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ELECTRONICS WORLD

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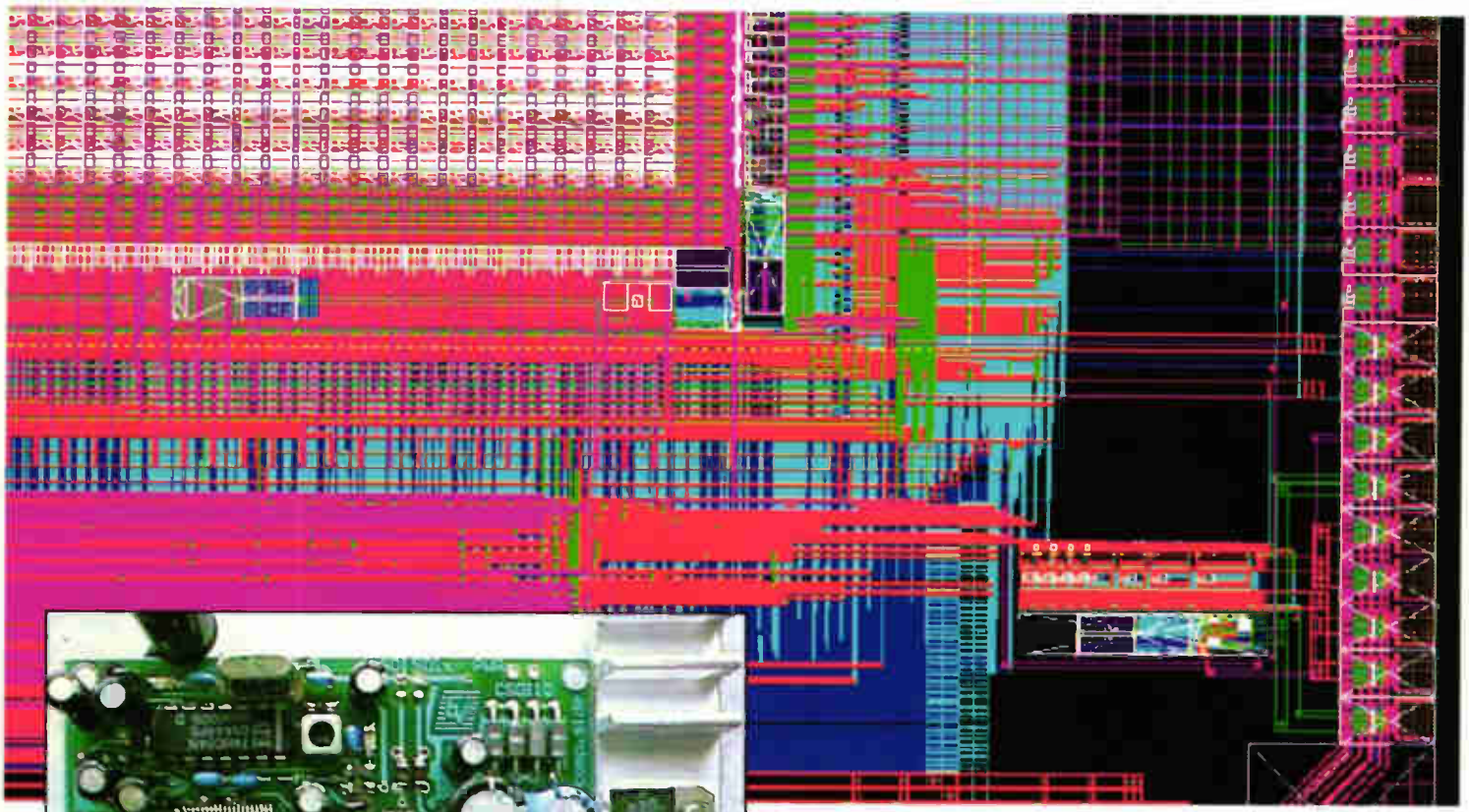
JANUARY 2004 £3.25

Douglas Self – analogue switching



Computing goes optical

Analogue to SDI converter project

Circuit ideas:

- Novel charger
- High-voltage fuse
- Digital potentiometer drive

Telnet

Quality second-user test & measurement equipment

Audio Precision – system 1 – Dual Channel Audio Test System	**£2200**
ENI 550L Amplifier (1.5 to 400MHz) 50 Watts	**£2500**
Hewlett Packard 3314A Function Generator 20MHz	**£750**
Hewlett Packard 3324A synth. function/sweep gen. (21MHz)	**£1950**
Hewlett Packard 3325B Synthesised Function Generator	**£2500**
Hewlett Packard 3326A Two-Channel Synthesiser	**£2500**
H.P. 4191A R/F Imp. Analyser (1GHz)	**£3995**
H.P. 4192A L.F. Imp. Analyser (13MHz)	**£4000**
Hewlett Packard 4193A Vector Impedance Meter (4-110MHz)	**£2900**
Hewlett Packard 4278A 1kHz/1MHz Capacitance Meter	**£3500**
H.P. 53310A Mod. Domain Analyser (opt 1/31)	**£3950**
Hewlett Packard 8349B (2 - 20 GHz) Microwave Amplifier	**£2000**
Hewlett Packard 8508A (with 85081B plug-in) Vector Voltmeter	**£2500**
Hewlett Packard 8904A Multifunction Synthesiser (opt 2+4)	**£1750**
Hewlett Packard 89440A Vector Signal Analyser (1.8GHz) opts AY8, AYA, AYB, AY7, IC2	**£9950**
H.P. ESG-D3000A 3GHz Signal Gen	**£6995**
Marconi 6310 – Prog'ble Sweep gen. (2 to 20GHz) – new	**£2500**
Marconi 6311 Prog'ble sig. gen. (10MHz to 20GHz)	**£2995**
Marconi 6313 Prog'ble sig. gen. (10MHz to 26.5GHz)	**£3750**
R&S SMG (0.1-1GHz) Sig. Generator (opts B1+2)	**£2500**
Rhode & Schwarz UPA3 Audio Analyser	**£1500**
Rhode & Schwarz UPA3 Audio Analyser	**£2250**
Fluke 5700A Multifunction Calibrator	**£12500**
Fluke 5800A Oscilloscope Calibrator	**£9995**

OSCILLOSCOPES

Gould 4068 150MHz 4 channel DSO	£1250
Gould 4074 100MHz - 400 Ms/s - 4 channel	£1100
Hewlett Packard 54502A - 400MHz – 400 MS/s 2 channel	£1600
Hewlett Packard 54520A 500MHz 2ch	£2750
Hewlett Packard 54600A - 100MHz – 2 channel	£675
Hewlett Packard 54810A 'Infinium' 500MHz 2ch	£2995
Intron 2020 - 20MHz. Dual channel D.S.O (new)	£450
Iwatstu SS 5710/SS 5702 -	from £125
Kikusui COS 5100 - 100MHz - Dual channel	£350
Lecroy 9314L 300MHz - 4 channels	£2750
Meguro MSO 1270A - 20MHz - D.S.O. (new)	£450
Philips 3295A - 400MHz - Dual channel	£1400
Philips PM3392 - 200MHz - 200Ms/s - 4 channel	£1750
Philips PM3094 - 200MHz - 4 channel	£1500
Tektronix 2213/2215 - 60MHz - Dual channel	£300
Tektronix 2220 - 60MHz - Dual channel D.S.O	£850
Tektronix 2221 - 60MHz - Dual channel D.S.O	£850
Tektronix 2235 - 100MHz - Dual channel	£500
Tektronix 2245A - 100MHz - 4 channel	£700
Tektronix 2430/2430A - Digital storage - 150MHz	from £1250
Tektronix 2445 - 150MHZ - 4 channel +DMM	£850
Tektronix 2445/2445B - 150MHz - 4 channel	£800
Tektronix 2465/2465A /2465B - 300MHz/350MHz 4 channel	from £1250
Tektronix 7104 - 1GHz Real Time – with 7A29 x2, 7B10 and 7B15	from £1950
Tektronix TDS 310 50MHz DSO - 2 channel	£750
Tektronix TDS 420 150 MHz 4 channel	£950
Tektronix TDS 520 - 500MHz Digital Oscilloscope	£2500

SPECTRUM ANALYSERS

Advantest 4131 (10kHz – 3.5GHz)	£3000
Advantest/TAKEDA RIKEN - 4132 - 100KHz - 1000MHz	£1350
Ando AC 8211 - 1.7GHz	£1500
Avcom PSA-65A - 2 to 1000MHz	£750
Farnell SSA-1000A 9KHz-1GHz Spec. An.	£1250
Hewlett Packard 182T Mainframe + 8559A Spec.An. (0.01 to 21GHz)	£2000
Hewlett Packard 853A Mainframe + 8559A Spec.An. (0.01 to 21GHz)	£2500
Hewlett Packard 3582A (0.02Hz - 25.5kHz) dual channel	£1500
Hewlett Packard 3585A 40 MHz Spec Analyser	£3000
Hewlett Packard 3585B 20 Hz - 40 MHz	£4500
Hewlett Packard 3561A Dynamic Signal Analyser	£3500
Hewlett Packard 8568A -100kHz - 1.5GHz Spectrum Analyser	£3500
Hewlett Packard 8590A (opt 01, 021, 040) 1MHz–1.5MHz	£2500
Hewlett Packard 8592B 9 KHz - 22 GHz	£7995
Hewlett Packard 8594 9 KHz - 2.9 GHz	£4500
Hewlett Packard 8596E (opt 41, 101, 105,130) 9KHz - 12.8GHz	£9950
Hewlett Packard 8713C (opt 1 E1) Network An. 3 GHz	£6000
Hewlett Packard 8713B 300kHz - 3GHz Network Analyser	£5000
Hewlett Packard 8752A - Network Analyser (1.3GHz)	£4995
Hewlett Packard 8753A (3000KHz - 3GHz) Network An.	£3250
Hewlett Packard 8753B+85046A Network An + S Param (3GHz)	£6500
Hewlett Packard 8754A - Network Analyser 4MHz -1300MHz)	£1500
Hewlett Packard 8756A/8757A Scaler Network Analyser	from £900
Hewlett Packard 8757C Scalar Network Analyser	£3500
Hewlett Packard 70001A/70900A/70906A/70902A/70205A - 26.5 GHz Spectrum Analyser	£7000
IFR A7550 - 10KHz-GHz - Portable	£1750
Meguro - MSA 4901 - 30MHz - Spec Anaylser	£600
Tektronix 492P (opt1,2,3) 50KHz - 21GHz	£3500
Wiltron 6409 - 10-2000MHz R/F Analyser	£1250
Tek 496 (9KHz-1.8GHz)	£2500

Radio Communications Test Sets

Hewlett Packard 8920B (opts 1,4,7,11,12)	£6750
Hewlett Packard 8922M + 83220E	£2000
Marconi 2955	£1250
Marconi 2955B/60B	£3500
Marconi 2955R	£1995
Motorola R2600B	£2500
Racal 6103 (opts1, 2)	£5000
Rohde & Schwarz SMFP2	£1500
Rohde & Schwarz CMD 57 (opts B1, 34, 6, 19, 42, 43, 61)	£4995
Rohde & Schwarz CMT 90 (2GHz) DECT	£3995
Rohde & Schwarz CMTA 94 (GSM)	£4500
Schlumberger Stabilock 4015	£3250
Schlumberger Stabilock 4031	£2750
Schlumberger Stabilock 4040	£1300
Wavetek 4103 (GSM 900) Mobile phone tester	£1500

MISCELLANEOUS

Ballantine 1620A 100Amp Transconductance Amplifier	£1250
EIP 545 Microwave Frequency Counter (18GHz)	£1000
EIP 548A and B 26.5GHz Frequency Counter	from £1500
EIP 575 Source Locking Freq.Counter (18GHz)	£1200
EIP 585 Pulse Freq.Counter (18GHz)	£1200
Fluke 6060A and B Signal Gen. 10kHz - 1050MHz	£950
Genrad 1657/1658/1693 LCR meters	from £500
Gigatronics 8541C Power Meter + 80350A Peak Power Sensor	£1250
Gigatronics 8542C Dual Power Meter + 2 sensors 80401A	£1995
Hewlett Packard 339A Distortion measuring set	£600
Hewlett Packard 436A power meter and sensor (various)	from £750
Hewlett Packard 438A power meter - dual channel	£1750
Hewlett Packard 3335A – synthesiser (200Hz-81MHz)	£1750
Hewlett Packard 3457A mull meter 6 1/2 digit	£850
Hewlett Packard 3784A - Digital Transmission Analyser	£2950
Hewlett Packard 37900D - Signalling test set	£2500
Hewlett Packard 34401A Multimeter	£500
Hewlett Packard 4274A LCR Meter	£1750
Hewlett Packard 4275A LCR Meter	£2750
Hewlett Packard 4276A LCZ Meter (100MHz-20KHz)	£1400
Hewlett Packard 5342A Microwave Freq.Counter (18GHz)	£850
Hewlett Packard 5385A - 1 GHz Frequency counter	£495
Hewlett Packard 6060A and B Electronic Load 300W	from £750
Hewlett Packard 6622A - Dual O/P system p.s.u	£950
Hewlett Packard 8350B - Sweep Generator Mainframe	£1500
Hewlett Packard 8642A – high performance R/F synthesiser (0.1-1050MHz)	£2500
Hewlett Packard 8656A - Synthesised signal generator	£750
Hewlett Packard 8656B - Synthesised signal generator	£995
Hewlett Packard 8657A - Synth. signal gen. (0.1-1040MHz)	£1500
Hewlett Packard 8657B - 100MHz Sig Gen - 2060 MHz	£3950
Hewlett Packard 8657D - XX DQPSK Sig Gen	£3950
Hewlett Packard 8901B - Modulation Analyser	£1750
Hewlett Packard 8903A, B and E - Distortion Analyser	from £1000
Hewlett Packard 11729B/C Carrier Noise Test Set	from £2500
Hewlett Packard 53131A Universal Frequency counter (3GHz)	£850
Hewlett Packard 85024A High Frequency Probe	£1000
Hewlett Packard 6032A Power Supply (0-60V)-(0-50A)	£2000
Hewlett Packard 5351B Microwave Freq. Counter (26.5GHz)	£2750
Hewlett Packard 5352B Microwave Freq. Counter (40GHz)	£5250
Keithley 220 Programmable Current Source	£1750
Keithley 228A Prog'ble Voltage/Current Source IEEE.	£1950
Keithley 237 High Voltage - Source Measure Unit	£3950
Keithley 238 High Current - Source Measure Unit	£3750
Keithley 486/487 Picoammeter (+volt.source)	£1350/£1850
Keithley 617 Electrometer/source	£1950
Keithley 8006 Component Test Fixture	£1750
Marconi 6950/6960/6960A/6970A Power Meters & Sensors	from £400
Philips 5515 - TN - Colour TV pattern generator	£1400
Philips PM 5193 - 50 MHz Function generator	£1350
Phillips PM 6654C System Timer Counter	£750
Rohde & Schwarz FAM (opts 2,6 and 8) Modulation Analyser	£2500
Rohde & Schwarz NRV/NRVD Power meters with sensors	from £1000
Tektronix AM503 - AM503A - AM503B Current Amp's with M/F and probe	from £800
Wayne Kerr 3245 - Precision Inductance Analyser	£1750
Bias unit 3220 and 3225L Cal.Coil available if required.	(P.O.A)
Wayne Kerr 3260A + 3265A Precision Magnetics Analyser with Bias Unit	£5500
W&G PCM-4 PCM Channel measuring set	£3750

All equipment is used – with 30 days guarantee and 90 days in some cases.
Add carriage and VAT to all goods.

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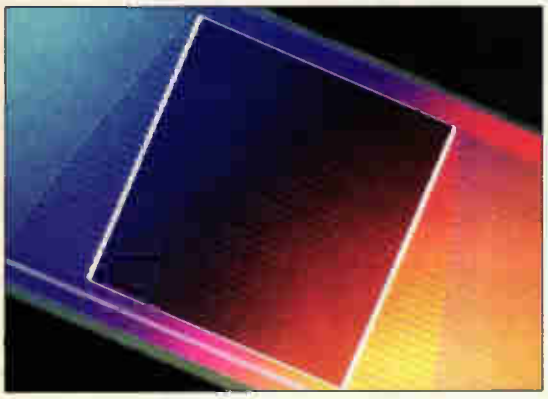

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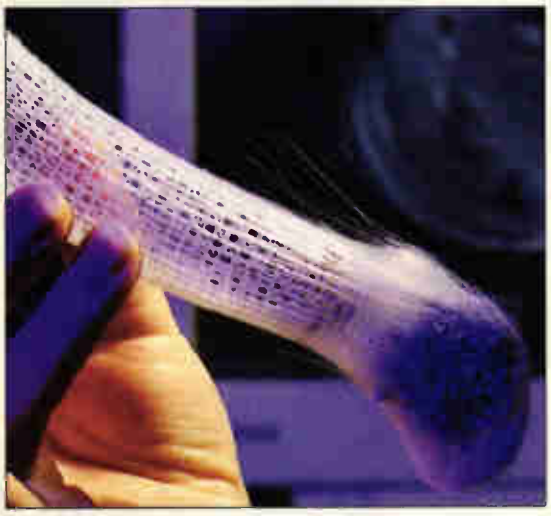

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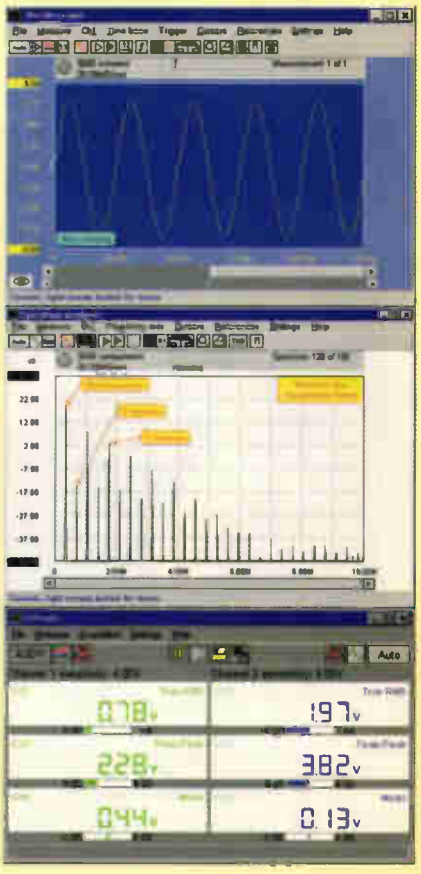
THERE IS INTERESTING NEWS

OSCILLOSCOPE | FFT ANALYSER | VOLTMETER | RECORDER

The Handyscope 3 is a powerful and versatile two channel measuring instrument with an integrated function generator.

- USB 2.0 connection (USB 1.1 compatible)
- sample speed up to 100 MHz per channel
- 8 to 16 bit resolution (6 µVolt resolution)
- 50 MHz bandwidth
- input sensitivity from 200 mVolt up to 80 Volt
- large memory up to 131060 samples per channel
- four integrated measuring devices
- spectrum analyser with a dynamic range of 95 dB
- fast transient recorder up to 10 kHz
- several trigger features
- auto start/stop triggering
- auto disk function up to 1000 files
- auto setup for amplitude axis and time base
- auto trigger level and hysteresis setting
- cursor measurements with 21 read-outs
- very extensive function generator (AWG) 0-2 MHz , 0-12 Volt



HANDYSCOPE HS3

WWW.TIEPIE.NL

for more information, demo software, software, source code and DLL's visit our internet page: http://www.tiepie.nl

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Oooooooh!

I'd like to thank all the readers who supported me in the 'O' or zero (old transistors conundrum) as it would appear that your editor was correct for once! Some of the substantiating evidence is on the *Letters* page.

And thanks also to the eagle eyed readers who attained better geography qualifications than all of us here. A selection of responses also appears in *Letters* but thanks also to Brian Johnson and Pigeon.

The Observant Electronics Datastation competition winner, drawn out of the hat by Caroline, was Richard Faulkner of the Bath Institute of Medical Engineering who will use the product in the development of assistive technology for people with dementia.

Following on from the 'Mickey Mouse' degree debate, an interesting press release came across my desk this month from the Disney Corporation. It appears that they are about to launch a range of audio and video products aimed at children. 'Mickey Mouse' electronic products are born – it's official. No longer will you be able say things like 'that looks a bit Mickey Mouse' for fear of litigation!

A more serious and worrying press release has crossed my desk, however. It comes from Envirowise, the UK environmental support and advisory service. It would appear that the 'powers that be' in good old Brussels have decided that the electronics industry needs to do something about the amount of pollution it causes. Two new directives – the WEEE (Waste Electrical and Electronic Equipment) and the ROHS (Restriction Of use of certain Hazardous Substances) – will have serious implications for all of us. They reckon the expense of re-design and the cost of compulsory recycling arrangements will cost the industry £655 million a year. This will mean a cost increase of 1-2% for many products and as much as 3-4% for "some larger and complex products".

