple 'super-regen' receiver.

In Figs 3a) and b), the ratio of C_1 to C_2 is one of the many design choices. Generally, the ratio is between 2 to 1 and 5 to 1. Increasing the value of C_1, while reducing that of C_2 to maintain the desired oscillator frequency reduces the loading of the transistor's input circuit on the tuned circuit. It also demands a higher gain from the transistor.

If the ratio of C_1 to C_2 is made too large, the circuit will not oscillate. Conversely, making C_2 larger, and reducing C_1 to maintain the desired oscillator frequency reduces the loading of the transistor's output circuit on the tuned circuit.

Again, go too far, and the circuit will not oscillate. Thus there are limits to the ratio, at both ends of the range, as discussed in reference 2. ■

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1. Sutcliffe, H, 'Transistor L.C. oscillator circuits with amplitude controlled by mean current'. *Electronic Engineering* July 1968, Vol. 40 No 485, 0. 388.
2. Shahzadi, B. 'Two distinct boundaries for feedback transistor oscillators'. *Electronic Engineering* Jan. 1965, pp. 32-34.

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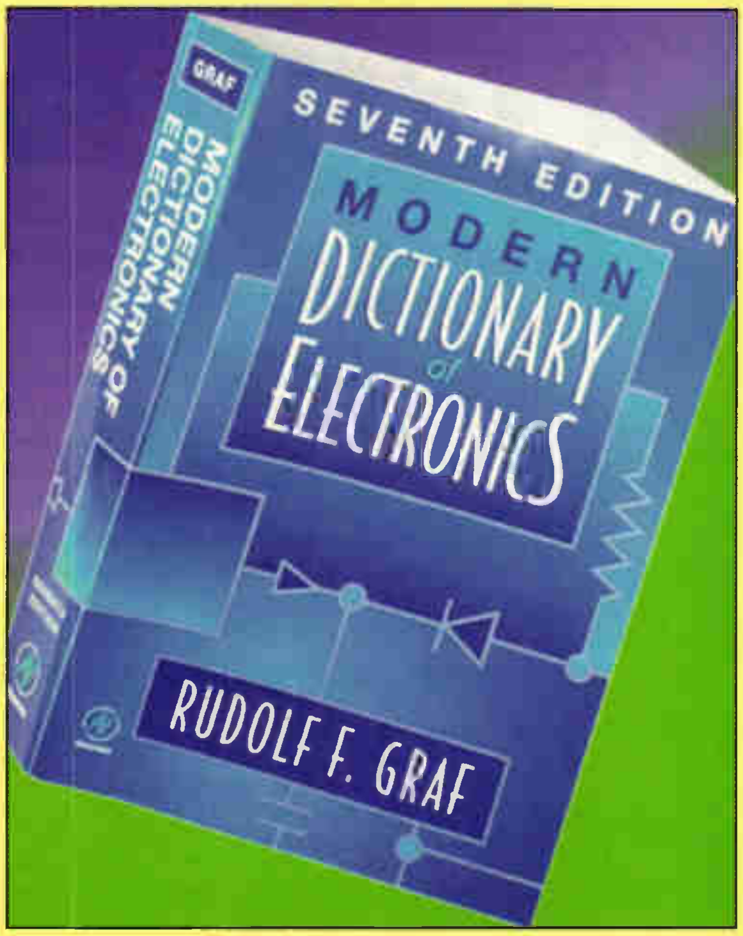

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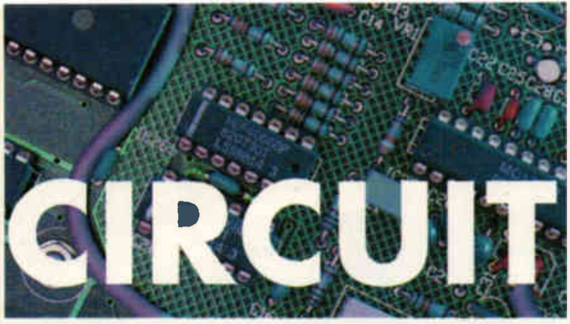

CIRCUIT IDEAS

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

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

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

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

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

Bike computer reads amps, amp.hours

A bicycle computer counts wheel revolutions, and displays speed and distance travelled. Depending on the model, it may also show the maximum and average speeds achieved.

The user has to program the computer with the wheel circumference C in metres, since velocity v in km/h is related to the frequency of wheel rotation, *Frps*, by $(v \div 3.6)/C=F$. Usually, a magnet attached to a wheel operates a reed relay to provide the count pulses, but in this application, a transistor switch is used.

Such a bicycle computer can be used for other purposes, such as measuring the charge rate and total charge stored in a solar panel accumulator charging set-up, **Fig. 1**. To achieve this, the charging current is monitored by a current shunt R_s, controlling a voltage-controlled oscillator.

The voltage controlled oscillator produces an output frequency such that a bicycle computer velocity reading of 120km/h indicates a current of 12A, and a trip reading of 2998.9 km indicates a charge of 299.89A/h.

The programmable value of C on the computer used was up to 2.999m. The VCO was designed to produce an output frequency of up to 13.7Hz for a 140mV input, corresponding to a 14A charging current. With this design of oscillator, **Fig. 2**, a circumference setting C of 2.671m worked well.

This application is limited by the lowest and highest frequencies that the bicycle computer can count, and by VCO offset and linearity errors.

A minimum output frequency of 0.1Hz is produced by the VCO, even when the drop across R_s is zero. But linearity errors up to the designed maximum, checked with a DVM and DSO, proved to be generally insignificant. There is a slight increase in error at the high frequency end of the range, due to the finite discharge time of C_1.

Heinz Zanke

GR-24002 Messenias

Greece

E10

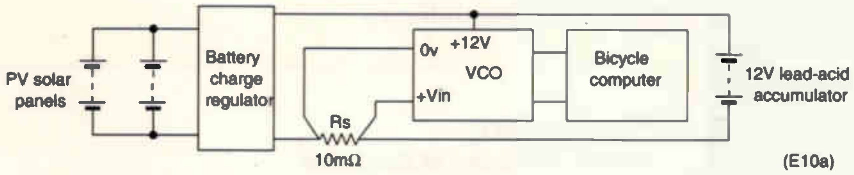

Fig. 1. Block diagram of charger metering system in a solar energy system, using a bicycle computer to monitor amps and A/h.