According to a Dr. Gibson "With many smaller electrical / electronics companies struggling to survive on profit margins that are little more than the cost of these increases, companies could find themselves sitting on a time bomb that they are too late to respond to. The longer they delay before taking action, the more costly it is going to be. Many companies, especially small ones, could go out of business unless they start to take action now."

While pollution is obviously important (or lack of it), I suspect a lot of companies will change direction away from electronics manufacture or may just shut up shop altogether. Thanks to all in Brussels who are helping to shrink the European electronics industry even further. Merry Xmas.

Phil Reed

New editorial and advertising address

The Highbury Business Communications office previously at Cheam, Surrey has moved to Swanley in Kent. All correspondence intended for the editorial and advertising departments should be addressed to:

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SiGe bipolar transistor matches CMOS-on-insulator

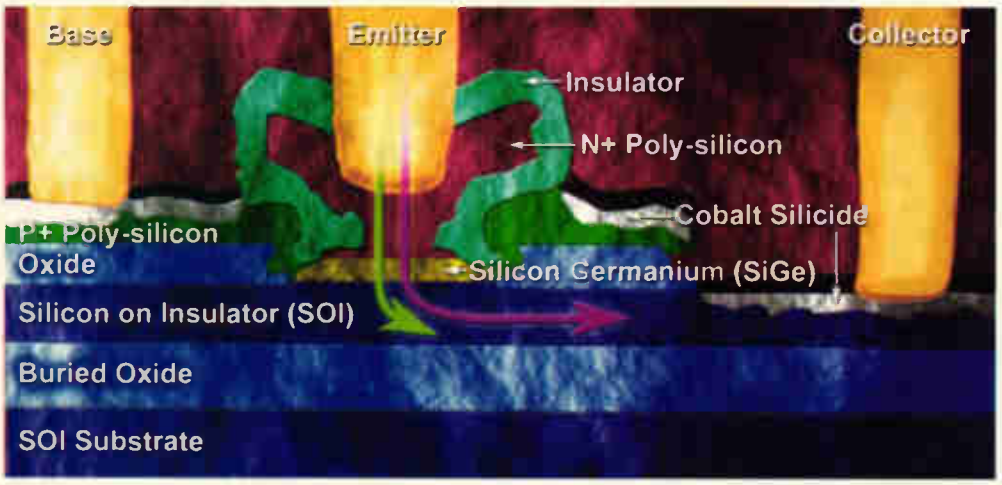
IBM researchers have designed a silicon-germanium (SiGe) bipolar transistor which can be integrated with silicon-on-insulator (SOI) CMOS.

SOI CMOS is used for the highest performance logic circuits and SiGe bipolar transistors save power and cut noise - compared to CMOS - in multi-GHz silicon RF circuits, and many analogue circuits.

Unfortunately, the two are not compatible.

BiCMOS processes are the usual way to combine CMOS and bipolar transistors, but bipolars cannot easily be made on SOI wafers.

IBM's most recent solution is shown in the diagram, with a silicon-germanium base built directly on the SOI wafer. The design solves the integration problem and also allows bipolar voltage capability to be traded for speed.



Electrons come down from the poly-silicon emitter, accelerate through the silicon germanium base, and make a turn in the SOI layer towards the collector contact electrode.

With zero or low voltage applied to the SOI wafer, the current path in the SOI is long, which results in low electric field in the SOI and makes the device suitable for high voltage applications (pink arrow).

With high positive voltage applied to the SOI wafer, the collector contact is virtually extended all the way to the back of the SOI layer under the emitter. The current path is thus shorter (green arrow), making the device suitable for high speed applications.

New watches for old

LED watches appeared and disappeared in the 70s and for fans have subsequently gained an air of retro-cool.

Now a UK firm has designed an LED watch for the new millennium.

"We are big fans and collectors of original 70's Pulsar, Compuchron and HP watches and have been trying to find a module manufacturer to bring them back for a few years," the firm told *Electronics World*. "They are as close as you could possibly get to the 70's ones with newly designed bodies and current technology modules."

Brightness is the same as older designs, but battery life is claimed to be better - although you still have to push a button to see the time.

The watches are made in Asia and are available, mail order only, from www.led-watch.com

Shake me for power

Self-powering circuits that derive their electricity from micro-electromechanical systems (MEMS) have been developed by IMEC, the Belgian research centre.

A prototype from IMEC uses vibrations to scavenge energy. Interlocking combs of fingers are etched into silicon and the change in capacitance is enough to generate some charge.

IMEC claims a power output of 180nW for a device measuring 0.04mm^3 which is vibrating at just over 1kHz. A larger area device, perhaps 20x20mm could yield 10μW, said IMEC.

The researchers are also looking to scavenge energy from the heat of human skin. Depending on activity level, waste heat density is between 4 and 14mW/cm^2.

Using the Seebeck effect, IMEC has managed to generate 4mW. The technique could be a 24hr replacement for solar cells.

Bargain supercomputer likely to hit top three slot

Virginia Tech's recently completed Terascale Cluster, made from 1,100 of Apple's recently introduced Macintosh G5 dual-processor desktop computers, is heading for the third spot in the world Top500 computer rankings.

According to www.supercomputingonline.com, the computer will rank third among the world's best machines if unconfirmed claims of 9.6 trillion operations per second (9.6Top/s) are true. NEC's 35Top/s Earth Simulator and Los Alamos' ASCI Q are likely to remain in front.

What is remarkable about the Virginia cluster, which has been nicknamed Big Mac, is that it cost little more than $5m and was put together in a few months by a production line of students.

The actual Top500 ranking will be announced by arbiter Jack Dongarra of the University of Tennessee at the 2003 Supercomputing Conference in Phoenix.

The Terascale Cluster under construction

Optical fibres found in sea creature

Scientists from Lucent Technologies' Bell Labs have found a deep-sea sponge whose skeleton is made from optical fibre similar to that used for communications.

"We believe this novel biological optical fibre may shed light upon new bio-inspired processes that may lead to better fibre optic materials and networks," said Bell Labs materials scientist Joanna Aizenberg.

The sponge, called *Euplectella*, lives in the tropics and grows to about 15cm in length.

Commonly known as the Venus Flower Basket, it has an intricate cylindrical mesh-like skeleton of glassy silica fibres, typically between 5 and 18cm long and about the thickness of a hair.

"These biological fibres bear a striking resemblance to commercial telecommunications fibres, as they use the same material and have similar dimensions," said Aizenberg.

Each fibre is composed of distinct layers with different optical properties. In them concentric silica cylinders with high organic content surround an inner core of high-purity silica glass.

The biological fibres are extremely resilient to cracks and breakage, said Bell labs, and conduct light but do not have the high transparency needed for telecommunication.

Scientists hope to learn some tricks if they can work out how the sponge chemically deposits fibres at seawater temperature.

The skeleton of the venus flower basket sponge is made from optical fibres

Tunnel diodes back in fashion

A professor at Ohio State University in the US has developed tunnel diodes to the point where they could make a commercial comeback.

Professor Paul Berger claims to have achieved a peak-valley current ratio of 4.0 at rooom temperature for the diode, made from silicon germanium.

"Our goal was to develop a tunnel diode that could be built directly onto a traditional computer chip at minimal cost," Berger said. "And we've achieved that."

Tunnel diodes have an area of negative resistance in their forward conducting region. This allows them to be used as a very fast switch.

The current ratio achieved by Berger is close to that of silicon devices, while the SiGe structure allows for devices with both low and high current densities.

Low current densities of around 10A/cm^2 would be suited to memory, while high densities of 50kA/cm^2 would suit RF devices.

Computing goes optical

A digital signal processor that works entirely in the optical domain has been developed by an Israeli firm. Lenslet claims its device is capable of performing eight trillion operations per second (8Tops), making it some three orders of magnitude faster than conventional DSPs.

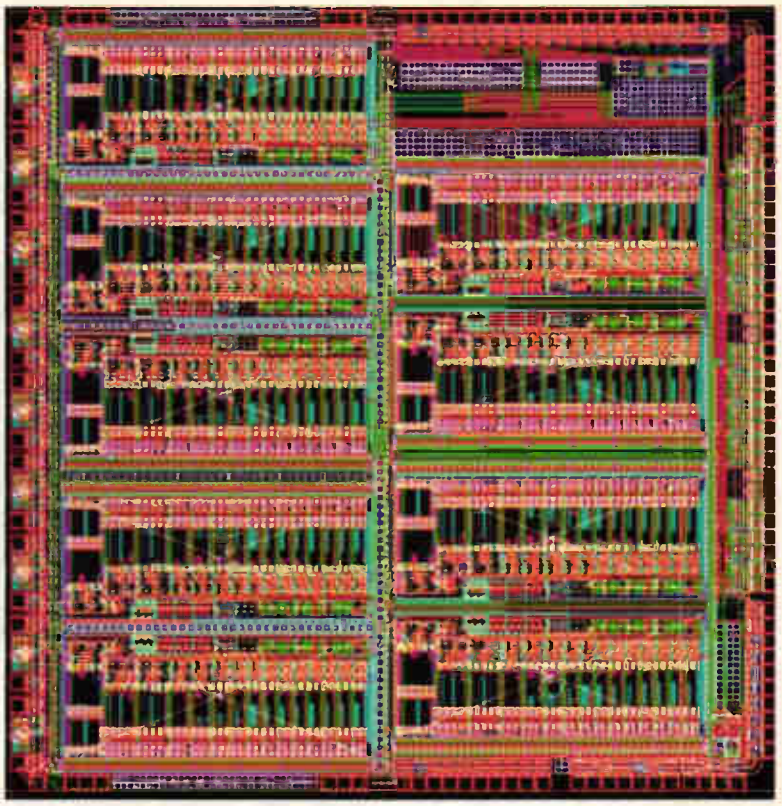
Applications are seen in noise-prone communications, synthetic aperture radar and digital beam forming.

The EnLight256 processor achieves its remarkable speed, using just 40W of power, by performing 256 vector multiply operations in parallel.

Data at 8-bit resolution is fed into the circuit via 256 vertical cavity surface emitting lasers. Light is passed through spatial light modulators to a 256x256 array of multipliers.

The array allows algorithms such as Fourier transforms, Euclid distance computations, picture motion detection and tensor manipulations to be calculated.

Photodiodes and analogue to digital converters are used to turn the output data back into electrical signals.



Dyson gets in a spin

A vacuum cleaner motor that spins at 100,000rpm has been developed by engineers at Dyson.

Having reached the power and efficiency limits of standard AC series motors, the firm decided to turn to a switched reluctance design, which can exceed 80 per cent efficiency.

The resulting motor, the X020, with its combined compressor fan, is capable of shifting 30 litres of air per second at a pressure of 30kPa, or 4.3psi.

Despite spinning three times faster than a conventional AC motor, the X020 weighs 25 per cent less at 1kg, said the firm.

Dyson used ideas from automotive turbochargers in the design, along with CAD techniques such as finite element analysis and computational fluid dynamics. These were crucial to ensure the high speed impeller does not self destruct.

The rotor and impeller assembly is made of polymer and steel plates, and at full speed the edge of the rotor spins at over 600mph within 0.3mm of the coils.

Switched reluctance motors are contactless and brushless, but use no permanent magnets. Magnetic flux created by coils in the stator affect steel plates on the rotor, and so can be used as the basis for a stepper motor. Coils do not need to be fed by sinusoidal currents, and this simplifies control, especially at high speeds.

Switching the control signals between coils requires feedback of the rotor's position, and Dyson uses an optical sensor. Sensor-less systems have been built, at the expense of more complex electronics.

Solar panels get Sharp

Rising European demand for solar panels, especially in the UK, has led Sharp to open a manufacturing plant in Wrexham, North Wales.

Photovoltaic modules manufactured at the plant will include a 72-cell 175W panel and a 48-cell 160W panel, said the firm.

"This decision will guarantee that the Wrexham facility will remain at the heart of Sharp's operations in Europe," said Hiroshi Sasaoka, chairman of Sharp Electronics (UK).

"The number of jobs created by this announcement will be 45 in 2004, rising to 90 in 2005," he added.

With production starting in the spring, the firm hopes to output panels capable of producing 20MW of power next year.

Panels produced in Wrexham will be suited to residential, commercial and industrial installations, said the firm, and will include grid-connected systems.

Last year, Europe installed over 100MW of solar cell power systems, while this year the expectation is that 160MW will be installed.

The UK share of this was just 6MW last year, but this figure should rise to 20MW within three years. Driving this is the Government's push towards renewable energy.

A target of generating ten per cent of the country's energy from renewable resources by 2010 has been set.

Typical system costs seems to be around £12,000 for a 2kW system, which would cover around 16m^2.

Residents and companies wishing to install photovoltaic cells can apply for a 50 per cent grant from the DTI. So far the Government has paid out over £2.5m to install 670kW of capacity.

In Germany, customers with grid-connected panels are paid up to ¤0.457/kWh for their electricity, a move yet to be copied in the UK.

www.est.org.uk/solar

Superconductor scans the skies

Extremely sensitive sensors that use superconductors to measure the wavelength of a single photon are being developed at NASA's Jet Propulsion Laboratory and the California Institute of Technology.

The devices, which could be formed into large arrays, will be used to accurately measure the background radiation of the universe or work as sensor in medical systems, said JPL.

A superconducting sensor can measure the wavelength of photons from optical source through to X-rays. It manages this because varying wavelength photons have differing effects on electrons in the superconductor.

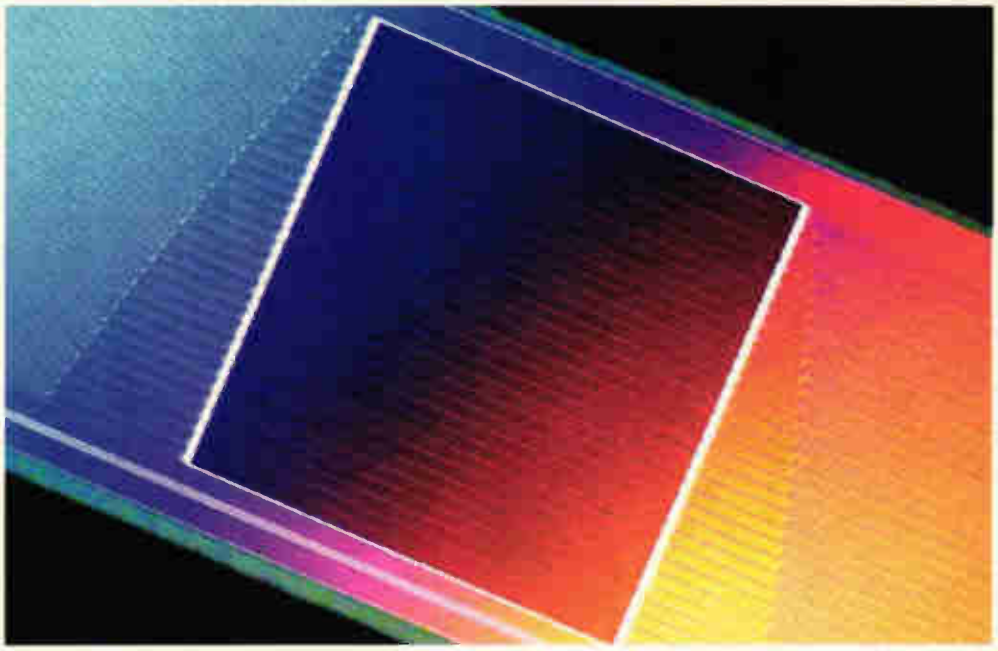
In a standard superconductor, electrons combine to form Cooper pairs. When the material, in this case aluminium at just 1.2K, is struck by a photon the energy is transferred through photoabsorption.

A cascade process occurs, with the single incident photon causing many Cooper pairs to be broken, thus reducing the superconducting properties.

A high frequency signal in the superconductor is delayed slightly, depending on the wavelength of the original photon. Thus energy and frequency of the photon can be calculated.

A 6keV X-ray photon can be resolved to within one part in 600. Eventually the JPL team hope to achieve one part in 2,000, while an optical photon should be resolved to within one part in 50.

A 32-element array is under development.



Taiwanese shrink the PC

The personal computer is about to get even smaller with Taiwanese firm VIA producing 15x15mm processors aimed at its forthcoming 'nanoITX' board.

NanoITX has an area of just 120x120mm, and is aimed at so-called 'cube' and fanless PCs. These products are typically using the mini-ITX format of motherboard, which measures 170x170mm.

The new processor is called the Eden-N, and has a thermal design power of 7W at 1GHz.

It integrates hardware security circuits, including an encryption engine and two random number generators. The latter circuits use electrical noise in the chip's substrate as the seed for the random number.

Despite the extra circuitry, VIA claims the whole processor fits in a 52mm^2 die.

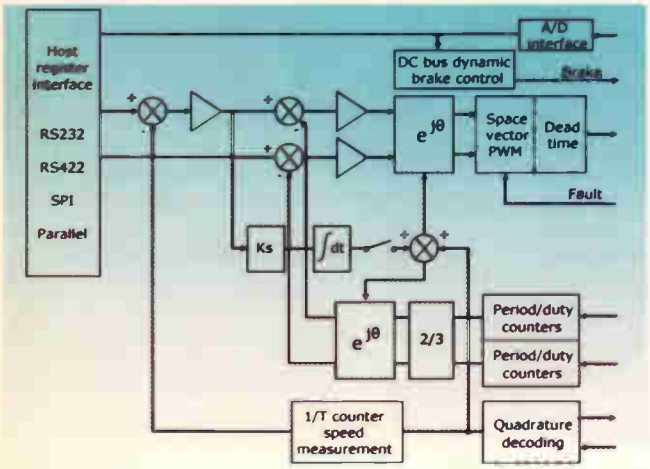
The firm has an associated chipset, the CLE266, supporting MPEG-2 decode, 5.1 channel surround sound, USB2.0 and 10/100 Ethernet.

Chips control AC motors without DSP

International Rectifier has introduced two digital motion control ICs for motor controls, completing its portfolio of motor control semiconductors.

Unlike most motor control chips, there is no DSP (digital signal processor) in the IR designs.

Instead control algorithms are hard-wired into a complex logic circuit called the 'motion control engine'. This "eliminates software programming. Motion control products developed with our digital control ICs can be brought to market faster, since commands are simplified to menu-driven selections", said IR.



To allow the chips to be used with different motors, gearboxes and loads, the chips include motor parameter registers - which are loaded at switch-on, either from a microcontroller or, for stand-alone applications, a serial EEPROM.

Supplied design software allows applications to be developed on a PC and parameters to be downloaded to a hardware development kit. The firm has begun a library of parameters for popular motors to give designers a head start.

One chip, the IRMCK201, is for velocity control of AC induction and permanent magnet servo motors with encoder feedback. It is designed to be used with IR's IR2175 linear current-sensing IC, the IR2136 three-phase inverter-driver IC and its IRAM 6A to 20A integrated power modules.

The second chip, the IRMCK203, is aimed at sinusoidal current-based sensorless control of permanent magnet AC machines and can be used for motors running at up to 100,000rpm and over a 20:1 speed range. The chip includes sensorless feedback circuits, a space vector PWM algorithm and a proprietary starting algorithm.

Both chips include a fast SPI (serial peripheral interface) port for connecting to low and medium-bandwidth external devices, and support communication over RS-232, RS-422, RS-485.

IR already makes power IGBTs and mosfets as well as high-voltage logic-to-transistor driver chips.

www.irf.com

Nano-prisms improve LCD frontlight

Japanese firm Omron has developed an LCD front lighting technology for handheld devices which it claims delivers three times more contrast than existing frontlights, and needs less power than backlights.

Like other frontlights, the device is an edge-lit plastic sheet which sits over the display. What makes it different is the embossed pattern of micro and nano-prisms on its surfaces.

The light source is actually an LED feeding from one corner. Light from the LED is directed towards the display by a pattern of 3μm deep triangular grooves in the upper surface - which act as totally-internal reflecting microprisms.

Each groove is curved, keeping a constant distance from the LED, and spaced to make illumination constant over the LCD.

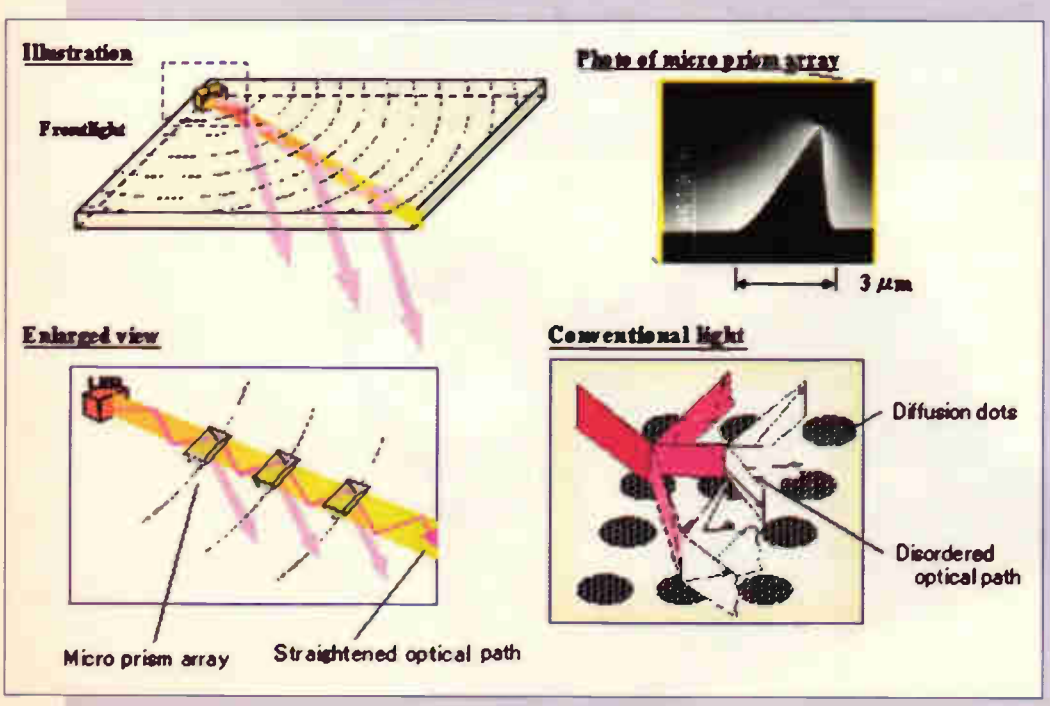
Using prisms, rather than the scattering dot pattern sometimes used, means light hits the display at right angles, not at random angles, and can be used more effectively.

However, the microprisms also reflect ambient light and can therefore cause a loss of contrast ratio, called optical noise by Omron - which is where the 200nm nanoprisms embossed on both sides come in - see diagram.

200nm is half the wavelength of the shortest (blue 400nm) visible light and, said Omron: "doing this enables control of the optical noise produced on the frontlight interface, realising a high-contrast, clear image display."

Omron sees the technology being applied to phones, cameras and PDAs. Electronics World saw the frontlight in action at Omron's 2003 Technology Fair in Kyoto, Japan. Although the exact figures cannot be confirmed, the resulting display is both evenly lit, bright and easy to read.

www.omron.com



Zetex gen5 transistors have improved metalisation

Oldham's Zetex has announced a fifth generation of its low saturation voltage bipolar transistors.

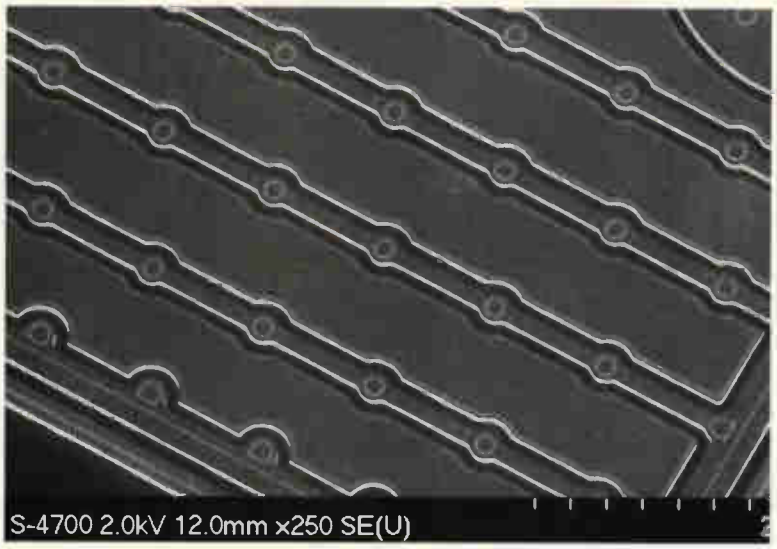
Aimed at switching power rails in handheld devices, the range includes an NPN device which saturates down to 0.113V_{CE} at 5A with 500mA base current - a forced gain (I_C/I_B) of 10.

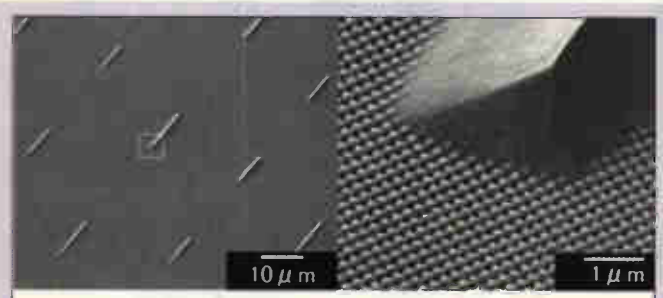
For portable applications, lower forced gains are more realistic and the devices are likely to saturate - to somewhat higher voltages - with (I_C/I_B) of 100 or more if they follow previous generations.

The first two devices will be the FCX2016 40V 4.5A PNP transistor, and the SOT89 40V 5.5A ZX5T3Z - the latter of which will saturate to 60mV at 1A (I_B=100mA), and 165mV at I_B=10mA.

Four NPN (25-100V and 7-6A) and four PNP (30-140V 5.5-4A) will be released in SOT223 packages.

For battery switching, some Gen5 transistors will be specified to leak only a fraction of a microampere with no base connection, said Zetex, allowing a PNP device to fully disconnect current in either direction when driven only by an open collector output.



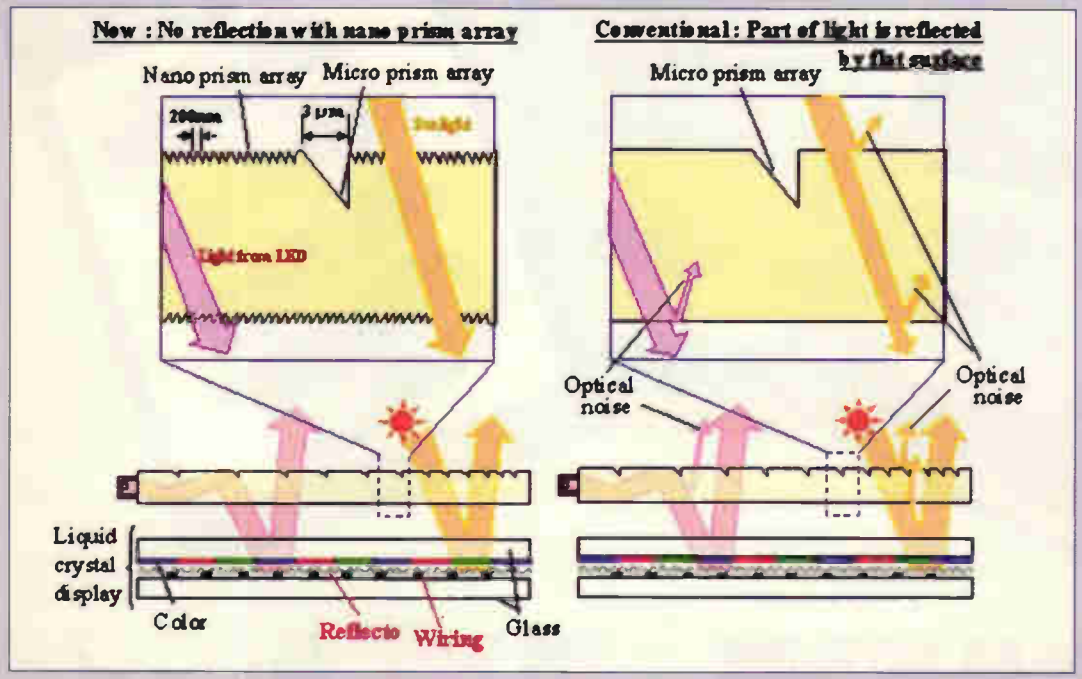


reduced enlarged

Hybrid integrated nano prism array and micro prism array

Table

Light type	Supply mA	Contrast	Brightness Cd/m²
Omron's front	20	55:1	100
Other front	30	15:1	30
Other back	60	50:1	100



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Electronic analogue switching

Part 1: CMOS gates

The switching and routing of analogue signals is a fundamental part of signal processing, but it is not easily implemented if accuracy and precision are required. This article focuses on audio applications, but the basic parameters such as isolation and linearity are equally relevant to many fields. Douglas Self elucidates

Any electronic switching technique must face comparison with relays, which are still very much with us. Relays give total galvanic isolation between control and signal, zero contact distortion, and in audio terms have virtually unlimited signal-handling capability. They introduce negligible series resistance and shunt leakage to ground is usually not even worth thinking about. Signal 'offness' can be very good, but as with other kinds of switching, this depends on intelligent usage. There will always be capacitance between open contacts, and if signal is allowed to crosstalk through this to nominally off circuitry, the 'offness' will be no better than other kinds of switching. (Throughout this article I use the word 'offness'- which is not found in any spell-checker but is widely used in the pro audio sector- as the quickest way of referring to the ratio in dB by which an unwanted input is suppressed.)

Obviously relays have their disadvantages. They are big, expensive, and not always as reliable as more than a hundred years of development should have made them. Their operating power is significant. Some kinds of power relay can introduce disastrous distortion if used for switching audio because the signal passes through the magnetic soft-iron frame; however such problems are likely to be confined to the output circuits of large power amplifiers. For small-signal switching the linearity of relays can normally be regarded as perfect.

Electronic switching is usually implemented with CMOS analogue gates, of which the well-known 4016 is the most common example, and these are examined first. However, there are many special applications where discrete JFETs provide a better solution, so these are dealt in the second article.

Analogue gates

CMOS analogue gates, also known as transmission gates, are quite different from CMOS logic gates, though the underlying process technology is the same. Analogue gates are bilateral, which means that either of the in/out leads can be the input or output; this is most emphatically not true for logic gates. The 'analogue' part of the name emphasises that they are not restricted to any fixed logic levels, but pass through whatever signal they are given with low distortion. The 'low' word there requires a bit of qualification, as will be seen later.

There is no 'input' or 'output' marked on these gates, as they are symmetrical. When switched on, the connection between the two pins is a resistance which passes current in each direction as usual, depending on the voltage between the two gate terminals. Analogue gates have been around for a long time and are in some ways the obvious method of electronic switching. They do however have significant drawbacks.

Analogue gates such as the 4016 are made up of two MOS FETs of opposite polarity connected back to back. The internal structure of a 4016 analogue gate is shown in Fig 1. The two transmission FETs with their protective diodes are shown on the right; on the left is the control circuitry. A and B are standard CMOS inverters whose only function is to sharpen up the rather soggy voltage levels that 4000-series CMOS logic sometimes provides. The output of B directly controls one FET, and inverter C develops the anti-phase control voltage for the FET of opposite polarity, which naturally requires an inverted gate voltage to turn it on or off.

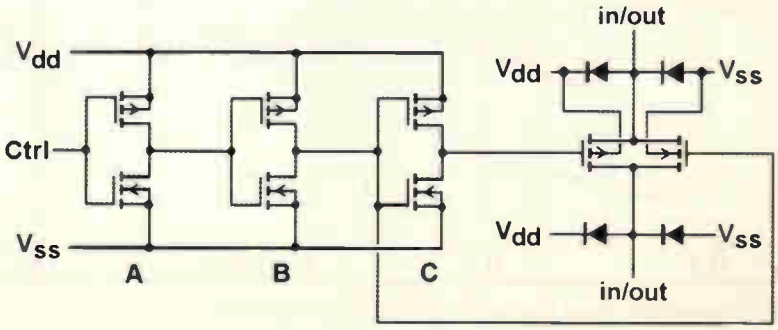

Fig 1. The internal circuitry of a 4000-series analogue gate.

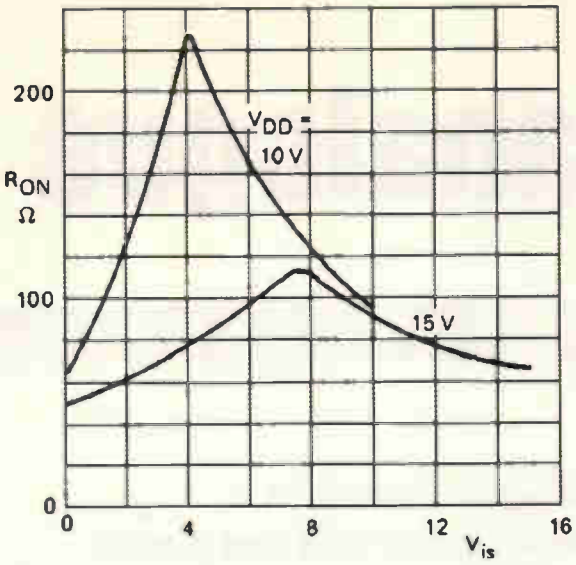

Fig 2. Typical variation of the gate series resistance Ron.

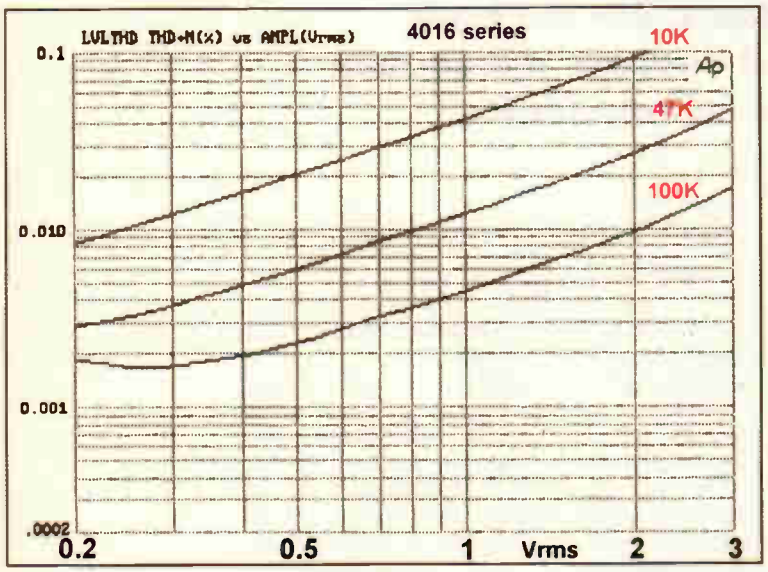

Fig 4. 4016 series-gate THD versus level, with different load resistances.

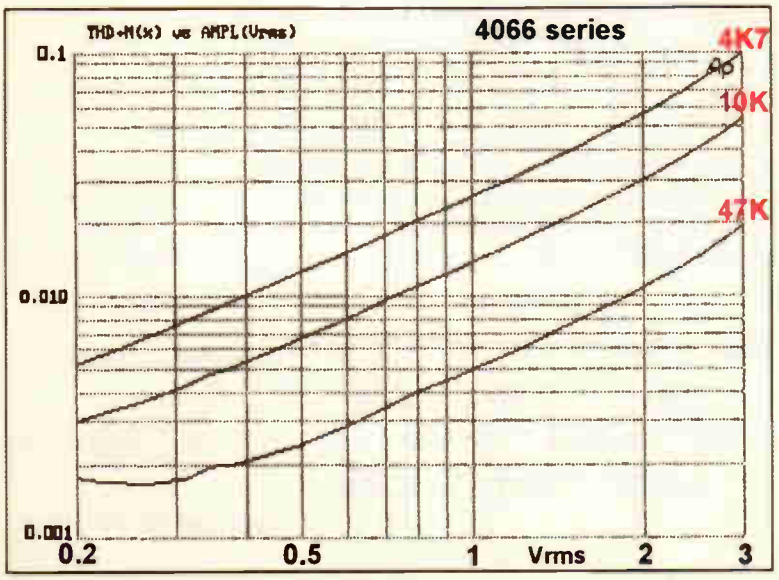

Fig 5. 4066 THD versus level, with different load resistances.

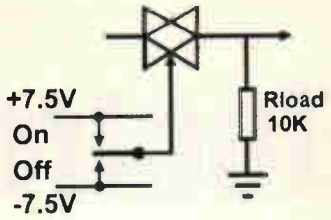

Fig 3. Voltage-mode series switching circuit using analogue gate.

MOS FETS are of the enhancement type, requiring a voltage to be applied to the gate to turn them on; (in contrast JFETs work in depletion mode and require gate voltage to turn them off) so the closer the channel gets to the gate voltage, the more the device turns off. An analogue gate with only one polarity of FET would be of doubtful use because R_{on} would become very high at one extreme of the voltage range. This is why complementary FETs are used; as one polarity finds its gate voltage decreasing, turning it off, the other polarity has its gate voltage increasing, turning it more on. It would be nice if this process cancelled out so the R_{on} was constant, but sadly it just doesn't work that way. **Figure 2** shows how R_{on} varies with input voltage, and the peaky R_{on} curve gives a strong hint that something is turning on as something else turns off.

Figure 2 also shows that R_{on} is lower and varies less when the higher supply voltage is used; since these are enhancement FETs the on-resistance decreases as the available control voltage increases. If you want the best linearity then always use the maximum rated supply voltage.

Since R_{on} is not very linear, the smaller its value the better. The 4016 R_{on} is specified as 115 Ohms typical, 350 Ohms max, over the range of input voltages and with a 15V supply. The 4066 is a version of the 4016 with lower R_{on}, 60 Ohms typical, 175 Ohms max under the same conditions. This option can be very useful both in reducing distortion and improving offness, and in most cases there is no point in using the 4016. The performance figures given below assume the use of the 4066 except where stated.

CMOS gates in voltage mode

Figure 3 shows the simplest and most obvious way of switching audio on and off with CMOS analogue gates. This series configuration is in a sense the 'official' way of using them; the snag is that by itself it doesn't work very well.

Figure 4 shows the measured distortion performance of the simple series gate using the 4016 type. The distortion performance is a long way from brilliant, exceeding 0.1% just above 2Vrms. These tests, like most in this section, display the results for a single sample of the semiconductor in question. Care has been taken to make these representative, but there will inevitably be some small variation in parameters like R_{on}. This may be greater when comparing the theoretically identical products of different manufacturers.

Replacing the 4016 gate with a 4066 gives a reliable improvement due to the lower R_{on}. THD at 2Vrms (10K load) has dropped to a third of its previous level. There seems to be no downside to using 4066 gates instead of the more common and better-known 4016, and they are used exclusively from this point on.

The distortion is fairly pure second harmonic, except at the highest signal levels where higher-order harmonics begin to intrude. This is shown in **Figs 5** and **6** by the straight line plots beginning to bend upwards above 2Vrms.

Analogue gate distortion is flat with frequency from

10Hz up to 30kHz at least, and so no plots of THD versus frequency are shown; they would merely be a rather uninteresting set of horizontal lines.

This circuit gives poor offness when off, and poor distortion when on. The offness is limited by the stray capacitance in the package feeding through into the relatively high load impedance. If this is 10K the offness is only -48dB at 20kHz, which would be quite inadequate for many applications. The load impedance could be reduced below 10K to improve offness- for example, 4K7 offers about a 7dB improvement- but this degrades the distortion, which is already poor at 0.055% for 3Vrms, to 0.10%.

Using 4066 gates instead of 4016 does not improve offness in this configuration. The internal capacitance that allows signals to leak past the gate seems to be the same for both types.

The maximum signal level that can be passed through (or stopped) is limited by the CMOS supply rails and conduction of the protection diodes. While it would in some cases be possible to contrive a bootstrapped supply to remove this limitation, it is probably not a good route to head down.

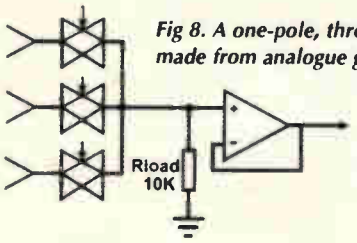

Fig 8. A one-pole, three way switch made from analogue gates.

Figure 8 shows a CMOS three-way switch. When analogue gates are used as a multi-way switch, the offness problem is much reduced, because capacitative feed-through of the unwanted inputs is attenuated by the low R_{on} looking back into the (hopefully) low impedance of the active input, such as an op-amp output. If this is not the case then the crosstalk from nominally off inputs can be serious. In this circuit the basic poor linearity is unchanged, but since the crosstalk problem is much less, there is often scope for increasing the load impedance to improve linearity. This makes R_{on} a smaller proportion of the total resistance. The control voltages must be managed so that only one gate is on at a time, if there is a possibility of connecting two op-amp outputs together.

It may appear that if you are implementing a true changeover switch, which always has one input on, the resistor to ground is redundant, and just a cause of distortion. Omitting it is however very risky, because if all CMOS gates are off together even for an instant, there is no DC path to the op-amp input and it will register its displeasure by snapping its output to one of the rails. This does not sound nice.