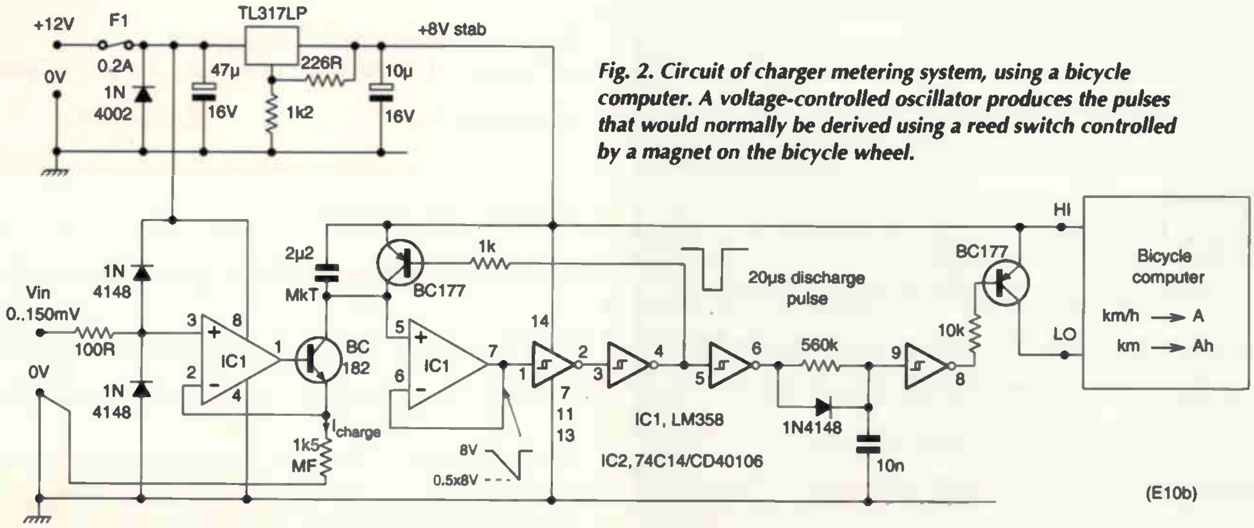

Fig. 2. Circuit of charger metering system, using a bicycle computer. A voltage-controlled oscillator produces the pulses that would normally be derived using a reed switch controlled by a magnet on the bicycle wheel.

One-at-a-time phones

Normally, where two or more phones share a single line, someone picking up another, when the first is in use, can overhear the other party. This circuit allows only one of the phones to be used at any given time.

One circuit must be connected in series with each phone. This idea relies on the fact that the off-hook line voltage is 48V, but only 10V on-hook. Incoming calls will ring all phones, as the ringing voltage is 80V RMS.

J M Brassart
Saint-Laurent-Du-Var
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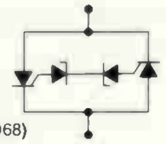
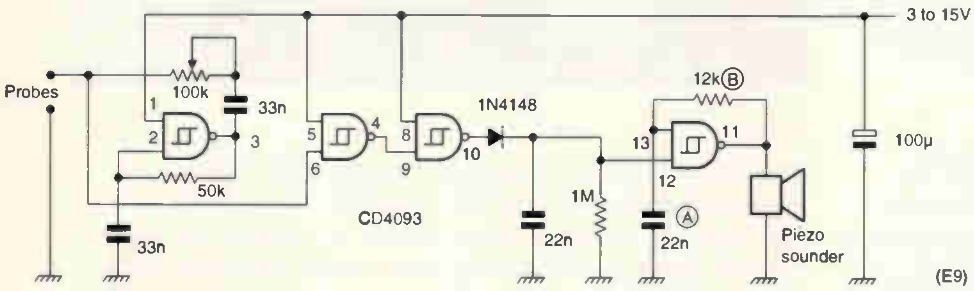
D68

In some countries, you are not allowed to connect any equipment that has not been formally approved by the service provider to telephone outlet – Ed.



Connected in series with each phone on a line, this circuit permits only one to be used at a time.

Economical liquid level sensor needs just 7µA



Low cost, long battery life and adjustable sensitivity are features of this liquid level sensor circuit. If you want a logic-level output, tie pin 13 to V_{cc} and omit components A and B.

This simple sensor circuit consumes only 7µA at 3V supply, and is ideal for battery operation. There is no dc through the sensing electrodes, minimising corrosion.

Sensitivity may be adjusted for different electrode sizes and fluid conductivities by means of the 100kΩ potentiometer. The circuit's cost is considerably less than the LM183, which is in any case currently unavailable.

David Tayler
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E9

Measuring femtofarads on traditional AC bridges

I needed to check the matching of two nominally 1pF capacitors, but all I had was a Marconi Instruments TF868B universal bridge having a lowest capacitance range of 0-100pF.

I could have made a transformer bridge, but this was not justifiable for the job in hand, given the extensive screening and constructional work required. The possibility of using the AC source and tuned detector in the TF868B was then considered; ideally without any alterations. Bridge resolution would have been adequate at 0.1 % at the top end of a range.

Figure 1 shows the basic bridge circuit for capacitance measurements; the earthing of one side of the detector allows three-terminal capacitance measurement to exclude strays. These are either across the detector or across C_2+R_1; so of insignificant effect at null.

If the voltage across C_2+R_1 were multiplied by a factor of, say 10, then a reduction in the unknown capacitor by a factor of 10 is needed to preserve bridge balance.

Amplification only need be applied to the voltage across the standard capacitor, since the magnified voltage is in phase with the original voltage across the capacitor, preserving phase balance to the detector, **Fig. 2**. No modifications are needed inside the TF868B.

Figure 3 shows the first circuit tried; to give a range of 0-10pF. The signal on 'Hi' never exceeded 1.2V pk-pk, so an op-amp with a gain of 10 could be used when driven from two 9V batteries in series; batteries were used to prevent earth loops and null offsets.

Op-amp IC_{1A} is a buffer – not strictly necessary – and R_5 protects the TL082 FET inputs against static build-up when the circuit is disconnected.

Components R_1, R_2 and C_1 gave a decoupled reference for the signal to swing about, R_4 and R_3 defining the gain. It worked as expected, giving a sharp null with a test 2.2pF capacitor at the 22pF dial reading.

Figure 4 showed the circuit tried to give 0-1pF range – with a resolution of 1fF at the top end; stray capacitances C_{S1} and C_{S2} are shown. I considered a transformer to be the easiest method for giving the voltage step-up needed. A tightly-coupled toroid transformer was found with the characteristics shown – and a self-resonant frequency well above the 1kHz of the bridge source.