Figure 9 shows the offness of a changeover system, for two types of FET-input op-amps. The offness is much improved to -87dB at 20kHz, an improvement of 40dB over the simple series switch; at the high-frequency end however it still degrades at the same rate of 6dB/octave. It is well known that the output impedance of an op-amp with negative feedback increases with frequency at this rate, as the amount of internal gain falls, and this effect is an immediate suspect. However, there is actually no detectable signal on the op-amp output, (as shown by the lowest trace) and is also not very likely that two completely different op-amps would have exactly the same output impedance. I was prepared for a subtle effect, but the true explanation is that the falling offness is simply due to feed-through via the internal capacitance of the analogue gate.

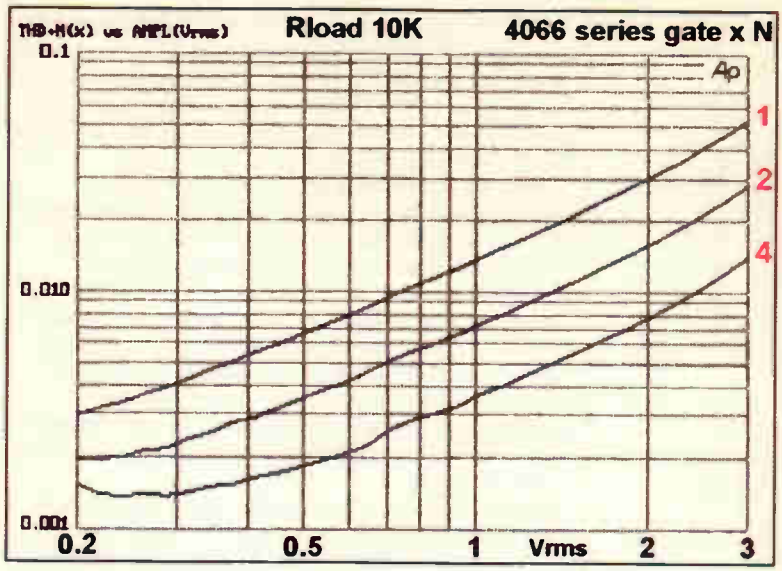

Fig 6. THD versus level, for different numbers of series 4066 gates.

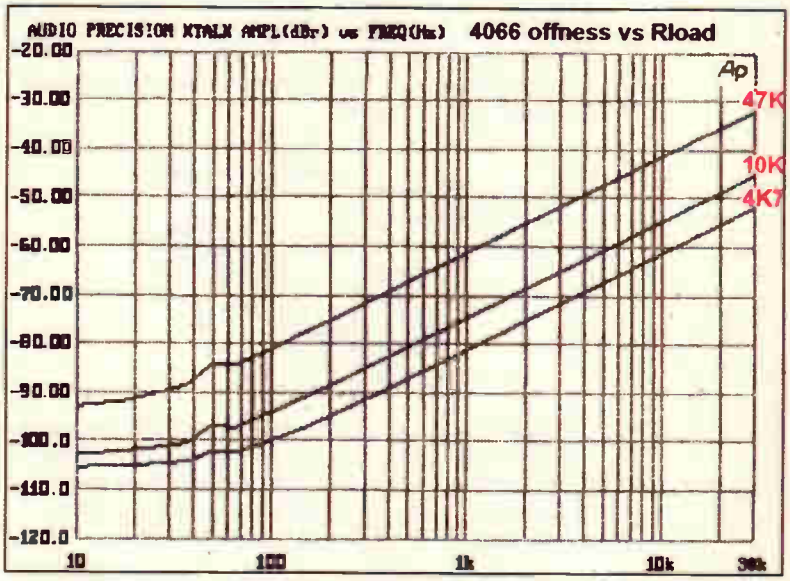

Fig 7. Offness versus load resistance. -48 dB at 20 kHz with a 10K load.

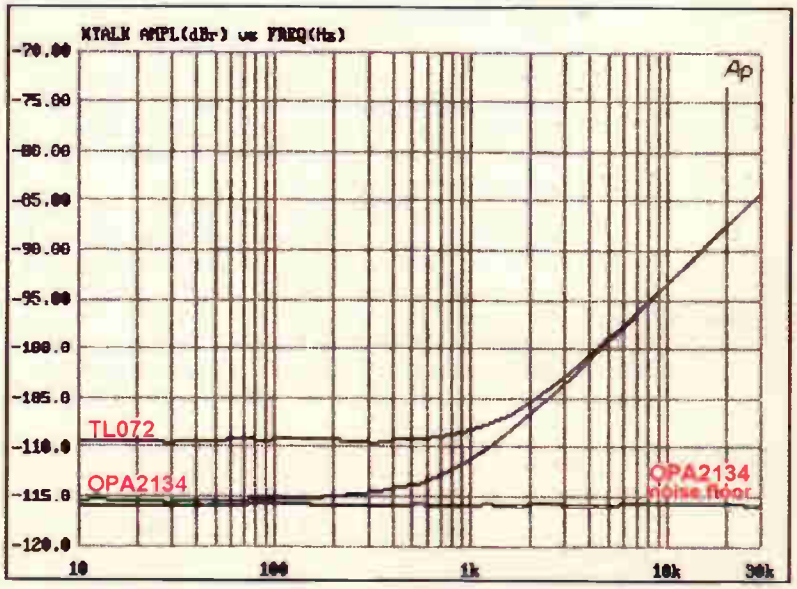

Fig 9. Voltage-mode changeover circuit offness for TL072 and OPA2134. Rload = 10K.

It now remains to explain why the OPA2134 apparently gives better offness in the flat low-frequency region. In fact it does not; the flat parts of the trace represent the noise floor for that particular op-amp. The OPA2134 is a more sophisticated and quieter device than the TL072, and this is reflected in the lower noise floor.

There are two linearity problems. Firstly, the on resistance itself is not totally linear. Secondly, and more serious, the on resistance is modulated when the gates move up and down with respect to their fixed gate voltages.

It will by now probably have occurred to most readers that an on/off switch with good offness can be made by making a changeover switch with one input grounded. This is quite true, but since much better distortion performance can be obtained by using the same approach in current mode, as explained below, I am not considering it further here.

Figure 10 shows a shunt muting circuit. This gives no distortion in the 'ON' state because the signal is no longer going through the R_{on} of a gate. However the offness is limited by the R_{on}, forming a potential divider with the series resistor R; the latter cannot be very high in value or the circuit noise will be degraded. There is however the advantage that the offness plot is completely flat with frequency. Note that the ON and OFF states of the control voltage are now inverted.

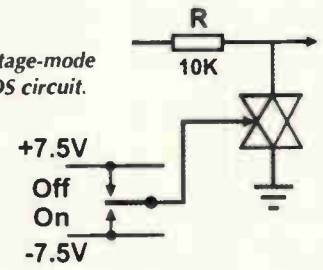

Fig 10. Voltage-mode shunt CMOS circuit.

Table 1 below gives the measured results for the circuit, using the 4066. The offness can be improved by putting two or more of these gates in parallel, but since doubling the number N only gives 6dB improvement, it is rarely useful to press this approach beyond four gates.

CMOS gates in current mode

Using these gates in current mode - usually by defining the current through the gate with an input resistor and dropping it into the virtual earth input of a shunt feedback amplifier - gives much superior linearity. It removes the modulation of channel resistance as the gate goes up and down with respect to its supply rails, and in its more sophisticated forms, can also remove the signal voltage limit and improve offness.

Figure 11 shows the simplest version of a current-mode on/off switch, and it had better be said at once that it is a bit too simple to be very useful as it stands. An important design decision is the value of R_{in} and R_{nfb}, which are often equal to give unity gain. Too low a value increases the effect of the non-linear R_{on}, while too high a value degrades offness, as it makes the gate stray capacitance more significant and also increases Johnson noise. In most cases 22K is a good compromise.

Table 1

N gates	Offness
1	–37dB
2	–43dB
4	–49dB

Table 2

	1kHz	10kHz	20kHz
THD via 4016, +20 dBu	0.0025%	0.0039%	0.0048%
THD: 4016 shorted, +20dBu	0.0020%	0.0036%	0.0047%
Offness	–68dB	–48dB	–42dB

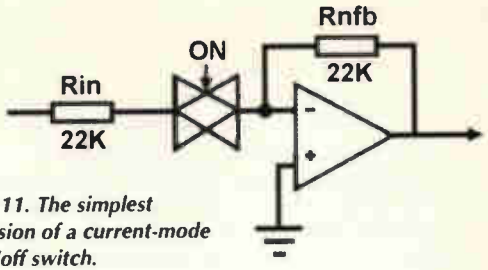

Fig 11. The simplest version of a current-mode on/off switch.

Table 2 gives the distortion for +20dBu (7.75Vrms) in/out, and shows that it is now very low compared with voltage-mode switchers working at much lower signal levels; compare Figs 5 and 6. The increase in THD at high frequencies is due to a contribution from the op-amp. However, the offness is pretty poor, and would not be acceptable for most applications. The problem is that with the gate off, the full signal voltage appears at the gate input and crosstalks to the summing node through the package's internal capacitance. In practical double-sided PCB layouts the inter-track capacitance can usually be kept very low by suitable layout, but the internal capacitance is inescapable.

In **Figs 11** and **12**, the CMOS gate is powered from a maximum of +/-7.5V. This means that in Fig 11, signal breakthrough begins at an input of 5.1Vrms. This is much too low for op-amps running off their normal rail voltages, and several dB of headroom is lost.

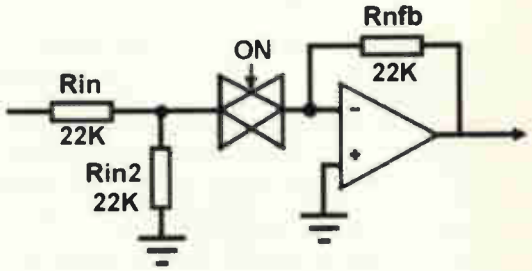

Fig 12. Current-mode switch circuit with breakthrough prevention resistor Rin2.

Figure 12 shows a partial cure for this. Resistor R_{in2} is added to attenuate the input signal when the CMOS gate is off, preventing breakthrough. There is no effect on gain when the gate is on, but the presence of Rin2 does increase the noise gain of the stage.

Series-shunt current mode

We now extravagantly use two 4016 CMOS gates, as shown in **Fig 13**.

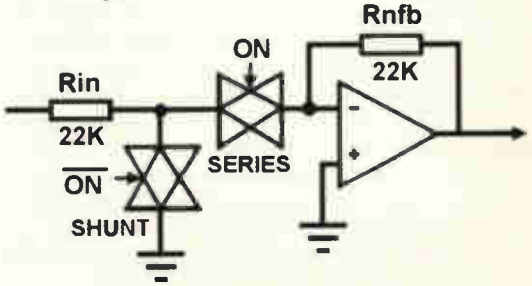

Fig 13. A series-shunt current-mode switch.

When the switch is on, the series gate passes the signal through as before; the shunt gate is off and has no effect. When the switch is off the series gate is off and the shunt

gate is on, sending almost all the signal at A to ground so that the remaining voltage is very small. The exact value depends on the 4016 specimen and its R_{on} value, but is about 42dB below the input voltage. This deals with the offness (by greatly reducing the signal that can crosstalk through the internal capacitance) and also increases the headroom by several dB, as there is now effectively no voltage signal to breakthrough when it exceeds the rails of the series gate.

Two antiphase control signals are now required. An excellent way to generate the inverted control signal is to use a spare analogue gate as an inverter, as shown in **Fig 14.**

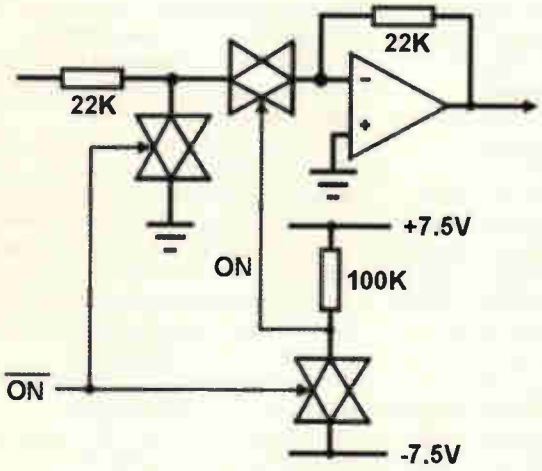

Fig 14. Generating the control signals with a spare analogue gate.

The distortion generated by this circuit can be usefully reduced by using two gates in parallel for the series switching, as in **Table 3** above; this gate-doubling reduces the ratio of the variable R_{on} to the fixed series resistor and so improves the linearity. Using two in parallel is sufficient to render the distortion negligible. (The higher distortion figures at 10kHz and 20kHz are due to distortion generated by the TL072 op-amp used in the measurements)

As before the input and output levels are +20dBu, well above the nominal signal levels expected in op-amp circuitry; measurements taken at more realistic levels would show only noise.

Table 3

	1kHz	10kHz	20kHz
THD via 4016 x 1, +20 dBu	0.0016%	0.0026%	0.0035%
THD via 4016 x 2, +20dBu	0.0013%	0.0021%	0.0034%
THD via 4016 shorted, +20dBu	0.0013%	0.0021%	0.0034%
Offness 4016 x 1	–109dB	–91dB	–86dB
Offness 4016 x 1, J111	Less than –116dB	–108dB	–102dB

Discrete FETs have lower R_{on} than analogue gates. If a J111 JFET is used as the shunt switching element the residual signal at A is further reduced, to about 60dB below the input level, with a consequent improvement in offness, demonstrated by the final entry in Table 3. This could also be accomplished by using two or more CMOS gates for the shunt switching.

There is more on discrete FETs in part two of this article.

Control voltage feedthrough in CMOS gates

When an analogue gate changes state, some energy from the control voltage passes into the audio path via the gate-channel capacitance of the switching FETs, through internal package capacitances, and through any stray capacitance designed into the PCB. Since the control voltages of analogue gates move snappily, due the internal inverters, this typically puts a click rather than a thump into the audio. Attempts to slow down the control voltage going into the chip with RC networks are not likely to be successful for this reason. In any case, slowing down the control voltage change simply converts a click to a thump; the FET gates are moving through the same voltage range, and the feed-through capacitance has not altered, so the same amount of electric charge has been transferred to the audio path - it just gets there more slowly.

The only certain way to reduce the effect of transient feed-through is to soak it up in a lower value of load resistor. The same electric charge is applied to a lower resistor value (the feed-through capacitance is tiny, and controls the circuit impedance) so a lower voltage appears. Unfortunately reducing the load tends to increase the distortion, as we have already seen; the question is if this is acceptable in the intended application. ■

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10-bit composite/component analogue video to SDI converter

The device presented here is a vital part of every studio infrastructure that is in transition from analogue to digital video formats. It takes as its input composite or component (YUV) analogue video formats and translates it to a Serial Digital Interface (known as the SDI standard – in the case of standard definition video). Emil Vladkov sets out his stall

The SDI outputs are digital video outputs, which are not compressed, so the bit-rate is high – 270Mbps. This bitstream can be supplied to any MPEG-compressor for reducing the bit-rate to a value suitable for transmission or statistical multiplex. The main advantage of using intra-studio uncompressed formats is that they are not subject to generation degradation. It means you can copy and move the material between storage media without any degradation of quality indefinitely. The use of such converters (known in the professional world also as 'decoders' because they decode the composite coded analogue video) was not possible in a pure analogue studio environment; otherwise they would have to build digital 'islands' in the analogue 'ocean'. As the infrastructure, distribution and routeing moved to digital, analogue-to-digital converters became a necessity, because many of the feeds of a typical studio are from analogue sources. It is unlikely that in the next few years (tens of years perhaps!) there will be no more analogue video sources, so the problem of feeding the composite and component signals in a digital studio will probably exist in the near future too. Although the components used in the design are not exactly cheap because of their professional origin, the project is relatively easy to implement and can solve the problems of small broadcast operators and individual video professionals, who are not willing to pay the price for a brand name equipment in the range of 1500 USD and above (especially in small quantities).

The internal structure of the video decoder

The main building blocks of the composite/component to SDI decoder are presented in Fig. 1. The device incorporates a video analogue-to-digital converter (dedicated IC[1]), which outputs 10-bit parallel words and a clock signal at 27MHz according to the standard ITU-R BT.601 / CCIR656. This parallel stream of digitised uncompressed video is fed to a video serialiser with an integrated cable driver and the last one outputs a valid SDI-stream. Actually there are two equivalent SDI-outputs, which is a feature of the integrated circuit used, so the CSDI10 serves also as a distribution amplifier. The whole device is controlled by an 8-bit 8051-based microcontroller. Actually the controller loads the configuration information to the analogue-to-digital converter. The serialiser is a pure hardware based architecture and so needs no configuration. The microcontroller has the following functions in this design:

- Loading configuration data to the A/D converter, which enables or disables specific features
- Storing the user configuration in an internal EEPROM, so the device boots automatically with the last valid configuration without the need of a new setting-up at every power-up;
- Communicating with a host PC for loading configuration data through a friendly user interface.

The last building block of the device is the integrated power supply, which accepts input power from an external adapter (wall cube).

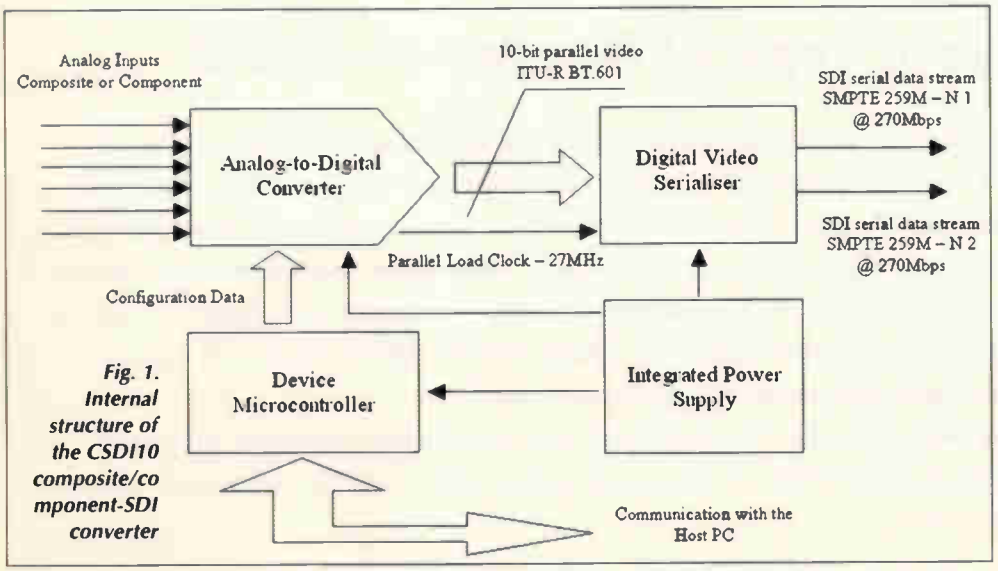


Fig. 1. Internal structure of the CSDI10 composite/component-SDI converter

Detailed schematics of the decoder

The circuit diagram of the analogue video to SDI decoder is presented in Fig. 2. There are six analogue inputs to the Analog Devices Professional NTSC/PAL Video decoder ADV7185 (U1). Actually only the Y1, U1, V1 inputs are used, but all others are also provided as ports on the PCB for greater design versatility. The inputs are marked J1 –J6 and are all terminated with 75Ω resistors (R13, R14, R15, R16, R17 and R23). All inputs are AC coupled to the A-D IC

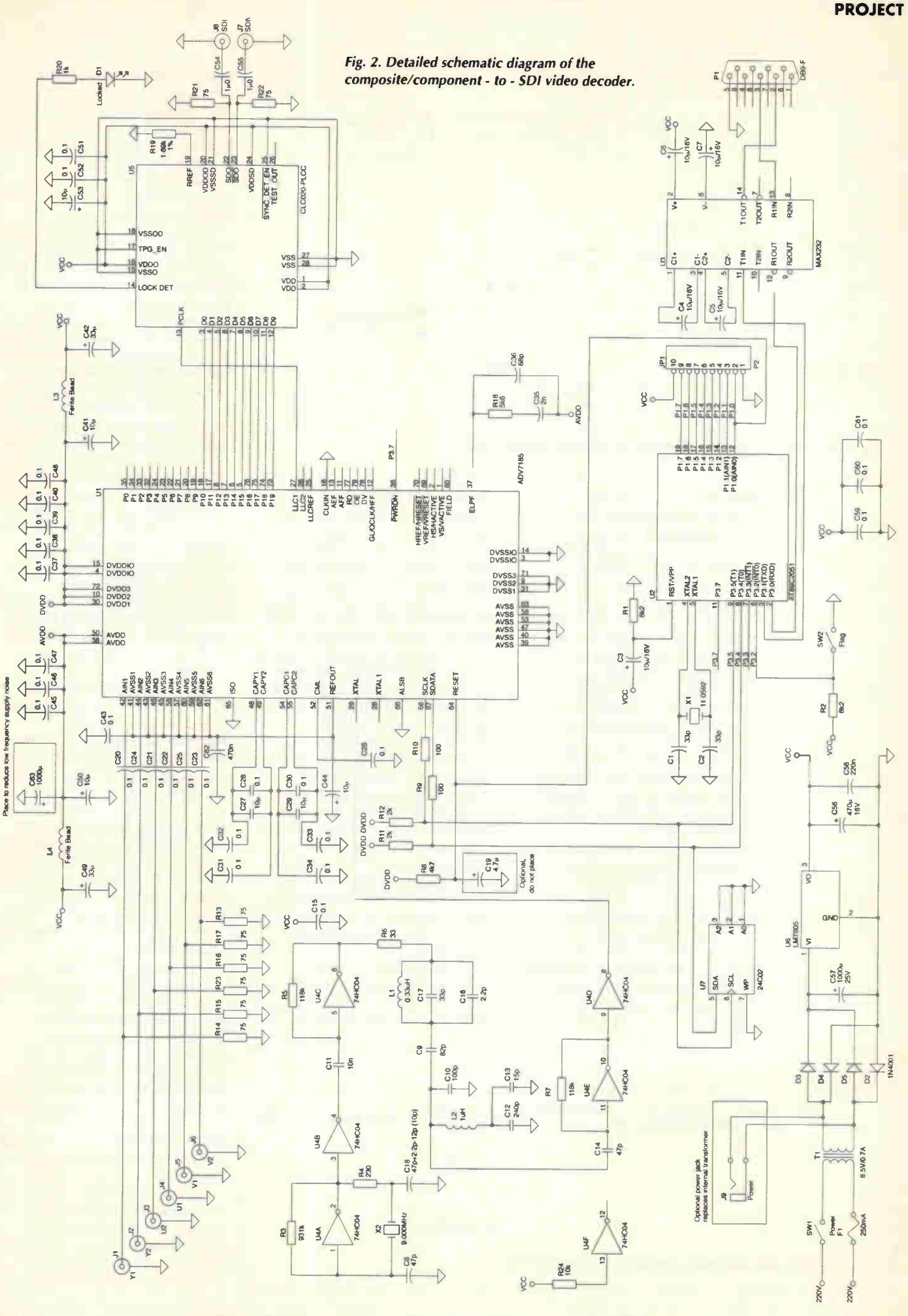


Fig. 2. Detailed schematic diagram of the composite/component - to - SDI video decoder.