Two op-amps – used as unity gain buffers – were connected in parallel to give the peak drive current of around 13mA. Individual and differential DC offsets, giving inter-amplifier and primary DC currents, were <1mV on the device used; should this be thought a problem capacitive coupling from the outputs to the transformer could be used at the risk of some phase shift; adding series resistance would also give significant phase shifts.

Supply current was around 11mA and the circuit worked well. Nulled 'tan-delta' setting was non-zero, due to phase shift from op-amp output impedance and primary inductance, but a sharp capacitive null was obtained.

Output voltage from the transformer was ≤120V pk-pk and at high impedance, but take care to avoid electric shock. Hand-capacitance effects were very significant and the high voltage output of the transformer secondary was placed at least 2 inches away from the 'Lo' terminal of the bridge. Even so, null readings of over 80fF were obtained, but use of a metallic screen between the high-voltage terminal and the 'Lo' terminal reduced this further. Low battery voltage is shown by a drifting null.

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D20

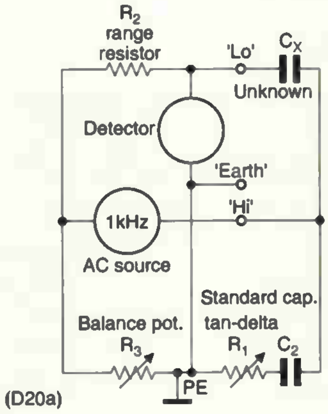

Fig. 1. Showing the basic bridge circuit for capacitance measurements.

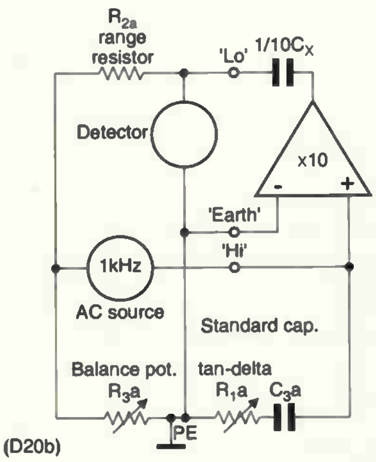

Fig. 2. Outline arrangement to extend the smallest range from 100pF maximum to 10pF maximum.

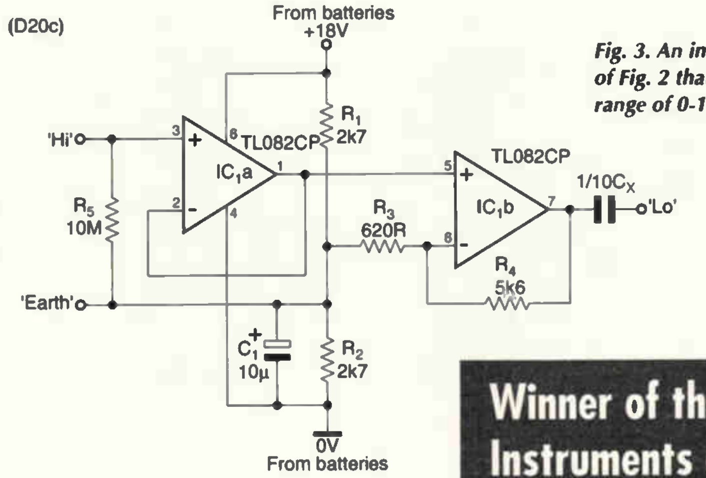

Fig. 3. An implementation of Fig. 2 that results in a range of 0-10pF.

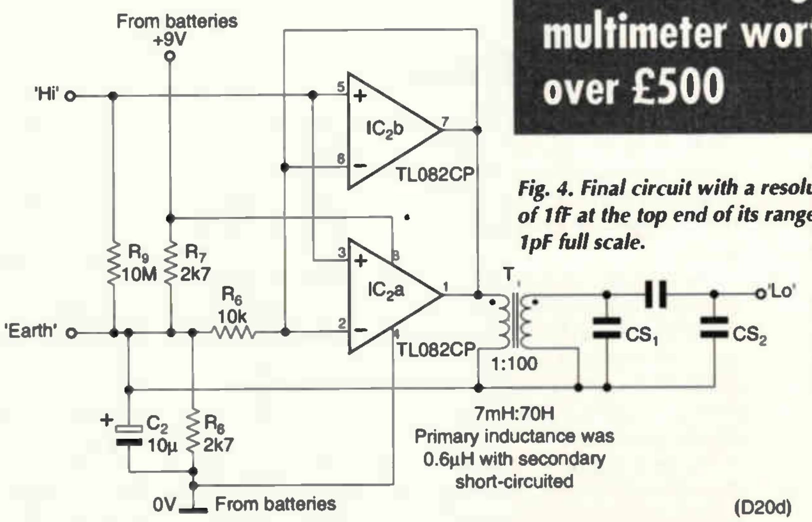

Fig. 4. Final circuit with a resolution of 1fF at the top end of its range and 1pF full scale.

£50 Winner

Simple charger for NiCd and lead acid batteries

This circuit is suitable for charging and discharging – i.e. refreshing – NiCd and lead-acid batteries. Despite being simple, it also promotes long battery life.

Charging current I_2 is constant. In switch position 'NiCd' the current flows for a period of 20ms, determined by R_7 and C_1. The next charge pulse will follow if the off-load voltage of the battery falls below 1.4V/cell.

Charging of NiCd cell with pulsed current has some advantages. Measuring the off-load voltage works well with old batteries having a higher internal resistance. The charging current used may be up to the C capacity of the battery; this permits a short charge time. On the other hand, the battery cannot be overcharged by a 20ms pulse.

With experience, the flash rate of the green LED can indicate the state of charge of the battery. With new and fully charged batteries, the period between flashes is several seconds. Use of high current pulses up to C avoids the memory effect in NiCd cells.

The semi-automatic discharge circuit is optional. Discharge is initiated by the push button, and the red LED lights, discharging the battery to, for example, 0.9V/cell. The circuit then automatically switches to the charge function.

In switch position 'Pb' for lead-acid batteries, a constant charge current is employed, until the battery voltage reaches, for example, 2.4V/cell. Current is then reduced and the green LED fades, indicating a change to constant voltage operation.

For correct operation, good battery contacts are required. In the event of a supply interruption, the battery will not be discharged.

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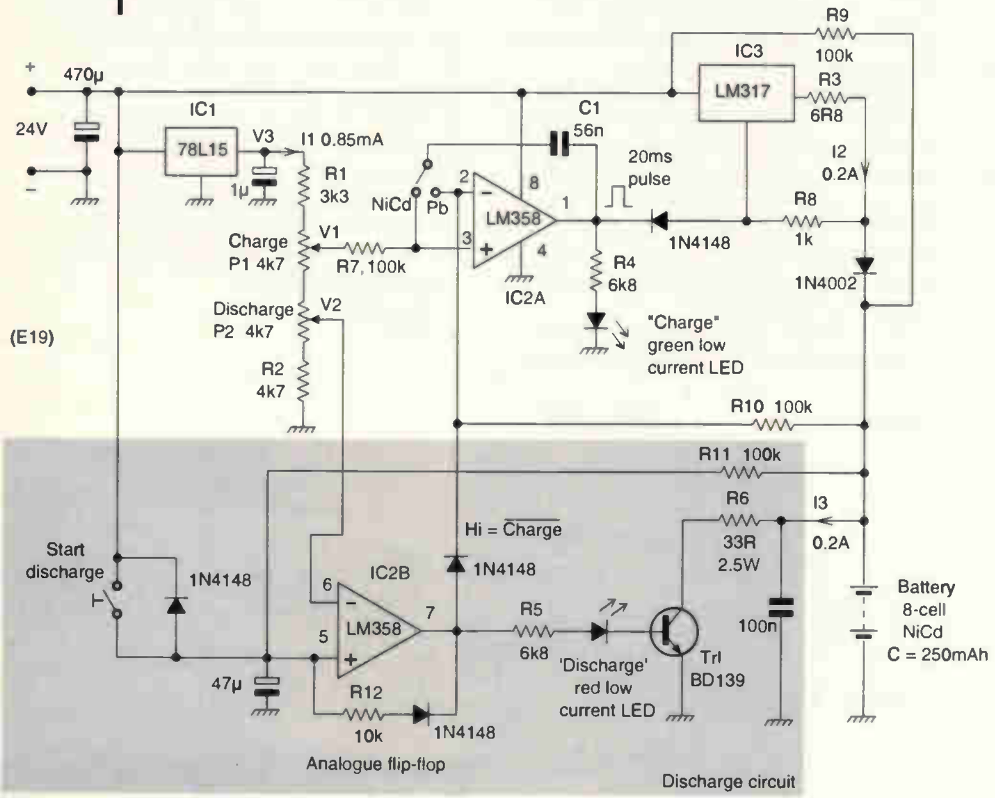
E19



This charger uses pulse charging for NiCds, constant current/constant voltage for lead acid. Supply voltage needs to be 3.4V higher than the highest expected battery voltage. Current I_1 is about a milliamp. Resistor R_3 is 1.35V/I_2. For lead-acid cells, V_1 is between 2.25 and 2.5V per cell and I_2 is less than C. For NiCd alternatives, V_1 is 1.4V per cell while V_2 is between 0.9 and 1.1V per cell. Current I_2 is roughly equal to both I_3 and C.

£50 Winner

Jingle softner

When watching TV, have you ever been annoyed by the increase in sound level during the advertisements? One way to deal with it is to press the mute key of the remote control unit. Another way is to plug this little unit into the Scart/Peritel socket at the back of your TV.

The schematic is given **Fig. 1**. The heart of this circuit is the MC3340P from Motorola, which is a voltage-controlled attenuator. This circuit offers a 80dB range attenuation when driven with a 3.5V control voltage range, but in order to obtain a reasonable THD (distortion) of less than 3%, the attenuation range must be limited to 40dB.

The input signal is reduced by a factor of 2 by resistors R_1 and R_2 because the maximum input voltage of IC_1 is 500mV. The signal attenuated by IC_1 appears at point B. The value of C_2 is not critical and limits the bandpass to 20kHz. Capacitors $C_1$and C_3 DC block IC_1. Values given for R_3 and C_3 produce a low-frequency cut-off close to 10Hz.

The signal at point B is amplified by IC_{2a} and P_1 permits the sound level to be adjusted as required. At point B, the signal follows a second path and is amplified by IC_{2b}. R_8 protects the output of IC_{2b}.

Components C_4, C_5, D_1 and D_2 form a doubler-rectifier, producing at point D a measure of the average value for the signal amplitude. The value of R_9 has been chosen to give a time constant big enough to avoid low frequency distortion. At point D, the voltage is equal to the average value of the signal amplitude at point C.

Op-amp IC_3 acts as a voltage follower and R_{10} and C_6 form an additional low-pass filter. The time constant is fixed at 50ms.

Voltage at point E controls the attenuation value of IC_1.