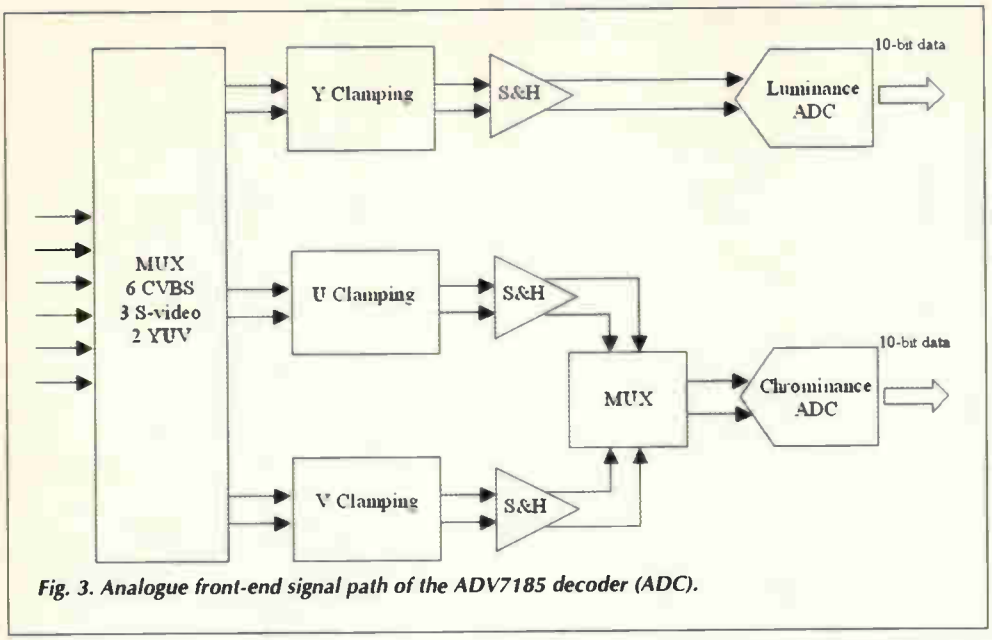


Fig. 3. Analogue front-end signal path of the ADV7185 decoder (ADC).

through C20 – C25. If single ended mode is implemented on the inputs (as is the case) the 6 analogue ground pins attached to every input AVSS1 – AVSS6 should be connected as shown to REFOUT and AC-connected to the device ground connection. The Analogue-to-Digital Converter capacitor networks are implemented as proposed by the manufacturer and consist of the C27 – C34 capacitors. The Input Switch Over (ISO-pin on the ADV7185), which is normally edge triggered, is not used in this design (connected to GND), as it is assumed that no frequent changes of the input video source will occur, so no reacquiring of the timing information will be necessary. The luminance/chrominance parallel multiplexed data is available on outputs P10-P19 of U1, which are connected directly to the parallel inputs of the serialiser circuit U5 – CLC020. The Line Locked Clock (LLC1 on U1) is a PLL stabilised version of the 27MHz clock feeding the A-D converter and is locked to the parallel data output. This clock output is used to clock-in the serialiser parallel data and so it is connected to the PCLK-pin of U5. As the LLC output to the serialiser is PLL filtered, the PLL External Loop Filter is implemented around the components R18, C35 and C36.

Special attention should be paid to the ADV7185 which has split power supplies – one for the analogue section, one for the digital section and one for the I/O interface. These power supplies should be provided to the device on separate paths, which are capacitor- and ferrite bead-filtered. The corresponding components are beads L3 and L4 and capacitors C37-C42, C45-C50. Although the manufacturer suggests using +5V for the analogue supply and +3.3V for all other digital supply voltages, in the design a common +5V supply is used. This results in somewhat greater power dissipation from the ADV7185 but the problem can be overcome by a small heatsink, mounted on top of the integrated circuit U1.

Clock is supplied to the A/D converter through the XTAL-pin of U1. The 27MHz crystal stabilized clock is produced by the 9MHz crystal oscillator built around the U4A and U4B buffers with X2, R3, R4, C8 and C18 passives, and the frequency tripler around U4C, U4D and U4E with the associated passive components. This idea of a 3x frequency multiplier is borrowed from the February 2001 issue of *EW* – the *Circuit Ideas* section[4].

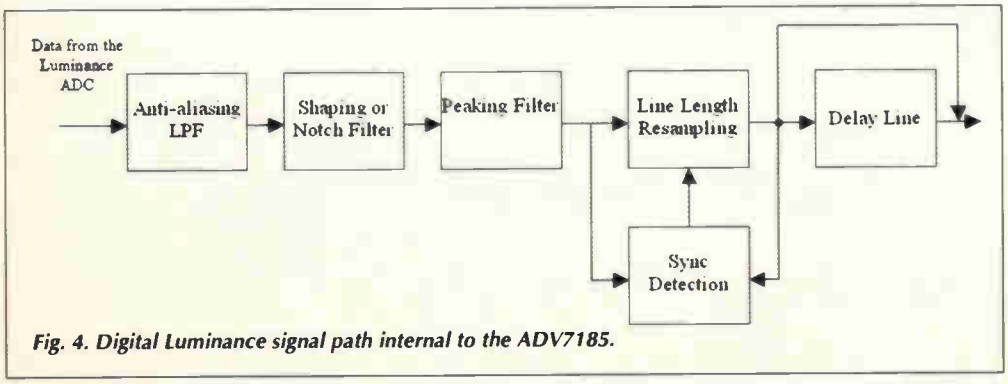


Fig. 4. Digital Luminance signal path internal to the ADV7185.

The digital video serialiser circuit U5 provides two complementary outputs SDO and SDO\, but as the SDI stream is NRZI-encoded it is polarity insensitive. So the two complementary outputs are fully equivalent SDI outputs. The R21 and R22 75Ω resistors are back matching resistors. The device manufacturer (National Semiconductor) states that no series back-matching resistors should be used with this design[2]. The SDI-signal is AC-coupled (C54 and C55) into the output BNC connectors (J7 and J8). When the internal PLL of the CLC020, which works at 10x the parallel load frequency of 27MHz (which is 270MHz), has locked to the parallel input data stream, the LOCK_DET output is activated, so the user is provided with this information through the D1 LED (and the associated R20 current limiting resistor). So the normal functioning of the device can be monitored visually.

The programming of the A/D converter is accomplished through the 2-Wire I^2C compatible microprocessor bus interface. The signals used are SCLK (Serial Clock) and SDATA (Serial Data). The same serial bus is used by the device microcontroller U2 (AT89C2051) to access the EEPROM-memory U7, where the configuration information is nonvolatile stored. With the ALSB-pin of the U1 set to Ground level the address at which the ADV7185 responds to the serial bus is 88 hex. The EEPROM memory responds at address A0 hex.

Although the Reset of the U1 converter can be accomplished by the use of a simple R-C network, I occasionally experienced some problems with a proper reset sequence. So I modified the first circuit diagram, so that RESET is provided by the microcontroller with the P1.0 line. The U2 microcontroller waits about 1 sec after power is applied to the system and then applies a low 100ms reset pulse to the ADV7185.

The firmware of the Composite/Component to SDI converter resides in the internal EEPROM of the microcontroller. The 2K-version micro used proved absolutely adequate to store the

whole firmware program. The microcontroller is clocked at 11.0592 MHz (X1 with C1 and C2) and is reset at power-up by R1-C3 network. The P2 port is provided for device extension purposes. The Host-PC communications are accomplished by the on-board integrated serial RS-232 compatible interface and the U3 RS-232 level translator with the corresponding components C4, C5, C6 and C7. The RS-232 port is available on the CSDI10 front panel in the shape of a DB-9 female connector (P1). Once configuration of the device different from the factory presets is needed, the only thing you need is to connect a standard serial cable to the PC and run the Graphical User Interface (GUI) program on the PC (more on this later).

Power can be supplied to the CSDI10 in 2 ways – by the use of an onboard power transformer or by an external power adapter, connected to the J9 power supply jack. The last one is the way used in the actual design – it has the advantage of being internationally universal: you can use the same CSDI10 module with different external power supply adapters in different mains standard countries around the world. The power jack input can be either polarity, which is accomplished by the use of D2, D3, D4 and D5 diodes. The external raw supply is filtered and regulated on-board by U6 three terminal regulator LM7805 and passive components C56, C57 and C58.

Choosing between the two input options (CVBS – composite and YUV – component) is done by correctly setting the SW2 (flag P3.2 of the U2 microcontroller) at power-up. If the switch is open (the default configuration) the device goes in component mode. Otherwise (switch closed) the composite mode is entered.

Main functions of the ADV7185 NTSC/PALvideo decoder

The ADV7185 consists of an analogue front-end and a digital processing section[1]. The analogue front-end (depicted in Fig. 3) has a six input multiplexer, which can be configured to accept different format channels – these are six CVBS, three S-video or two YUV component channels. There are three clamping circuits for DC restoration and three sample and hold amplifiers in front of the two Analogue-to-Digital Converters. As the standards for video digitising state that the chrominance information occupies

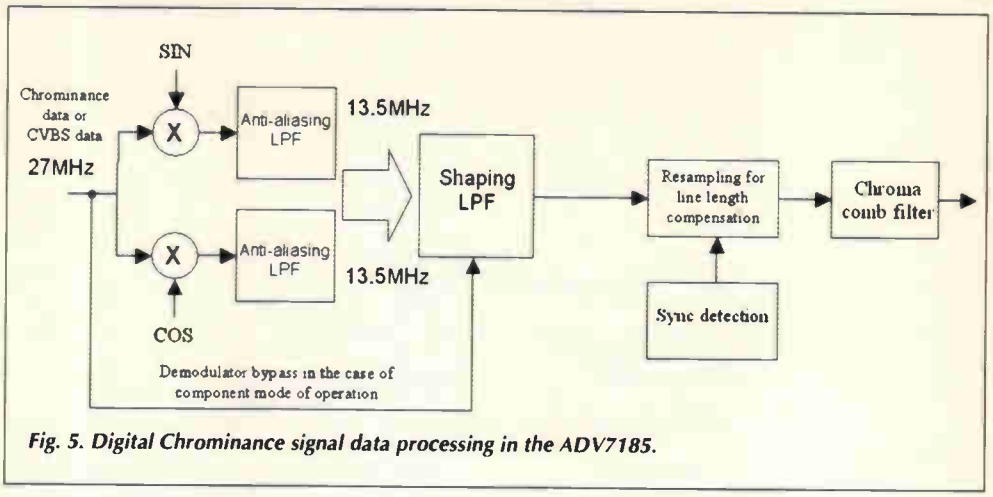


Fig. 5. Digital Chrominance signal data processing in the ADV7185.

two times less bandwidth than the luminance information and so is coded in the 4:2:2 manner (one Y sample, one Cr sample, one Y sample, one Cb sample), the second A/D converter has multiplexed inputs – it converts both chroma channels at half sampling rate each. All analogue video processing can be done in single-ended and differential approach (the last one obtaining the highest fidelity possible). In the current design the single-ended approach is used.

The luminance path of the signal after passing the A/D-converter continues in the ADV7185 as depicted in **Fig. 4**. First, a low-pass anti-aliasing filter is applied to the digital stream, then the data is passed to shaping and notch filters. The shaping filter is used in the component mode of operation and it is basically a low-pass type filter with different cut-off frequencies (shapes), which can be configured in software. In the composite mode of operation the notch filter is used in place of the shaping one with the purpose of removing the unwanted chrominance information, overlaid on the luma signal around the colour subcarrier frequency. The peaking filter follows next on the digital signal path and it has the function of controlling the image luminance sharpness. This filter can be bypassed if not needed. A resampler follows with outputs exactly 720 pixels per line for both PAL and NTSC, so that line length variations in the input video can be compensated. The resampler is controlled by a sync-detection block. The end of the digital luminance path is a 2-line delay line, which is switched on if the comb filter is used in the chrominance data path. In this case the chrominance data is delayed in respect to the luma information, so a delay in the luminance path has to be used to provide the correct output.

The chrominance data path depicted in **Fig. 5** operates on the Y ADC data in the case of the composite mode of operation and on the chrominance data (U, V) in the case of component mode. The data is demodulated through multiplication with a local oscillator (two multipliers in quadrature) and then low-pass filtered to remove the higher order products of the synchronous detection. The demodulation is bypassed in component mode. The demodulated/pure chrominance data is then passed to an anti-aliasing filter and then to a shaping filter. As with the luminance data this shaping filter is a low-pass filter with selectable cut-off frequency, so that the bandwidth of the chroma data can be limited according to the user needs. A resampler follows, which corrects for variations in the line length of the video line. The last stage in the data path of the chrominance is the comb filter, which is used to better separate chrominance and luminance data, allowing the luminance data to have greater bandwidth (no low-pass filtering of the luma information is needed in this case prior to composite encoding).

The CLC020 digital video serialiser

The National Semiconductor CLC020 IC serialises the 10-bit (or 8-bit) parallel data, but also incorporates many additional functions necessary for the channel coding specified for the SDI interface[2]. This includes the data scrambling with the polynomial $1+X^4+X^9$ and the data format conversion from NRZ to NRZI (Non-Return-to-Zero-Inverted). It also incorporates a coaxial line driver circuit. The output is a serial digital stream compliant with the SMPTE 259M standard. The internal Test Pattern Generator (TPG) is not used with this design. The Sync_Detector

Table 1: Configuration Commands for the CSDI10 Video Decoder (All Commands end with CR + LF)

Command Group	Operation	Mnemonics
Default Configuration	Loads the factory preset configuration	DEFAULT
Standard Selection	Input Standard PAL (BGHID), (default)	R00D80 CVBS 89 YUV
	Input Standard NTSC (M)	R00D40 CVBS 49 YUV
	Auto Detect Input Standard	R00D00 CVBS 09 YUV
Video Quality Control	Broadcast Quality Input (default)	R01D08
	TV Quality Input	R01D09
	VCR Quality Input	R01D0A
	Surveillance Quality Input	R01D0B
Video Enhancement Control	Peaking +4.5dB boost	R02D02
	Peaking +1.25dB boost	R02D03
(Peaking Filter – luma peaking) around the color subcarrier)	No peaking (default)	R02D04
	Peaking -1.25dB attenuate	R02D05
	Peaking -1.75dB attenuate	R02D06
	Peaking -3.0dB attenuate	R02D07
Output Format Control	10 bit 4:2:2 CCIR 656 (default)	R03D00
	8 bit 4:2:2 CCIR 656	R03D0C
Contrast Control	Contrast adjustment 00-FF hex (80 – default)	R08Dxx (80)
Saturation Control	Saturation adjustment 00-FF hex (80 – default)	R09Dxx (80)
Brightness Control	Brightness adjustment 80-FF & 00-7F hex (00 – default)	R0ADxx (00)
Hue Control	Hue adjustment (00 – default)	R0BD00
Shaping Filter Control	Auto Narrow Notch Shaping Filter (default)	R17D01
	Auto Wide Notch Shaping Filter	R17D00
Comb Filter Control	Chroma Comb Adaptive 2 Lines (default)	R19D18
	Chroma Comb Adaptive 1 Line	R19D14
	Chroma Comb Non-adaptive 2 Lines	R19D08
	Chroma Comb Non-adaptive 1 Line	R19D04
	No Chroma Comb Filter	R19D00
Pixel Delay Control	Chroma +2 pixels (early)	R27D48
	Chroma +1 pixel (early)	R27D50
	No delay between luma and chroma samples (default)	R27D58
	Chroma -1 pixel (late)	R27D60
	Chroma -2 pixels (late)	R27D68
	Chroma -3 pixels (late)	R27D70

is enabled in the design, so the synchronizing words in the digital stream (parallel format) – the Timing Reference Signal, are recognized in a valid video signal. LSB clipping is performed in cooperation with the Sync_Detector, so all data values in the TRS in the range 000h – 003h are forced to 000h and all TRS values in the range 3FCh – 3FFh are forced to 3FFh. With such high sophistication and so many standard-defined functions available in the CLC020 the designer has nothing to worry about the standards compliance of the final device. A more detailed discussion on the Serial Digital Interface (SDI), the SDI-signal path and characteristics and the structure of the digital video stream can be found in an earlier article of mine, published in the February issue 2002 of *Electronics World*[3].

Programming the video decoder / Instruction Set

As mentioned earlier, the ADV7185 video decoder responds with a unique slave address on the I²C-bus. The master of this bus is the host microcontroller, running the firmware (for a copy of the firmware code in both object code and assembler code please contact *EW*). The ADV7185 is a highly integrated and complicated device with many functions, so its mode of operation is controlled through 60 dedicated registers, which are grouped by the manufacturer Analog Devices Inc. in a 'Basic Block' and in an 'Advanced Block'. Every register has a unique subaddress, which is a HEX-value (for example the 'Output Control Register' resides at location 03 hex). The microcontroller sends on the I²C-bus first the subaddress of the register accessed and then reads the data in the register or writes new configuration data to it.

The main strength of the design presented here is that its programming is future proof. Although the user is provided with a friendly user interface (described later), the communication port of the CSDI10 offers access to every register of the video decoder. Of course the user should obey a specific syntax of the commands sent through the COM-port. Basically information about the subaddress of the register and the register data is provided in a very simple ASCII-character form. The data message format is:

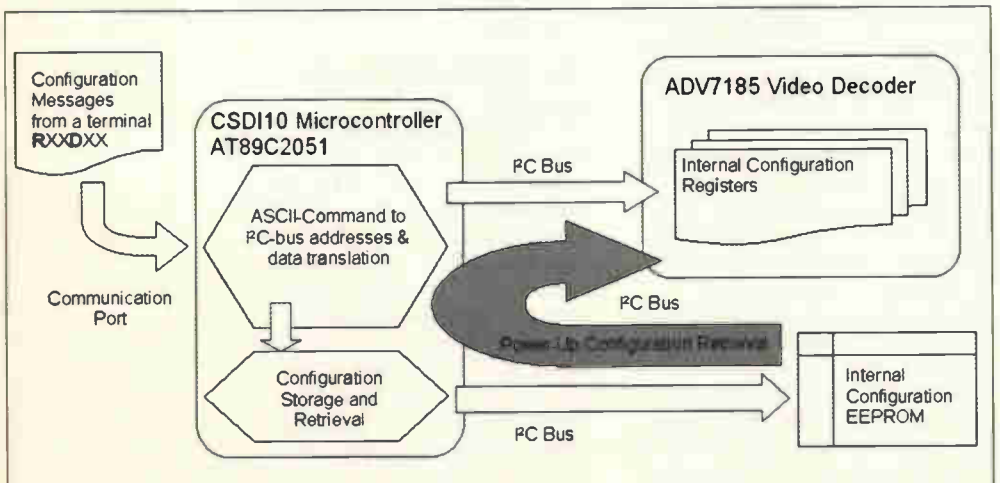


Fig. 6. Configuration Data Flow of the CSDI10 Video Decoder ('open system concept').