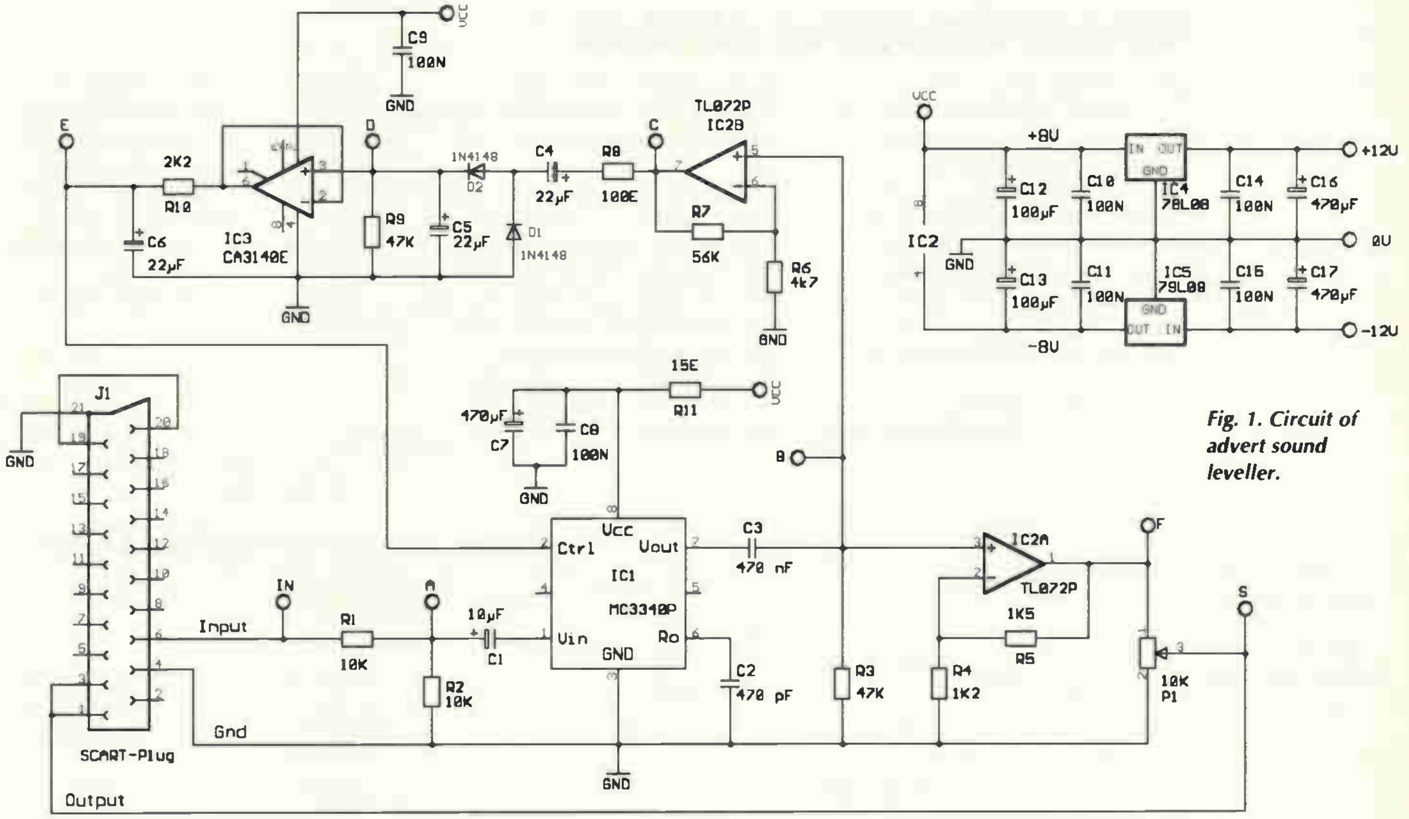


Fig. 1. Circuit of advert sound leveller.

When the sound becomes louder, voltage at *D* increases, increasing the attenuation. Thus the sound level at B remains constant.

The functional block diagram is given in **Fig. 2** and the attenuation curve of the MC3340 in **Fig. 3**. An 8V supply is used, so the attenuation curve gives a 40dB/V attenuation factor for control voltages greater than 2V.

Gain variation of the device can be derived from the curve,

$$G = -40\times(V_d - 2) = 20\log.$$

Static gain for IC_1 is 13dB and amplification is 4.5x.

When V_e is under 2V, IC_1's amplification value is close to 4 and V_b is $4\times V_a$. As $V_e = V_{c(max)}$ and V_c is $13\times V_b$, you can work out, $V_e = 13\times4.5\times V_{a(max)} = 58\times V_{a(max)}$.

When V_e is less than 2V, $V_{a(max)}$ is less than 35mV and $V_{in(max)}$ is less than 70mV (50mV RMS). Thus, when V_{in} is under 50mV RMS, the circuit doesn't limit and the value of V_f is given by the linear expression.

$$V_f = 0.5\times4.5\times2.25\times V_{in} = 5\times V_a$$

Gain is 14dB.

When V_{in} is greater than 50mV RMS, V_e increases in order to maintain V_b at the same value. Voltage V_f keeps its former value. $V_f = 5\times V_{in}$=250mV RMS. Voltage V_s can be adjusted between 0 and 250 mV RMS.

Note that if another value of the starting point for limitation, currently 50mV RMS, is needed, this can be achieved by changing the values of R_7 and/or R_6.

$$\frac{R_7}{R_6} = \frac{2000}{4.5\times0.5\times V_{in(RMS)}\times1.14} - 1$$

J M Terrade
63100 Clermont-Ferrand
France
D51

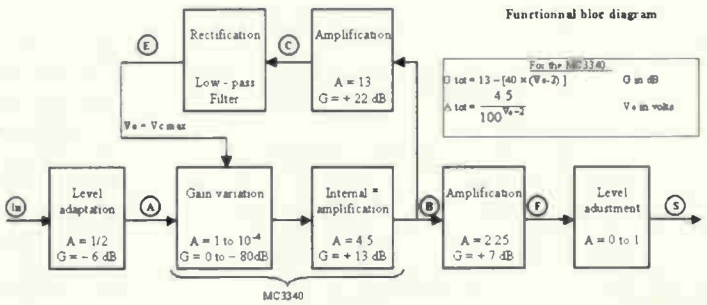


Fig. 2. Block diagram of advert sound leveller.

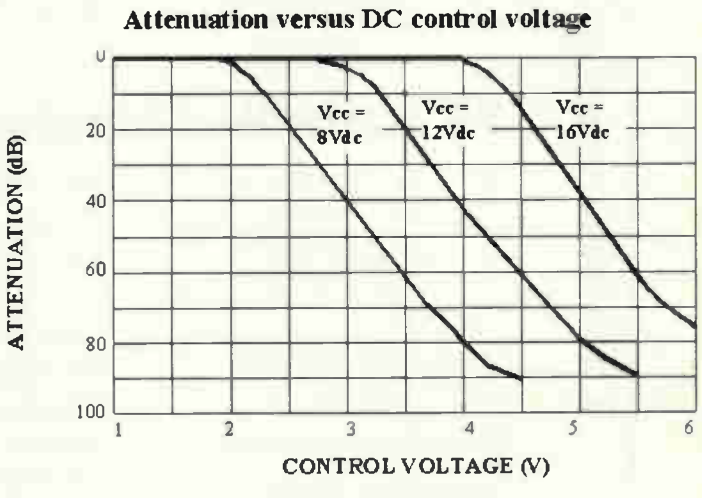


Fig. 3. Attenuation characteristic of Motorola MC3340

Circuit simulates glitches

Circuit **Fig. 1a)** can be used to simulate glitches and waveforms encountered in electronic circuits during operation. Glitches can arise due to various causes such as bouncing of mechanical switch contacts, switching transients, timing mismatch, electromagnetic interference, etc. Digital logic can interpret glitches as valid logic level transitions and erroneous operation can occur.