`RXXDXX + Carriage Return + Line Feed`

XX stands for the address (first) and data (second) fields in ASCII HEX representation. A typical valid example is `'R01D08 + Carriage Return + Line Feed'`. Not all possible register values are used by the GUI interface, it can be stated that the video decoder provides far more versatility and functions than used (and needed) in an actual implementation. The straightforward thing is that the firmware residing in the microcontroller does not limit the user to specific registers or register values. It translates every command with correct syntax received on the COM-port to an I^2C-bus register access to the ADV7185. So you can access every register and write to it every value. Not all values bear meaning; some of them will even disrupt the normal operation of the encoder (the GUI prevents you from making such annoying mistakes). This is the bad news, the good one is that if the need should arise in the future, to change the modes of operation of the CSDI10, you will not need to change the hardware, you will not need even to open the device to upgrade the firmware. The only necessary thing in this case will be to write the correct new configuration commands to the device (in every terminal program), which will automatically store the new data not only in the ADV7185 registers but also in the EEPROM configuration memory. So at next power-up you will have a completely new (and hopefully working) device, best suited to your needs. If something goes wrong you simply type the service command 'DEFAULT' and the device enters the 'good' factory programmed configuration. So the device has a very 'open' architecture. The terminal settings for accessing the device through the COM-port and a terminal program on the host PC are '9600,8,N,1'. Some of the most important configuration commands are presented in **Table 1.** The configuration is valid for the current state of the CVBS/YUV switch on the back panel of the device (this is equivalent to the SW2-flag in the circuit diagram in Fig. 2). Two independent configurations can be stored for CVBS-input mode and for YUV-input mode in the internal EEPROM, so depending on switch position one from these is loaded at power-up. The 'open system' concept and the configuration data flow of the CSDI10 are presented in **Fig. 6.**

All registers of the ADV7185 with their addresses and values for both

Table 2: Register values for composite and component modes of operation

Register Address	Register Description	Register Value	
		YUV	CVBS
00 hex	Input Control	89 hex	80 hex
01	Video Selection	08	08
02	Video Enhancement Control	04	04
03	Output Control	00	00
04	Extended Output Control	0C	0C
05	General Purpose Output	60	60
06	Reserved (Misc Control Register)	FF	FF
07	FIFO Control	04	04
08	Contrast Control	80	80
09	Saturation Control	80	80
0A	Brightness Control	00	00
0B	Hue Control	00	00
0C	Default Value Y	10	10
0D	Default Value C	88	88
0E	Temporal Decimation	00	00
0F	Power Management	00	00
10	Status Register (Read only)	FF	FF
11	Info Register (Read only)	FF	FF
12	Reserved	FF	FF
13	Analogue Control Internal	FF	FF
14	Analogue Clamp Control	18	18
15	Digital Clamp Control 1	60	60
16	Digital Clamp Control 2	00	00
17	Shaping Filter Control	01	01
18	Reserved	FF	FF
19	Comb Filter Control	FF	18
1A	Scaling/Cropping MSB	FF	FF
1B	Active Video Desired Lines	FF	FF
1C	Active Video Vertical Begin	FF	FF
1D	Vertical Scale Value 1	FF	FF
1E	Vertical Scale Value 2	FF	FF
1F	Active Video Horizontal Begin	FF	FF
20	Active Video Desired Pixels	FF	FF
21	Horizontal Scale Value 1	FF	FF
22	Horizontal Scale Value 2	FF	FF
23	Color Subcarrier Control 1	E5	EF
24	Color Subcarrier Control 2	43	FF
25	Color Subcarrier Control 3	00	FF
26	Color Subcarrier Control 4	41	FF
27	Pixel Delay Control	58	58
28	Manual Clock Control 1	00	00
29	Manual Clock Control 2	00	00
2A	Manual Clock Control 3	00	00
2B	Auto Clock Control	FF	FF
2C	AGC Mode Control	CE	CE
2D	Chroma Gain Control 1	FF	FF
2E	Chroma Gain Control 2	FF	FF
2F	Luma Gain Control 1	FF	FF
30	Luma Gain Control 2	FF	FF
31	Manual Gain Shadow Control 1	FF	FF
32	Manual Gain Shadow Control 2	FF	FF
33	Misc Gain Control	E3	E3
34	Hsync Position Control 1	0F	0F
35	Hsync Position Control 2	01	01
36	Hsync Position Control 3	00	00
37	Polarity Control	00	00
44	Resample Control	FF	FF
45	Reserved	FF	FF
F1	Reserved	FF	FF
F2	Reserved	FF	FF
32	Manual Gain Shadow Control 2	FF	FF
33	Misc Gain Control	E3	E3
34	Hsync Position Control 1	0F	0F
35	Hsync Position Control 2	01	01
36	Hsync Position Control 3	00	00
37	Polarity Control	00	00
44	Resample Control	FF	FF
45	Reserved	FF	FF
F1	Reserved	FF	FF
F2	Reserved	FF	FF

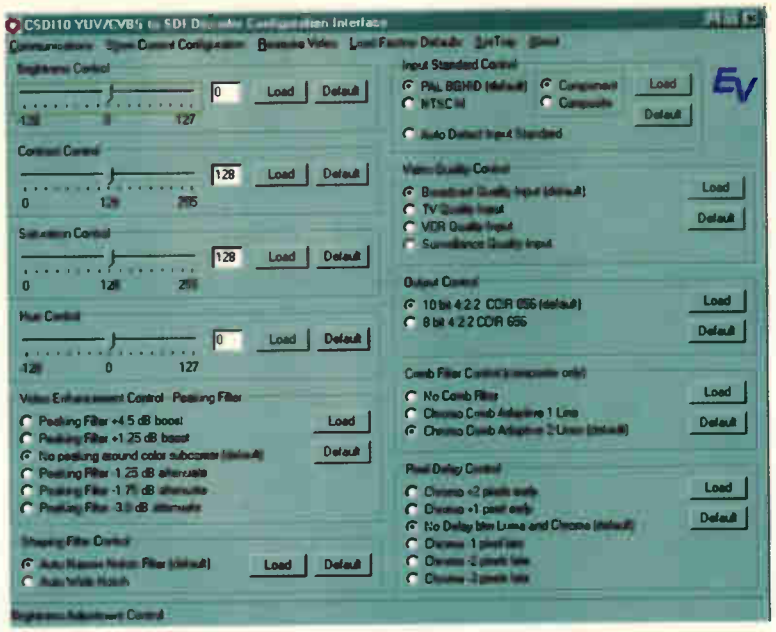


Fig. 7. The Graphical User Interface for Configuration of the CSDI10 Video Decoder.

modes of operation of the device – composite and component, are presented in **Table 2.** The values are based on a careful study of the ADV7185 data sheet, where every bit in every register is explained as a function. Some additional experimental work helped with finding the most suitable values for the power-on (default factory preset) configurations. The register values equal to FFh should not be modified from their initial state (which can be different from FFh!). Please use this table with the values given as a reference (together with the manufacturers datasheet), when you start experimenting with changing individual bits and bytes

GUI and CSDI10 configuration options

A snapshot of the GUI configuration program running on all Win9X platforms is presented in **Fig. 7.** From this figure all important configuration options and design features become visible, which are listed:

The Input Standard Control: The user can select both input standards: PAL and NTSC depending on the country in which the Video Decoder is implemented. A powerful option exists allowing the device to autodetect the incoming video standard. Component or Composite mode can be specified, but this setting will override the SW2-position at Power-Up.

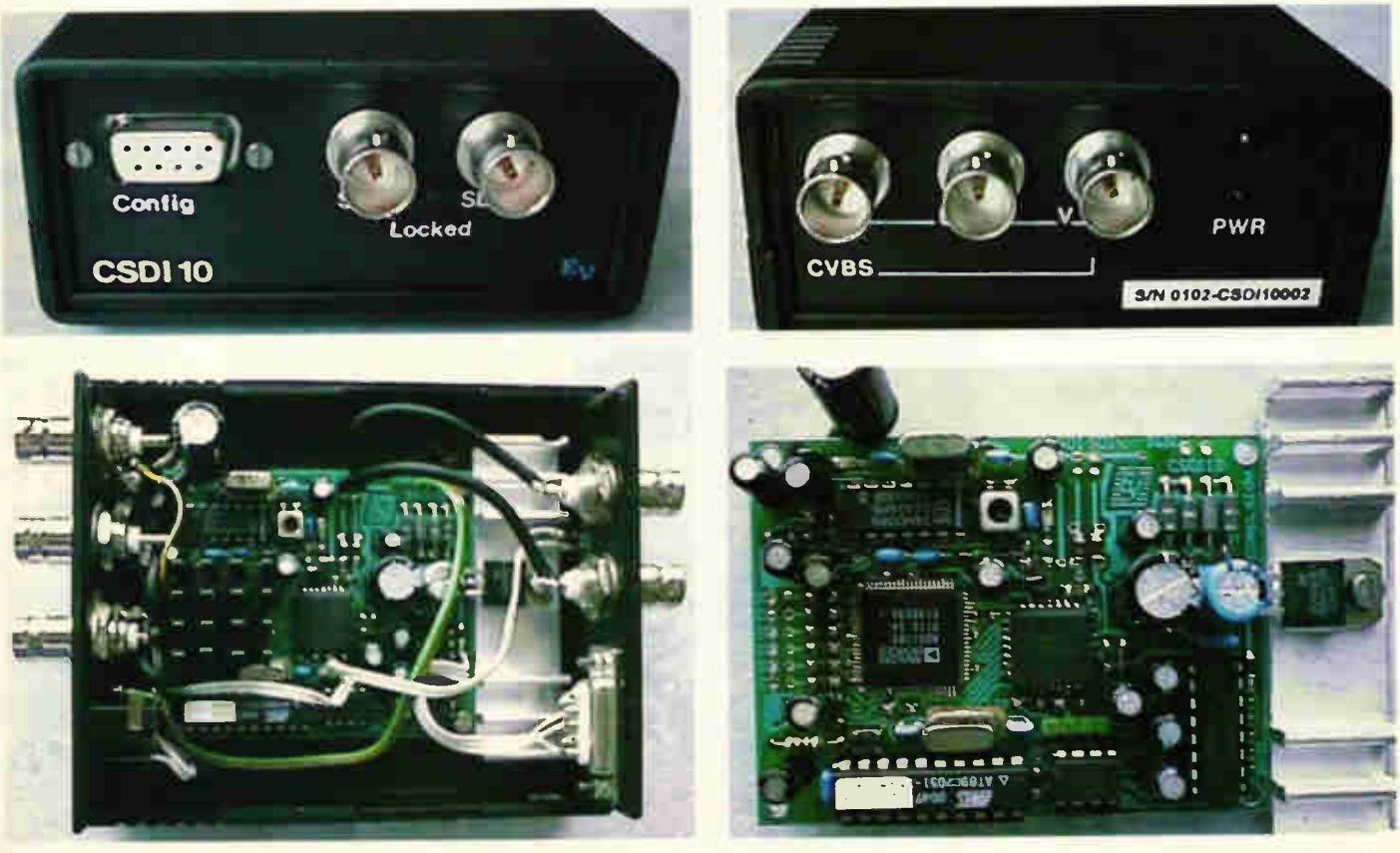
The Video Quality Control: The user can select between Broadcast (highest), TV, VCR or surveillance (lowest) quality.

The Output Control: 2 options are available, 8-bit and 10-bit (default) quantization of the analogue video signal. This conforms to CCIR 656.

The Comb Filter Control is available only for the composite mode of operation as it helps for better separation of chroma and luma information. The filter can act on 1 line or 2 lines of the video signal.

The Pixel Delay Control allows shifting the chroma data relative to the luma pixel data in both directions with 2-3 pixels. So the video data mismatches between luma and chroma can be compensated. In this case the Video Decoder acts as a video processing enhancer to the video signal.

Fig. 8. Photos of the CSDI10 Video Decoder author's prototype.



The Shaping Filter Control allows selection between Narrow and Wide notch filters.

The Video Enhancement Control applies different amounts of peaking (boost or attenuation) to the luma path of the video, allowing the user to compensate the effect of the colour subcarrier separation from the composite signal.

There are many sliders on the GUI for controlling the brightness, contrast, saturation and hue of the signal. The user should experiment with the different values (which are the decimal numbers, corresponding to the hex-representation register contents of the respective ADV7185 registers) of the controls to obtain the subjective best picture.

The program allows you to return to an original factory preset mode of operation through the 'Load Factory Defaults' menu, if the changes made are not satisfactory or resulted in improper device operation. The current device configuration can be displayed by choosing the 'Show Current Configuration' menu, but in this case the device should be powered down and up again (the CSDI10 outputs the contents of all its registers to the serial port at start-up). For a copy of the GUI-interface (warning: large files!) please contact *EW*.

Device assembly and availability

The photos of the assembled and cased CSDI10 composite/component-SDI decoder are presented in **Fig. 8.** The device can be housed in a plastic (shielded or unshielded depending on electromagnetic compatibility issues) case or in a 19'' rack mounted box (many decoders/encoders in a one 1RU case). The bare PCB can be ordered through *Electronics World* office at the price of £50. Please allow four weeks for production and delivery.

Note: chips are not inclded.

References

1. Analogue Devices, 'ADV 7185 Professional NTSC/PAL Video Decoder with 10-bit CCIR656 Output', Rev. PrF, 09/2001.
2. National Semiconductor, 'CLC020 SMPTE 259M Digital Video Serialiser with integrated cable driver, February 2000.
3. Vladkov, E., 'Analyse your SDI', *Electronics World*, February 2002, pp. 18-28.
4. Wells, J.N., 'CMOS tripler needs no adjustment', *Electronics World*, February 2001, p. 140.

The CSD110 technical specifications

Main Features

PAL/NTSC YUV or CVBS Video to ITU-R BT601/656 SDI component decoder;

Two high quality 10-bit video ADCs, 12-bit range;

Adaptive digital line-length tracking;

Multi-standard video input support – composite (CVBS) and component YUV;

Integrated automatic gain control & clamping – analogue and digital clamp control;

Integrated digital anti-aliasing low-pass filter;

Integrated shaping/notch and peaking filter;

Integrated two line chroma comb filter;

Programmable video source modes (through the RS-232 configuration interface): broadcast TV / VCR / camcorder / security / surveillance;

Digital output formats: YUV 4:2:2 CCIR 601/656 8-bit or extended 10-bit;

Programmable video controls – Pk-White / Hue / Brightness / Saturation / Contrast (through the RS-232 interface);

Locked Indicator on the Front/Back Panel;

Simple Reconfiguration through the integrated Serial RS-232 Interface and ASCII Terminal Commands / Configuration GUI;

Sampling

10 bit precision 4:2:2 to ITU-R BT.601/656

Analogue Inputs

Impedance	75Ω BNC
Input Return Loss	> 40dB to 5.5MHz

Video Serial Digital Outputs (2)

Applicable Standards	ITU-R BT.601/656
Format	EBU Tech 3267-E and SMPTE 259M-C
Number of Outputs	2 BNC
Output Impedance	75Ω
Return Loss	>15dB, 5-270MHz
Amplitude	800mV p-p unterminated
Data Rate	270Mbps
Data jitter	220ps at 270Mbps
Rise and Fall times	800ps
Drive Capability	Up to 250m (Belden 8281)

Input Standards

PAL-B/D/G/H/I/M/N, NTSC-M/N

Video Performance

Gain Stability	±0.05dB
S/N Ratio	> 60dB unweighted
Analogue front-end channel crosstalk	> 60dB
Hue accuracy	1.0 Degree
Colour Saturation Accuracy	1%
Colour Gain Control Range	-6 to +18 dB
Luma Brightness Accuracy	1%
Luma Contrast Accuracy	1%
Luminance Nonlinearity	0.7%
Chroma Non-Linear Gain	0.7%
Differential Gain	0.5%
Differential Phase	0.5 Degree

Frequency Response

Y	–0.1dB to 5.5MHz
Chroma	–0.1dB to 2.5MHz

Power Requirements

Power 9-12VDC, 600mA

Mechanical & Climatic

Height	45mm (1.75 inches)
Width	90mm (3.5 inches)
Depth	130mm (5 inches)
Weight	0.25Kg
Temperature	+50C to +350 C
Humidity	96% maximum

The practicalities of microwave cavities

The theory and practice of Cylindrical Microwave Cavities represented state of the art technology almost half a century ago. Resurrecting the need for this old technology brought to the fore a desperate shortage of practical guidance on even the most fundamental aspects of the subject. John Share explains

In this article experiences with cavities intended for operation at 9GHz and 14GHz are detailed with descriptions of the instrumentation developed to evaluate designs for our specific application.

The current or voltage distribution within a cavity is termed its 'mode'; this determines its radius and length. Our specific application required a point of maximum current at the centre of the cavity. As shown in **Fig. 1**, this specified a mode 'TE011' cavity.

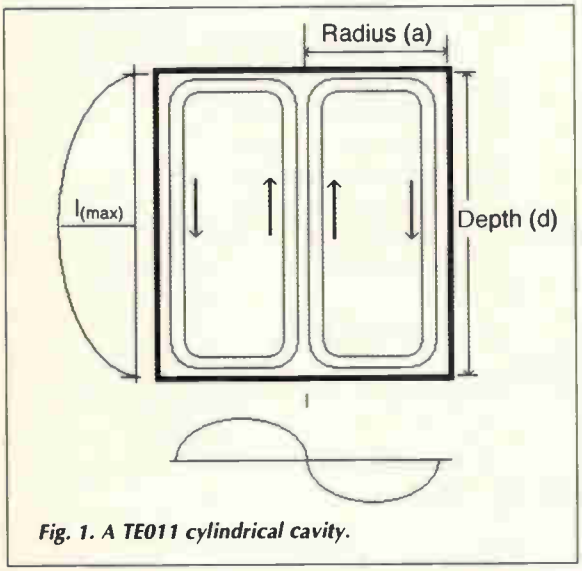


Fig. 1. A TE011 cylindrical cavity.

Table 1: Cavity Radius – v – Resonant Frequency

Radius (mm)	Frequency (MHz)
21.67	9116.166
21.68	9111.961
21.69	9107.761
21.7	9103.563
21.71	9099.369
21.72	9095.180
21.73	9090.994

Confusingly, convention dictates that the bore is regarded as radius, symbol 'a', and the length is depth, symbol 'd'. The propagation constant 'p' has different specific values dependent on the mode of the cavity, for TE011 the value is 3.832. In this mode there is a peak in the Q factor value when the diameter and depth are equal.

$$f = \frac{c}{2\pi}\sqrt{\left(\frac{p}{a}\right)^2 + \left(\frac{\lambda}{d}\right)^2}$$

f = resonant frequency, c = speed of light, d = depth, a = radius. p = propagation constant.

The precision required in machining to produce a cylindrical cavity of the required resonant frequency is shown in **Table 1.** An error of 0.01mm (10 microns) results in a resonant frequency error of 4MHz for a cavity nominally resonant at 9.1GHz.

Machining to this tolerance is expensive and difficult. In reality it should not be necessary to require the resonant frequency to be defined to this level of accuracy. The diameter can be machined to a far more relaxed tolerance and the error compensated by modifying the depth. It is essential to know the penalties of such a solution particularly where mechanical adjustment is introduced into the design.

The currents in the walls of a TE011 cavity are minimal at the junction of the end caps and the barrel, these are voltage maximum regions and as such do not demand high electrical conductivity. A cavity was made in three parts as shown in **Fig. 2,** the barrel machined to as close a tolerance as was reasonably practicable and the two end caps were machined to be a close fit to the bore but initially somewhat over length. The flat face was machined to provide a wall thickness of <1 mm for a nominal 5mm aperture, eventually a wave-guide flange would be fitted to this surface.

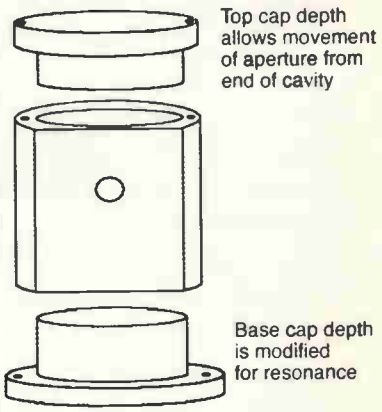


Fig. 2. A development cylindrical cavity.

By modifying the penetration of the top cap the position of the aperture could be moved relative to the end of the cavity. Similarly the base cap could be machined to vary the cavity depth.

Determining the Cavity Resonance

A classical method described by Montgomery, **Fig. 3,** was used to determine the resonant frequency of the cavity and obtain some indication as to its Q value. For a TE011 mode cavity the Q peaks when the diameter and depth are equal. The cavity under

test is mounted onto the broad face of a length of wave-guide and coupling is through a small aperture. At the resonant frequency power is absorbed by the cavity and this is indicated by a change in the level of power reaching the dummy load.

In order to maximise the reactance tuning it is necessary for the flat face on the cavity to become the internal face of the wave-guide broad wall and for the two items to be electrically bonded. If these criteria are met, the attenuation approaches 100% and no power reaches the dummy load. The signal generator could be presented with a short circuit load under these conditions and the Variable Attenuator provides some degree of isolation.

A far more practical solution is to place the cavity flat face on the outside of the wave-guide and to hold it in place using a sprung clamp assembly (**Fig. 4**). This requires an aperture in the cavity and another aperture in the centre of the wave-guide. It is convenient to use round apertures in both the wave-guide and cavity, however during testing it was found desirable to utilise a slot aperture to define TM or TE mode within the cavity. With a similar slot machined in the wave-guide there were a number of spurious responses as detailed in **Table 2.**

This configuration requires a number of specific components that might not be available across a number of wave-guide sizes or frequency ranges.

Alternative Measuring System

By measuring the amplitude of the signal within the cavity rather than measuring the power reaching the dummy load it is possible to eliminate the directional coupler and power meter. Using very small coupling between the wave-guide and the cavity the signal generator essentially sees only the dummy load termination. Whilst this system, as shown in **Fig. 5.** does not lend itself to absolute measurements it is of sufficient accuracy and repeatability to render it suitable for comparative evaluation of cavities.

The critical factor is that the coupling between the cavity and the wave-guide is very small, this results in a correspondingly small signal from the detector and DC measurements are somewhat difficult even using a high resolution digital voltmeter. The HP8673 Signal Generator will deliver +10dBm and detector voltages of 10mV are typical.

An option is to amplitude modulate the signal generator and to measure the audio frequency modulation of the detected signal. This signal also contains wideband noise and 50Hz and requires filtering. The design of a system built around MF8CCN active filters with the modulation frequency and band-pass filter centre frequency locked together is shown in **Fig. 6.**

Detector Diodes

The specific microwave detector used for this work was obtained from surplus as a complete assembly with an N type input connector and a BNC output connector. Such assemblies are not difficult to acquire. **Figure 7** shows the generic form of the design for a detector and suitable diodes are manufactured by Philips.

Pick-up probe presence inside the cavity causes perturbation, the resonant frequency will be lowered as a consequence and can be approximated as equal to the change in cavity volume. Provided the probe is of a small diameter and its projection is minimal, 1mm diameter extending 2mm into the cavity, its presence in a 9GHz cavity is virtually invisible.

The position of the probe is not vitally important. For convenience it

Table 2: Spurious responses and aperture-to-aperture geometry.

Slot / Slot	Slot / Round	Round / Round
14,506	*	*
14,206	*	*
13,944	13,943	13,943
13,460	*	*
13,075	*	*

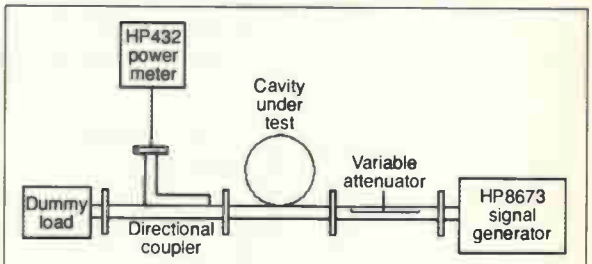


Fig. 3. Montgomery reactance resonance method.

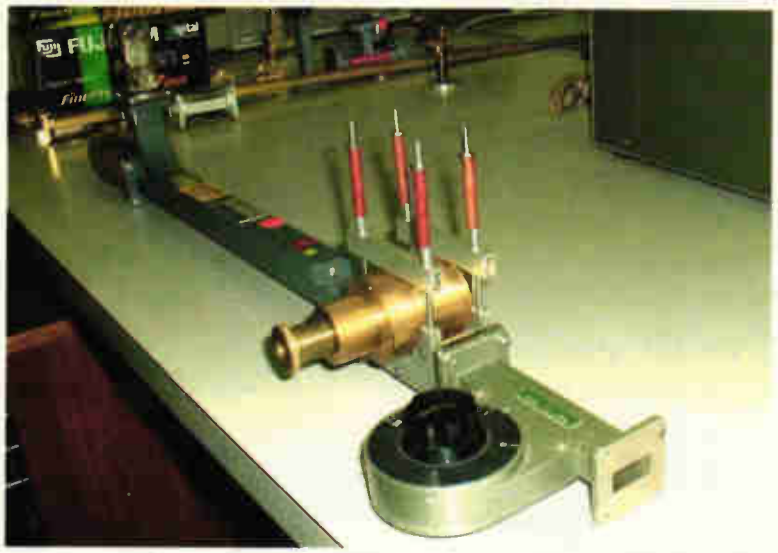


Fig. 4. Practical realisation of the reactance resonance technique.

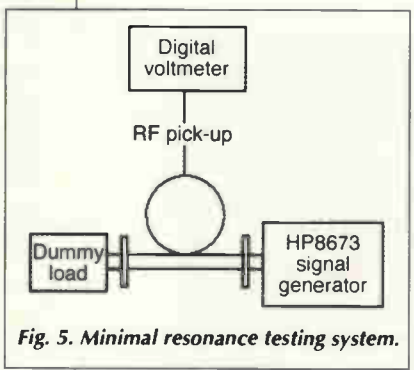


Fig. 5. Minimal resonance testing system.

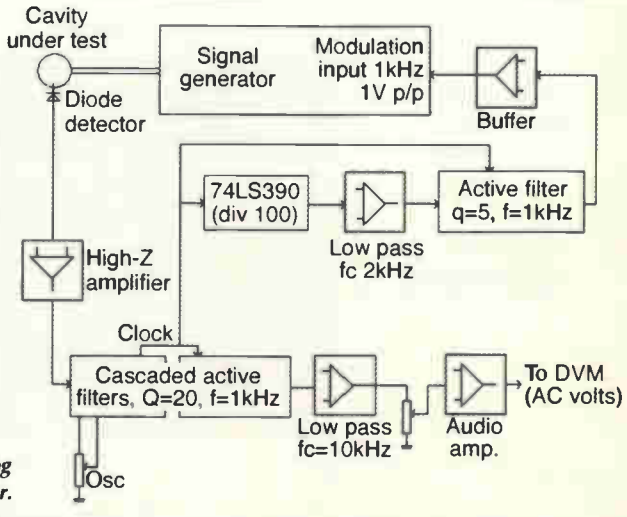


Fig. 6. Active filter tracking modulator/demodulator.

was mounted on the barrel rather than an end wall because these were subject to changes during development. A suitable point is at a quarter of the depth of the cavity and the probe access aperture diameter should be minimal.

Wave-guide dummy loads

These items are not plentiful and tend to be expensive, attempts at making wave-guide dummy loads proved most disappointing. In an attempt to use a commercial load on different, though similar, wave-guides a tapered transition with an overall length of several wavelengths was constructed for a nominal frequency of 14,000MHz. The resulting standing wave ratio was found to be rather uneven and not acceptable.

An alternative is to use a wave-guide/coaxial transition and coaxial attenuators open circuited at the far end. Attenuators tend to be plentiful and inexpensive and a 40dB unit has a return loss of 80dB, effectively making the open circuit invisible to the wave-guide. It is important to use attenuators designed for the frequency range in use.

The SWR plot shown in **Fig. 8.** demonstrates that this technique provides a quite acceptable termination to the wave-guide for this application. The adaptor was a Flann 17094-NF10 (s/n 109903), Series 1 plot is with an HP 8491A Attenuator (s/n 10681), calibration marked 19.8dB/14GHz, whilst the Series 2 plot using a Marconi 8534/4 Attenuator (s/n 3080), calibration marked 20dB/12GHz, shows a slightly higher SWR at this frequency range.

There were occasions when it was necessary to recheck a cavity resonance or relative Q after a wave-guide flange had been fitted to the flat face. The void between the wave-guide and the cavity is approximately 6mm and represents a significant spacing at 10GHz. Where this is unavoidable a solid filler with an identical aperture should be fitted into the flange void. This does not provide perfect results but it does suppress irregularities due to the flange.

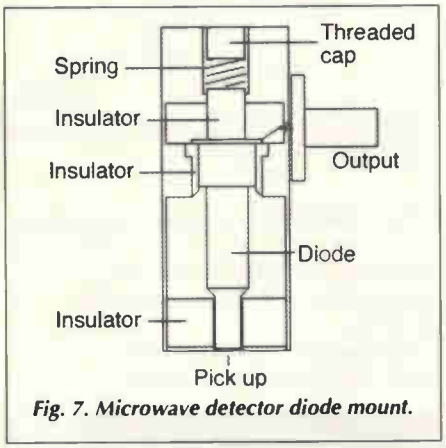


Fig. 7. Microwave detector diode mount.

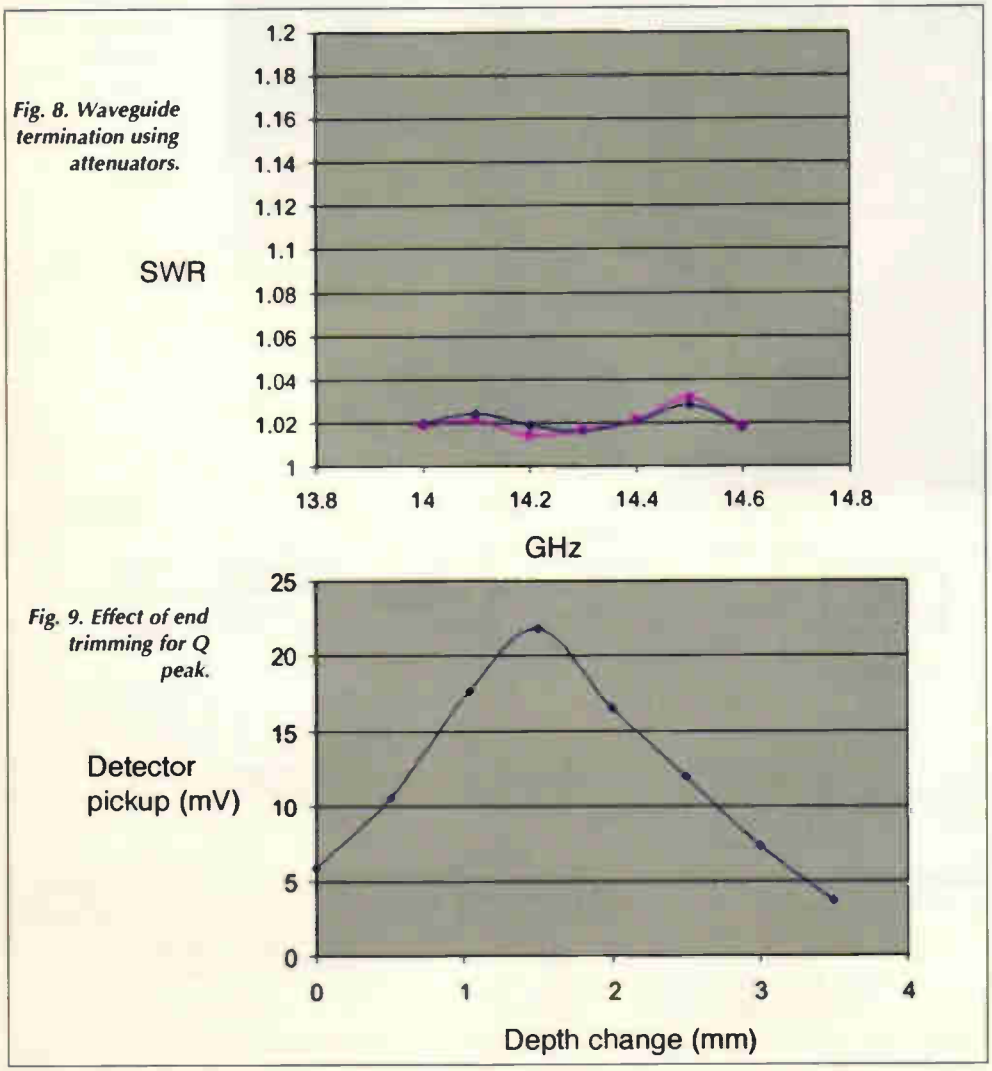


Fig. 8. Waveguide termination using attenuators.

Fig. 9. Effect of end trimming for Q peak.

Optimising the Cavity

Accepting that the diameter had been machined as closely as possible to the desired dimension, it remained necessary to optimise the depth. This was determined by measuring the value of detected signal for various depths of penetration of the base end cap. This component had been initially made over length and the face was skimmed in a lathe.

As shown in **Fig. 9.** the signal peaks quite distinctly when the depth is adjusted in 0.5mm steps with final trimming to an optimum value requiring 10 micron precision.

This particular cavity was resonant at 9,109.25MHz, each millimetre of depth trimming represented a frequency shift of 30MHz.

Having optimised the cavity to meet the 2a=d criterion it was possible to determine a relative value for the Q by taking measurements at 50kHz intervals and estimating the bandwidth at the –3dB points.

From the plot (**Fig. 10.**) the half voltage bandwidth of this cavity is c. 600kHz, the resonant frequency 9,109.250MHz and the unloaded Q of 15,182. This is a realistic value but it must be emphasised that it is not an absolute value and is used as a datum for comparison.

Customising the Cavity

In a microwave demagnetiser the rock sample is introduced into the cavity via an access hole in the bottom end cap, it is positioned at the centre of the cavity for processing and then exits via a further hole in the top end cap for measurement. The samples have magnetic, conductive and dielectric properties that cause perturbation within the cavity causing the resonant frequency to alter. It was important to evaluate the effects of the access and exit holes and the means of compensating for the presence of the samples within the cavity.

For our specific application it was necessary for a rock sample to enter

and exit the cavity though a 6mm diameter aperture in the end plate as shown in **Fig. 11.** A feature of the 'TE011' cavity is that at the centre of the end plates there is a current minimum, removing material from this area should not have a significant effect on the Q of the cavity.

The effect on the Q factor was determined by first measuring the cavity without the aperture and then with the aperture, the depth of insertion of the end plate into the cavity was re-optimised for peak Q.

The initial Q was 15,180, with a 6mm aperture 14,198, representing a 6% loss.

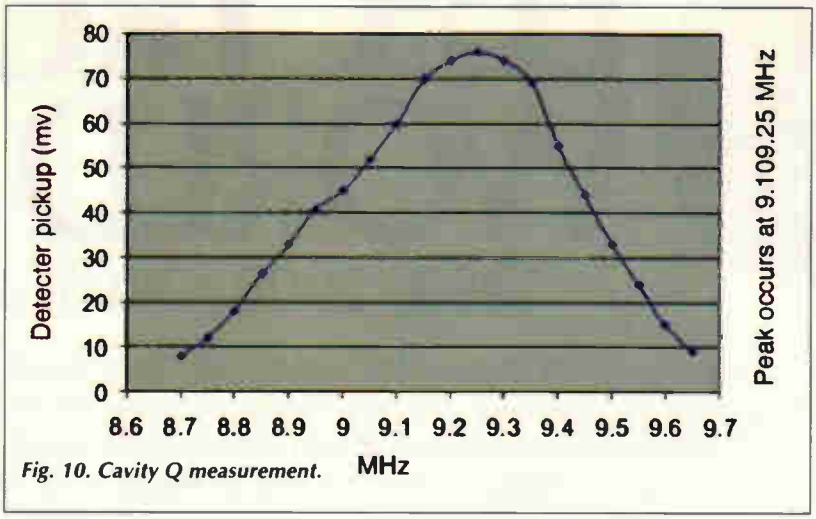


Fig. 10. Cavity Q measurement.

End Plate Adjusters

In order to retune the cavity due to the presence of a sample it was necessary to vary the depth. An adjuster plate of diameter very close to that of the cavity was mounted onto a fine threaded lead screw that had a diameter 75% that of the cavity (**Fig. 12**). This adjuster can be seen in the photograph of the test jig shown in Fig. 4.

The initial Q was 13,100, with a 6mm aperture 12,435. an 18% loss. It was obvious that there was leakage around the adjuster plate as a number of small spurious responses were noticeable. These could be suppressed by a resonant groove around the circumference of the adjuster plate and the back volume filled with a conducting material.

To confirm that there was a problem with the back volume, a similar plate was mounted on to a much smaller lead screw (**Fig. 13.**) and this provided a significant back volume.

As had been anticipated the Q plunged to 11,770 and, with the aperture, 11,064 representing a loss of 25%. Spurious responses were very evident. This experiment provided the opportunity to investigate filling the back volume although with little success.

On the basis that the change in frequency due to the sample was relatively small it is possible to arrange for a small change in depth of the cavity without introducing the plate and the problem of back volume by simply making a small adjustment to the cavity volume. This is exploiting the perturbation and is shown in **Fig. 14.**

The end cap was machined to create a cavity depth 1mm greater than the optimum value, the adjuster was a lead screw with a diameter half that of the cavity and protruded 1 or 2mm into the cavity. As expected a Q of 14,806 was close to that obtained with the plain end cap. From earlier experiences it was predicted that aperture would reduce the Q by 6% and in fact the resultant value of 13,917 was very close indeed to prediction. This design is by far the most superior means of adjusting the cavity for small frequency changes.

Conclusions

From an initial outset of having absolutely no means of determining any of the parameters regarding cylindrical cavities, it has been shown that the development of suitable instrumentation is entirely possible on a minimal budget. No claim has been made for absolute accuracy, neither is it necessary to produce absolute values when the objectives can be evaluated by comparison as has been demonstrated.

In our application, resonance at a specific frequency was not a criteria because of the availability of bandwidth in the microwave amplifier. It has been shown that an optimised cavity can be fabricated by machining the diameter without recourse to extremes of tolerance and then modifying the depth to arrive at a resonant frequency sufficiently close to the target value for our purposes.

In our specific application the requirement for sample access and exit apertures are unavoidable and it has been possible to determine that such apertures do not have a particularly detrimental effect. Adjusters need to be considered with care if they are not to compromise the cavity. The minimisation, indeed the elimination, of volume behind an adjuster plate should be a point of serious attention if spurious cavity responses are to be avoided.

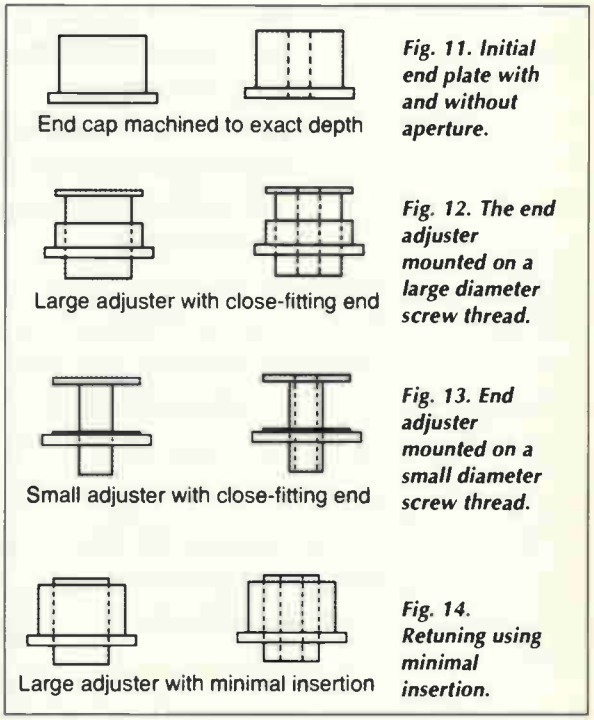


Fig. 11. Initial end plate with and without aperture.

Fig. 12. The end adjuster mounted on a large diameter screw thread.

Fig. 13. End adjuster mounted on a small diameter screw thread.

Fig. 14. Retuning using minimal insertion.

Appendix.

This listing gives sources of information, instrumentation and microwave hardware used during our development of microwave cavities. It is intended for guidance.

Barlow, 'Microwaves and Wave Guides', Constable and Company Ltd, 1947 (out of print)

Flann Microwave Ltd, Dunmere Road, Bodmin, Cornwall, PL31 2QL. http://www.flann.com. Microwave hardware, couplers, flanges, attenuators, Wave Guide.

Huxley, L.G.H., 'A Survey of The Principals & Practice of Wave Guides', University Press, Cambridge, 1947 (out of print).

Johns Radio, Whitehall Works, 84 Whitehall Rd East, Birkenshaw, Bradford, BD11 2ER., surplus hardware and instrumentation.

Montgomery, 'Techniques of Microwave Measurements', McGraw-Hill, 1947 (out of print)

Pozar, D.M., 'Microwave Engineering (2nd edition)', 1998, ISBN 0-471-17096-8, John Wiley & Sons, Inc. (http://www.wiley.com/college) (currently in print).

Rollet & Co. Ltd, Unit 10, Cotswold Mews, 30 Battersey Square, London, SW11 3RA. Solid and flexible Wave Guide, Flanges.