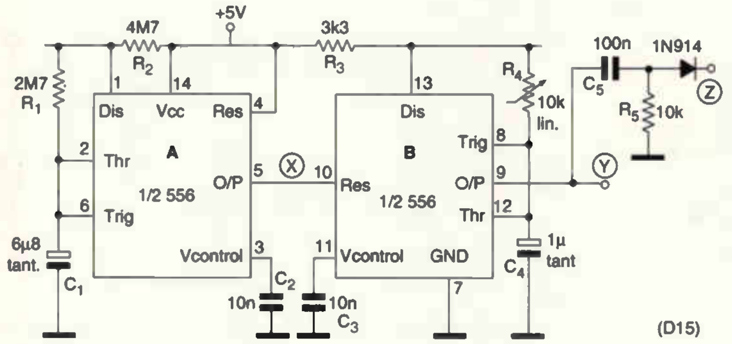
To avoid such spurious operation, glitch detectors and glitch elimination techniques are used. Nonetheless, glitches find a way into digital circuits and false operations do occur in some situations.

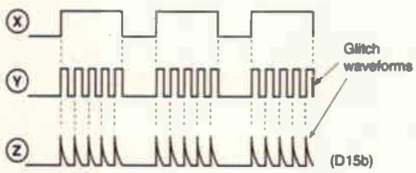
To test and validate a glitch detector or a glitch eliminator, or even a circuit designed to be immune to glitches, a test circuit that simulates glitches is needed. The circuit described here can be used to produce such a glitch-simulating test waveform.

The circuit consists of a 556 dual timer, both halves of which are configured as astable multivibrators. The frequency of the A half is kept very low, at around 1Hz. On the other hand, the frequency of the second half, B, is kept high so that B produces narrow short-duration pulses.

Output of A is connected to the reset input of B so that when the output of A is high, the astable B oscillates and when the output of A is low, the reset input of B is kept low and it does not oscillate.

This generator simulates glitches, contact bounce, etc., as an aid to making sure that a circuit is glitch proof.





The duty-cycle of the output of B can be adjusted by varying the resistor R_4. Waveforms involved are shown in Fig. **1b)**. Output at point Y consists of a sequence of short-duration pulses coming in bursts and this can be used to test logic circuits. Output at Y is differentiated by a *C-R* differentiator and rectified by diode *D* to get positive going spikes at point Z.

The output of the differentiator can also be used to test the immunity of a digital logic circuit to voltage spike type of glitches. Note that the reset function is asynchronous, so that R_4 can be set to make the last pulse in a train very short or glitch like.

Frequencies of the astable multivibrators are given by,

$$f_A = \frac{1.44}{(R_2 + 2R_1)C_1}$$

and,

$$f_B = \frac{1.44}{(R_3 + 2R_4)C_4}$$

Duty-cycle of the astable B is given by,

$$w = \frac{R_4}{(R_3 + 2R_4)}$$

To simulate different types of glitches, the duty-cycle can be altered. For example, to simulate contact bounce, you can make *w* around 10ms.

Making R_4 variable will enable change of pulse-width as required. The time-constant of the differentiator should be chosen such that $RC<T/10$ where *R* and *C* are the values of the differentiator's resistor and capacitor and *T* is the time period of the input waveform of the differentiator i.e., at Y.

Since the circuit is quite simple, it can be packaged in the form of a handy testing tool. Usage of a low-power timer such as XR-L556, or equivalent, will enable operation on batteries.

V. Lakshminarayanan
Bangalore
India
D15

£50 Winner

PC controlled bipolar stepper motor

A PC can control the speed and direction of a bipolar, i.e. two-phase, stepper motor, in a simple way as described below.

Circuit **Fig. 1** interfaces a bipolar stepper-motor to the parallel LPT port of an IBM-compatible PC. It consists of complimentary transistors connected in bridges. One bridge is required for each phase winding of the stepper motor.

Each bridge transistor should be installed on 5W rated heat sink. Diodes in each bridge are used to provide free-wheeling action. Two 2N2222-type transistors interface each bridge to the parallel port of the PC.

Data bits D_0 and D_1 on pins 2 and 3 of the parallel port, are used to drive a bridge circuit. Bits D_2 and D_3, pins 3 and 4, drive the second bridge. Pin 25 of the parallel port connects to the ground of the bridge power supply.

Manufacturers data on a 3.6V, 4A/phase, 1.8°, two-phase stepper motor is provided in **Fig. 2**. A power-supply design, capable of driving two similar motors is provided in **Fig. 3**.

Figure 4 gives the data sequence required at the parallel port to drive the stepper motor in one direction. When a low on data bit D_0 and a high on data bit D_1 is sent, this switches transistors $Tr_{1,3}$ on.

The result is a current flow through the motor's R-Y phase in one direction. When a high on data bit D_0 and a low on data bit D_1 is sent this switches transistor Tr_2 and transistor Tr_4 on. The result is a current flow through the motor's R-Y phase in the opposite direction.

Continued on page 745...

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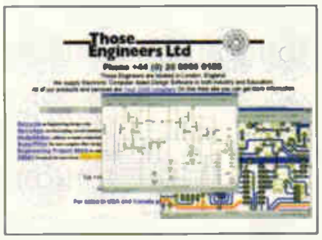

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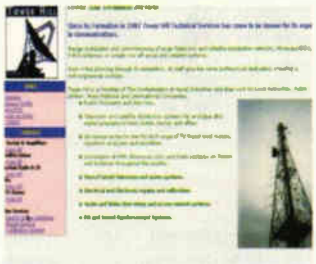

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PC controlled bipolar stepper motor

...continued from page 740

A simple QuickBasic program to run the stepper motor at any speed and in any direction is given. The parallel port's address is 378_{16}. Duration of execution of four 'FOR TO NEXT' loops in subroutines CW and CCW determines the speed of the motor. If the duration of these loops is increased, by increasing the variable D, the motor speed will reduce.

Pressing function key F1 should move the motor in a clockwise direction, while the motor should move in counter clockwise direction by pressing F2.

This circuit can be used to drive two-phase stepper motors of various sizes. ■

M T Iqbal
Rawalpindi
Pakistan.
E23

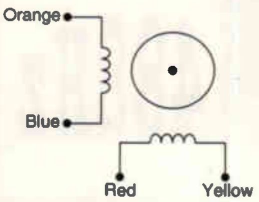
Listing. Quick Basic program for driving a stepper motor from a PC.

```
CLS:D=1000
PRINT  "Press F1 for clockwise rotation
PRINT  "Press F2 for counter clockwise"
PRINT  "Press F3 to change speed"
ON KEY(1) GOSUB CW
ON KEY(2) GOSUB CCW
ON KEY(3) GOSUB Direction
KEY(1) ON
KEY(2) ON
KEY(3) ON
10 GOTO 10
END

CW:
OUT &H378,9
FOR I=1 TO D : NEXT I
OUT &H378,10
FOR I=1 TO D : NEXT I
OUT &H378, 6
FOR I=1 TO D : NEXT I
OUT &H378, 5
FOR I=1 TO D: NEXT I
RETURN

CCW
OUT &H378, 5
FOR I=1 TO D : NEXT I
OUT &H378, 6
FOR I=1 TO D : NEXT I
OUT &H378, 10
FOR I=1 TO D : NEXT I
OUT &H378, 9
FOR I=1 TO D : NEXT I
RETURN

Direction:
INPUT "Please enter new speed"; D
RETURN
```



Step	Red	Blue	Yellow	Orange
1	–	–	+	+
2	+	–	–	+
3	+	+	–	–
4	–	+	+	–

Stepper motor Sanyo Denshi model 103H8223–5141, 3.6V, 4A/phase, 1.8°/step

Fig. 2. Connections for the stepper motor used in the author's prototype.

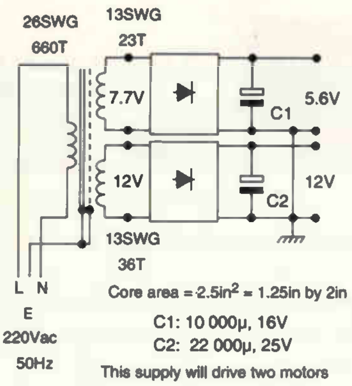
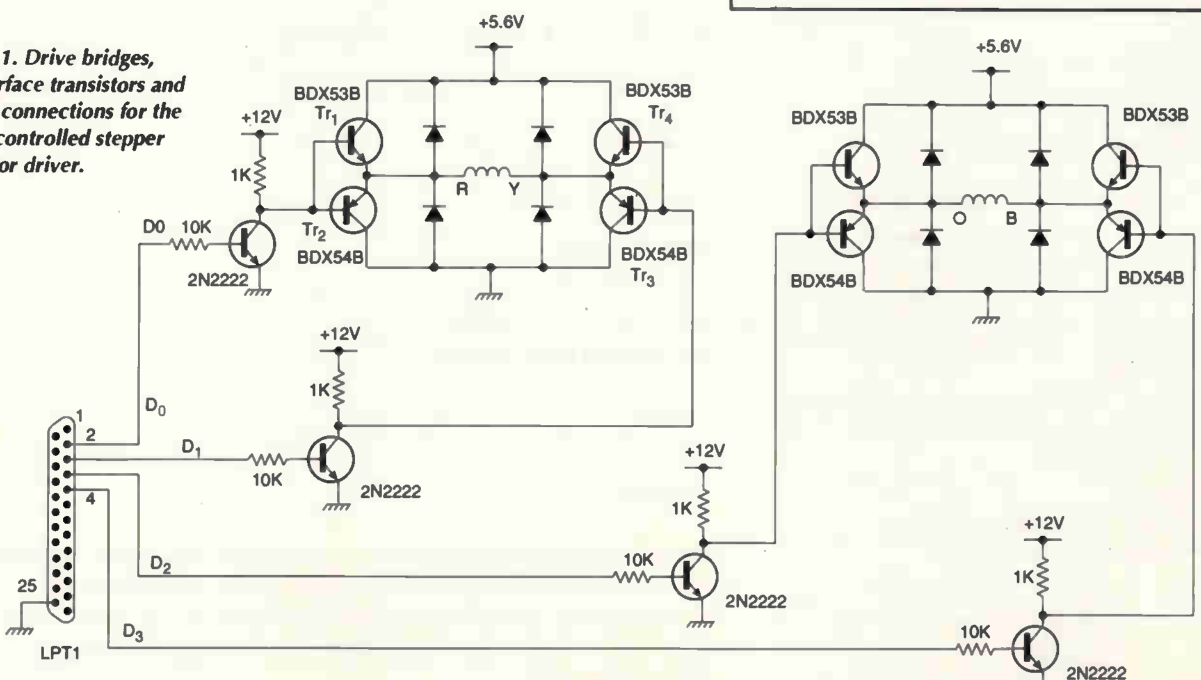


Fig. 3. Power supply for the drive circuits.

(E23d)

Steps	D0	D1	D2	D3	
1	H	L	L	H	CW
2	L	H	L	H	
3	L	H	H	L	
4	H	L	H	L	

Fig. 4. Data sequence on parallel port for clockwise rotation.



All diodes type 1N5401

Fig. 1. Drive bridges, interface transistors and LPT connections for the PC-controlled stepper motor driver.

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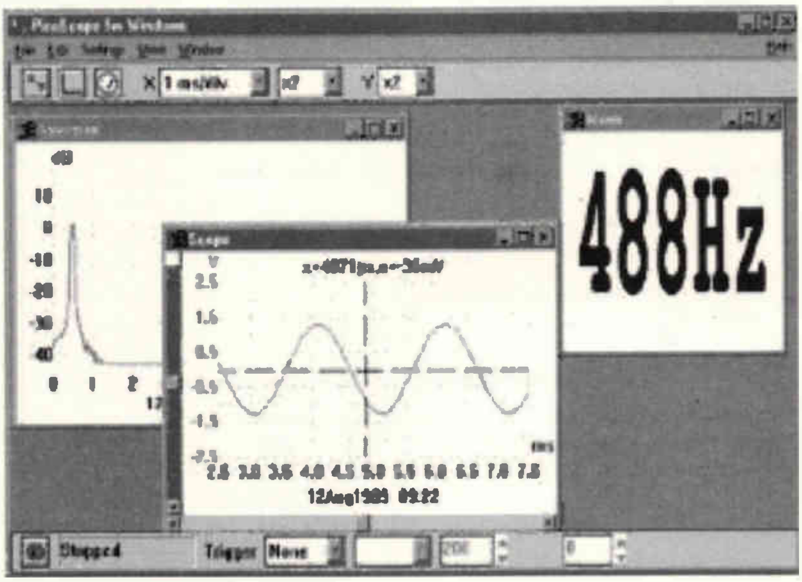

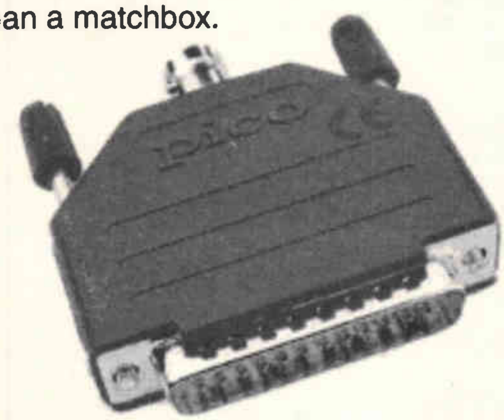

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THD is meaningless

Anthony New's article in the June issue has much merit. It would be easy to pick up and make much of a few errors, but that would do the matter a great injustice.

What, however, is worthy of comment, I think, is that the criticisms of current practices are actually well-known to many audio designers, but nevertheless to far too few.

Mr New stands to receive similar criticism to that levelled at Matti Otala when he published his first paper in English on transient intermodulation distortion. Top designers said, more or less, 'This is all old stuff, we know all about it and any claims of novelty are false.' But lots of designers didn't know about it, and designed amplifiers that exhibited it in various degrees.

A limited defence of harmonic distortion measurements is that the object of design is to achieve linearity. A design exhibiting no THD must be completely linear and cannot produce intermodulation distortion either.

Simple theory can show that there is a strict numerical relationship between the amounts of THD and IMD produced by a given non-linearity. But simple theory is too simple: there are many conditions necessary if the numerical relationship is to be preserved.

These conditions are rarely fulfiled. This is discussed in BS EN 60268-2, which is identical to IEC 60268-2. But if designers don't read these standards...

Reliance on THD control, regardless of its psychoacoustic significance, if any, worked quite well in historical times, when linearity was not very good, but it is indeed too 'broad-brush' now that linearity is excellent.

Having said that, an authoritative source of information on measurements of audio amplifiers is BS EN 60268-3, which is identical to IEC 60268-3. This has, throughout its editions – a third is imminent – included standard methods of measurement for intermodulation distortion.

The fact that designers don't use these methods, or, at least, IMD does not feature in published specifications, is regrettable. I can demonstrate the relative inaudibility of harmonics compared with even low-order intermodulation products, using contrived circuits that can produce one without the other.

John Woodgate
Via e-mail

I had intended to present a cross-section of responses to Anthony's article on IMD, but I ran out of space again, sorry. I'll try to find space again next month. Ed

Sizzling hard drives

The interesting correspondence regarding hard-drive failures due to overheating, and Ed Dell's "fridge" remedy, suggest that the problem is bearing-related.

A number of years ago, I encountered a regular failure, on very hot days, in the drive of a Transtec HD2 machine. It was similarly cured by lowering the ambient temperature around the unit and giving it a 'nudge'.

Our office had a large plate-glass window facing on to the street. Close to high-noon, the programmers' desk used to get alarmingly hot. Invariably, the HD2 would fail at around 2pm! A quick burst with the office fan and a tap on the casing of the computer got it healthy again.

Many drives rely on the precise alignment of a shaft in a set of precision agate, ruby, or similar bearings. These have built in compensation for temperature changes.

If this sensitive equilibrium is upset, the shaft can expand axially and produce heat in the bearing surfaces. This then compounds the problem because some of the heat is transferred to the shaft, which then expands even more, and so on.

If the overheating continues unchecked, radial carrier-shaft expansion can become significant and lead to bearing failure. The drive is usually unusable after this point has been reached.

In what we laughingly call 'The-Good-Old-Days', mainframe rooms were air-conditioned and temperature-controlled. However, with the proliferation of PCs, we seem to have overlooked the fact that computers, like people, have a optimum temperature at which they are most comfortable.

Chris Eccles
Research Director
Gardner Watts Ltd

Thoughts from a long-standing reader

I have been a reader of *Wireless World* since 1943 and a subscriber since 1949. During my working life, mainly in engineering research and development, I found the magazine an invaluable source of technical information.

Now I am well retired, I still read it with interest, though not having the hardware available, I do not find the computer articles very much help. Obviously they help other people in the way that earlier articles helped me.

I have always had a strong practical interest in audio and sound recording. Since I do quite a lot of headphone listening, for various reasons, I was very interested in John Watkinson's article in the November 1999 issue, particularly in the headphone shuffler circuit. I had hoped that he would produce a practical circuit for this, but since he did not, I decided to try it myself.

As I am in my mid-seventies and my maths is well on the way to rusting up solid, I adapted the group delay circuit from Bill Hardman's 'Precise active crossover' in the August issue. I based the system on the equilateral triangle spacing of speakers and listener. This required a delay of around 200μs. This was easy to obtain up to about 2kHz, but the delay fell off to about 80μs at 10kHz.

I made a shade attenuator within ±1dB of the graph given by John Watkinson. This contributed to the delay that now ranged between 260 and 95μs.

By recording short sections with and without the shuffler in circuit on a cassette recorder, I was able to assess its effect. First impressions were that the bass was increased, as you would expect from the response of the shader attenuator.

The 'in-head' effect was reduced, but the sound only seemed to be shifted about three inches forward, not comparable with actual loudspeaker listening. However, it was an interesting experiment. I would welcome seeing a proper circuit from John Watkinson. I fear my circuit is not up to the standard I expect from *Electronics World*.

I was also very interested in Anthony New's article in the June issue; I think he was a bit too sweeping in dismissing THD measurements. I have always regarded THD and IMD as two interdependent facets of non-linearity in amplifiers, IMD being the worst effect of the two.

In valve technology, as I used it in the 1950s, there was a commonly held view that IMD values had a simple relationship to THD, so that knowledge of THD gave an index of IMD. No doubt the incidence of higher-order distortions than we had in those days has affected this relationship and possibly led to the disregard for IMD that Anthony New is trying to redress.

I also enjoyed reading John Linsley Hood's reminiscences. They reminded me in places of my own experiences in electronics; though I was more interested in sound recording, building my own disc recorder on a shoe-string in the 1950s.

I look forward to reading *EW* each month.

A T Granger
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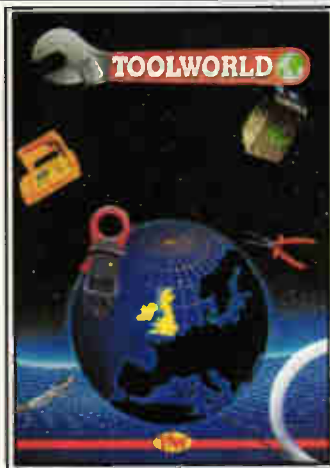

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