Farnell Elect Components, stockist of Philips Microwave Detector Diodes.

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8-bit microcontroller with 16kbyte flash

Toshiba has introduced a general-purpose 8-bit microcontroller with 16kbyte of on-chip NAND-type flash memory. Supplied in a 44-pin LQFP, the TMP86FH47UG microcontroller has an input voltage range of 2.7 to 5.5V and is based on the company's 8-bit TLCS-870/C CPU core. This core has 731 basic instructions and operates at up to 16MHz. The device also has 512byte of RAM. The microcontroller incorporates on-board peripherals including LED driver output pins, 10-bit timer-counter, two-channel 16-bit timer-counter and programmable watchdog timer. Serial connectivity comprises a single-channel UART and a high-speed 8-bit serial interface.

There is dual-clock operation and, in addition to the built-in stop mode, it has eight power-down modes. These comprise slow 1 and 2, idle 0, 1 and 2, and sleep 0, 1 and 2.

Toshiba
www.toshiba-europe.com
Tel: +44(0) 49 211 5296 254

Photovoltaic connector system

Tyco Electronics has introduced the second generation of Solarlok connector systems, a flexible system solution for interconnection between photovoltaic modules and the DC/AC converter. The junction box concept is based on a

flexible open system structure, which allows serial as well as parallel interconnection via direct wire connection or separable connectors. Within the junction box, the number of termination positions is flexible and can be tailored to specific customer requirements. The robust construction combined with the proven 'squeeze to release' connection system ensures reliable connection in harsh outdoor environments with extreme temperature variations.

Tyco Electronics
www.tycoelectronics.com
Tel: +44(0) 49 6232 30214

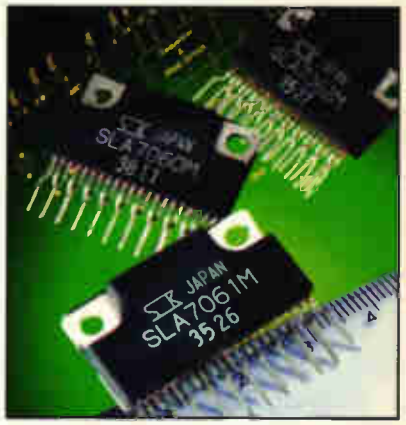
Hardware probe for embedded Linux development

Wind River Systems has added Linux support to its hardware bring-up tools. Fully able to run with either VxWorks, embedded Linux, the vision PROBE II hardware is available for the IBM PowerPC 405 family. The intention is to support Linux developers by providing visibility and access to the virtual and physical address space. It allows developers to track Linux virtual access and to control their physical peripherals and memory.

Wind River
www.windriver.com
Tel: +44(0) 121 3590999

Two-phase unipolar microstepping motor

Allegro Microsystems has a family of unipolar microstepping two-phase constant-current motor driver ICs with built-in sequencing



capabilities. The devices provide a unipolar microstepping motor driver, including a built-in translator to operate unipolar stepper motors with a simple step input. There is no need for phase-sequence tables, high-frequency control lines, or complex interfaces to program. Combining low-power CMOS logic with high-current, high-voltage power FET outputs, the series SLA706xM translator/drivers provide complete control and drive for a two-phase unipolar stepper

Smallest GPS surface mount antenna

Sarantel has announced a surface mount version of its GeoHelix GPS antennas, which only requires 10x14mm of board space. The balanced antenna requires no ground-plane and according to the supplier offers a wider beamwidth than conventional patch antennas, with a 3dB contour up to 120° being typical. A consequence of the wider beamwidth is that more GPS satellites can be seen simultaneously, resulting in greater accuracy and positioning reliability. In addition, the antenna's small near-field means the weak signals from the satellites are not de-tuned when the handset is close to human tissue such as the hand or head. The GeHelix-SMP antenna, which operates on the GPS L1-band (1575.42MHz) measures only 35x14.0x11.2mm (uncapped) and uses a patented design in which copper tracks deposited into a small ceramic cylinder are individually and automatically laser trimmed for optimum frequency response.

Sarantel
www.sarentel.com

Please quote *Electronics World* when seeking further information

motor with internal fixed off-time and pulse-width modulation (PWM) control of the output current.

Allegro Microsystems
www.allegromicro.com

In-line EMC compliance measurements

The LISN1600 from TTI is a line impedance stabilisation network which provides compliance level measurements of conducted EMC emissions at the supply input of any piece of electrical or electronic equipment operating from a single-phase AC supply. The instrument is designed for use in conjunction with a spectrum analyser or a measurement receiver. The instrument fully meets the requirements of CISPR 16 for measurements in both band A (10kHz to 150kHz) and band B (150kHz to 30MHz). A switchable 150kHz filter is incorporated to limit low-frequency signals (particularly AC line frequency) when measuring in band B. This reduces the dynamic range requirements of the measuring device.

Thurlby Thandar Instruments
www.tti-test.com
Tel: +44(0) 1480 412451

Touch screen monitor responds to gloves

A 3M touch monitor from Secure Retail has the touch technology designed and manufactured as a part of the monitor itself to accredited ISO9001 procedures rather than retro-fitted to a standard monitor. The monitor is available with a choice of durable capacitive or more sensitive resistive touch screen technology. There is a resistive 5-wire version which will respond to the touch of fingernails, credit cards, gloved hands and fingers.

Secure Retail
www.tsecure-retail.co.uk

38mm DC fans outstrip AC versions

The ADDA range of 80mm DC fans, available in 38mm thickness as well as 15, 20 and 25mm, are intended as an upgrade for the 80x80x38mm AC fans currently being used in many designs. A DC fan of this size can operate at 4,000 rev/min and is capable of airflow of over 29 litre/sec (61.7CFM). This is compared to 11.3 litre/sec (24CFM) delivered from a similarly sized AC fan that has its speed limited by the AC frequency to typically 2,750 rev/min. There is also speed control by varying the input voltage making them suitable for use in intelligent cooling systems that are required to cool complex electronic systems.

Aerco
www.aerco.co.uk
Tel: +44(0) 1403 260 206

Low closure force shielding gasket

Chomerics is offering a low closure force gasket in over 90 different profiles, which is suitable for door seal, faceplate and backplane shielding applications in products such as

CD head units, navigation systems and entertainment systems. Called Soft-Shield 3500, it consists of nickel-plated nylon wrapped around a urethane foam core and has a shielding effectiveness of greater than 90dB between 50MHz and 10GHz. The construction gives a closure force of typically less than 1lb/in (0.175N/mm), which the firm says allows the gasket to be used with thin walled plastic enclosures. The extensive range of profiles that includes rectangular, C-fold, D-shape and P-shapes. It is available in standard or cut lengths or kiss-cut to custom shapes. The operating temperature range is –40 to +70°C.

Chomerics
www.chomerics.com

High performance FM radio data link

RF Solutions has introduced an 868MHz version of its RTFQ1 and RRFQ1 transmitter and receiver pair. Designed to provide FM radio data link in applications such as wireless security systems, car alarms, remote and data capture. Suitable for both one-to-one and multi-node applications, RTFQ1 and RRFQ1 are already available in 433.92MHz and 315MHz versions. Key features include a data rate of up to 9.67kbit/s and a range of approximately 75 metres indoors and 250 metres across open ground. The DIL packaged transmitter measures 20.3x11.4mm and operates from a 3V supply. The receiver is housed in a 38.3x18.3mm SIL package and requires a 5V supply in operating mode. RTFQ1 and RRFQ1 modules feature a laser trimmed thick-film ceramic hybrid construction that results in extremely stable performance across an industrial operating temperature range of –20 to +85°C.

RF Solutions
www.rfsolutions.co.uk
Tel: +44(0) 1273 488880

Ready-to-use IP66 enclosures

Rolec is offering custom enclosures fully finished and ready for final assembly of the customer's components. In addition, the supplier also offers a design modification service where it can evaluate and implement technical modifications to both electrical and electronic assemblies. The firm manufactures a range of IP66 enclosures in diecast aluminium, polyester,

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polycarbonate and ABS. CNC milling, drilling and tapping can allow for the mounting of cable glands, connectors and switches into the case walls. Special performance EMC and IP gaskets can be added, as can hinges, DIN rails and earth points. Further customisation can machine large apertures for LCD modules into the lid of the enclosure. In order to customise the look and feel of the enclosures, special materials, silk-screen printing and non-standard colours are available on demand.

Rolec
www.rolec-enclosures.co.uk
Tel: +44(0) 1489 583858

Temperature-compensated transistors for audio

The Sanken SAP 16 range of 150W bipolar temperature-compensated Darlington transistors with built-in emitter resistors, is available from Allegro MicroSystems Europe. For audio applications, the devices use a proprietary technology to offer temperature compensation using an on-chip diode. Both n- and p-type transistors are available. In addition to the collector power dissipation of 150W, absolute maximum ratings include collector-base and collector-emitter voltages of 150V, collector current of 15A, base current of 1A, and diode forward current of 10mA. Other key electrical parameters include collector and emitter cut-off current of 100μA maximum, collector-emitter voltage of 160V minimum, DC current transfer ratio from 5000 to 20,000, collector-emitter saturation voltage of 2V maximum, and base-emitter saturation voltage of 2.5V maximum. The SAP 16 series is available in conventional TO-220 and TO-3P power packages as well as single-inline configurations.

Allegro Microsystems
www.rallegromicro.com
Tel: +44(0) 1799 520022

Handheld spectrum analyser

Anritsu's latest handheld spectrum analyser has improved sweep speed, and includes a built-in pre-amplifier that increases the analyser's sensitivity and dynamic range. The MS2711D, which covers a specified range of 100kHz to 3.0GHz, has a –135dBm noise floor. Full span sweeps from 9kHz to 3GHz can be performed in less than 1.1s, while the sweep speed in zero span can be set from less than 50μs up to 20s. The analyser will tolerate input signals up to +43dBm without any damage. There is also a dynamic attenuation feature, which tracks the input signal level and automatically adjusts the input attenuation level to protect the MS2711D in the presence of high signal levels.

Anritsu
www.eu.anritsu.com
Tel: +44(0) 1582 433433

Radio receiver for wireless RS-232

RF Solutions has introduced a radio transceiver in an 18-pin DIP/SOIC package for use in lower cost wireless RS-232, cable replacement, alarms and communications systems. Capable of standalone operation and with a direct microcontroller interface, the RF600T requires a 3.0V to 5.5V supply (2.0V to 5.5V optional). It features a 190byte buffer that is used to either hold host data prior to

transmission, or buffer data received over the radio link before it is transmitted to the host. Control lines are used to handle the flow of data to and from the host. Data packet generation is automatically performed along with 'Manchester' encoding and CRC-based error checking. Further hardware features include two digital telemetry lines and an asynchronous serial host interface.

RF Solutions
www.rfsolutions.co.uk
Tel: +44(0) 1273 898000

Windows XP enters test department

Agilent's line of Infiniium oscilloscopes will run Windows XP. The company has also added three new application programs for the upgraded scopes. If you own an Infiniium scope with Windows XP or you upgrade your 54830-series scope from Windows 98 to Windows XP, you can then use the three new software packages. Called "My Infiniium", the software lets you perform data analysis on the scope rather than forcing you to export data to a PC. You can now make direct data links to programs such as Excel or Matlab because of the open Windows platform. My Infiniium also adds intensity-grade variable persistence amid lowpass/highpass digital filters. A new button, called "Quick Execute" (new scopes only) lets you run custom programs that you can write in any Windows-based language. You can also use IVI instrument drivers for automated tests. A "save all waveforms" feature lets you save current scope data in ASCII format so you can import data into any data-analysis software package. With an operating system version 3.1, you can connect an external monitor to the scope and run the scope in dual-monitor mode.

Agilent
www.agilent.com
Tel: +44(0) 1344 366666

32-bit ARM MCU in 64-pin package

Philips Electronics has added two microcontrollers to its range of ARM based microcontrollers. The LPC2114 and LPC2124 32-bit MCUs feature up to 256kbytes of embedded flash memory, 10-bit A/D converters, 16k of SRAM, pulse width modulation (PWM), timers, UARTS, serial peripheral interface (SPI) and up to 46 general purpose I/O in a small outline 64-pin package. Anticipated applications include motor control, servo loop control, power management and data acquisition. The LPC21 xx family uses a 0.18μm CMOS

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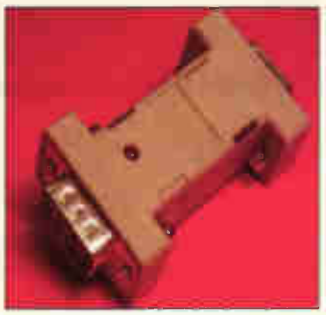

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embedded Flash process enabling ultra-low 1.8V voltage operation.
Philips
www.semiconductors.philips.com

DC-DC boost converters

Texas Instruments has a range of synchronous 1.5A switch DC-DC boost converters with linear regulation functionality for battery-powered electronics. Offering up to 96 per cent power conversion efficiency from an input range of 0.9V to 5.5V, the TPS610x family provide 200mA from a 0.9V input when using a single-cell alkaline battery. The TPS6102x can also be used for generating 5V at 500mA from a 3.3V rail or a LI-ion battery at up to 96 per cent power conversion efficiency. The boost converters generate a stable output voltage that can be adjusted by an external resistor divider or fixed internally on the chip. It uses a synchronous rectifier based on a pulsewidth modulation (PWM) controller and according to the supplier, it can operate in a down-mode with linear regulation if the input voltage reaches or exceeds the output voltage. In this mode the control circuit changes the behaviour of the rectifying integrated switching transistor. The converter then operates as a linear regulator by setting a voltage drop across the transistor high enough to regulate the output voltage. As a result, the output voltage remains regulated even if the input voltage exceeds the nominal output voltage. For dual output applications, the TPS611xx family of DC-DC boost converters with integrated LDO features two independent voltage regulators in a small 4x4mm QFN package.
Texas Instruments
www.ti.com

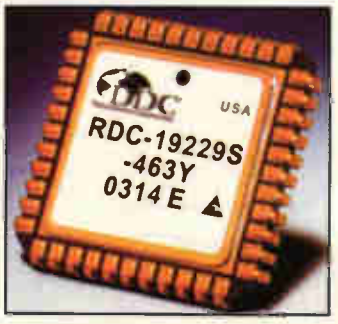
Mosfets and bipolar transistors

Rohm is offering Mosfets and bipolar junction transistors (BJTs) in TSMT packages that measure 2.9x1.6x1.0mm. Available in N-channel and P-channel formats, Rohm's 20V and 30V TSMT Mosfets have current ratings up to 4.5A and power ratings up to 1.25W.

Typical on-resistance values are down to 31mΩ at a V_{GS} of 4V (dependent on device chosen). The TUMT-packaged Mosfets have drain currents up to 3A, while on-resistance values are 50mΩ at a V_{GS} of 4V. Power dissipation for TUMT devices is rated at 0.8W. Both series include single- and dual- Mosfet options, each of which can be supplied with or without a built-in Schottky diode.
Rohm
www.rohm.co.uk

Class K Resolver-to-digital converter

Data Device Corporation has released a class K resolver-to-digital monolithic converter, the RDC-19229S-4XX series. The converter accepts analogue inputs from electromechanical angular transducers, which provide a digital output of up to 16-bit resolution. The converter provides features such as programmable 10, 12, 14, and 16-bit resolution, programmable bandwidth and tracking rates, +5V only input power, and internal synthesised reference.



Typical space applications include motor control, positioning and reaction wheels.
Data Device Corporation
www.ddc-web.com
Tel: +44(0) 1635 811140

Programmable micrometer for cables

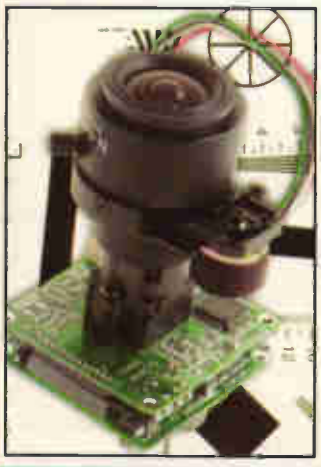
Precision instrument specialist Tinsley has introduced a programmable micrometer intended for use in cable resistance measurement, contact resistance measurement of connectors, switches, relays and similar components in accordance with standards. Operating from mains supply or optional rechargeable batteries the unit provides 4-wire resistance measurement from 0.1μΩ to 200kΩ with an accuracy of 0.03 per cent and resolution to 0.1μΩ.
Tinsley
www.tinsley.co.uk

Waterproof USB connectors

Bulgin's latest range of IP68 waterproof USB connectors are four-pole hot pluggable connectors supporting USB version 2 and data rates up to 480Mbit/s. Housed in an over-mouled body using UL94V-0 rated PVC, the connectors are environmentally sealed to IP68. Three formats of single- and double-ended cables offer a number of options for environmentally sealed connections between PC and

High resolution colour PCB camera module

A colour PCB camera module for CCTV applications is based around a 1/3in. CCD and offers an aperture of F1.5. Available from Stortech Electronics, the CB60HVAI is supplied with an integrated, variofocal lens, with a focal length betwwen 4mm and 9mm. Incorporating a DC controlled auto-iris; the device has a horizontal picture resolution of 450TVL and is designed to deliver images over a wide range of light levels, down to a minimum illumination of 1 Lux (at scene), Featuring electronic auto exposure settings of between 1/50th and 1/80,000th of a second, and a signal to noise ratio of greater than 45dB, the camera requires a regulated 12V DC power supply. The video output is a standard PAL composite video signal at 1V peak to peak (75 ohm load). A standard audio line-level signal is also provided at 1V peak-to-peak.



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Motor Drivers/Controllers

Here are just a few of our controller and driver modules for AC, DC, unipolar/bipolar stepper motors and servo motors. See website for full details.

DC Motor Speed Controller (5A/100V)
Control the speed of almost any common DC motor rated up to 100V/5A. Pulse width modulation output for maximum motor torque at all speeds. Supply: 5-15VDC. Box supplied. Dimensions (mm): 60Wx100Lx60H.
Kit Order Code: 3067KT - **£12.95**
Assembled Order Code: AS3067 - **£19.95**

NEW! PC / Standalone Unipolar Stepper Motor Driver

Drives any 5, 6 or 8-lead unipolar stepper motor rated up to 6 Amps max. Provides speed and direction control. Operates in stand-alone or PC-controlled mode. Up to six 3179 driver boards can be connected to a single parallel port. Supply: 9V DC. PCB: 80x50mm.
Kit Order Code: 3179KT - **£9.95**
Assembled Order Code: AS3179 - **£16.95**

PC Controlled Dual Stepper Motor Driver

Independently control two unipolar stepper motors (each rated up to 3 Amps max.) using PC parallel port and software interface provided. Four digital inputs available for monitoring external switches and other inputs. Software provides three run modes and will half-step, single-step or manual-step motors. Complete unit neatly housed in an extended D-shell case. All components, case, documentation and software are supplied (stepper motors are NOT provided). Dimensions (mm): 55Wx70Lx15H.
Kit Order Code: 3113KT - **£15.95**
Assembled Order Code: AS3113 - **£24.95**

NEW! Bi-Polar Stepper Motor Driver

Drive any bi-polar stepper motor using externally supplied 5V levels for stepping and direction control. These usually come from software running on a computer.
Supply: 8-30V DC. PCB: 75x85mm.
Kit Order Code: 3158KT - **£12.95**
Assembled Order Code: AS3158 - **£26.95**

Most items are available in kit form (KT suffix) or assembled and ready for use (AS prefix).

Controllers & Loggers

Here are just a few of the controller and data acquisition and control units we have. See website for full details. Suitable PSU for all units: Order Code PSU203 £9.95

Rolling Code 4-Channel UHF Remote

State-of-the-Art. High security. 4 channels. Momentary or latching relay output. Range up to 40m. Up to 15 Tx's can be learnt by one Rx (kit includes one Tx but more available separately). 4 indicator LED 's. Rx: PCB 77x85mm, 12VDC/6mA (standby). *Two and Ten channel versions also available.*
Kit Order Code: 3180KT - **£41.95**
Assembled Order Code: AS3180 - **£49.95**

Computer Temperature Data Logger

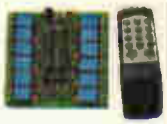
4-channel temperature logger for serial port. °C or °F. Continuously logs up to 4 separate sensors located 200m+ from board. Wide range of free software applications for storing/using data. PCB just 38x38mm. Powered by PC. Includes one DS1820 sensor and four header cables.
Kit Order Code: 3145KT - **£22.95**
Assembled Order Code: AS3145 - **£29.95**
Additional DS1820 Sensors - **£3.95 each**

NEW! DTMF Telephone Relay Switcher

Call your phone number using a DTMF phone from anywhere in the world and remotely turn on/off any of the 4 relays as desired.
User settable Security Password, Anti-Tamper, Rings to Answer, Auto Hang-up and Lockout. Includes plastic case. 130x110x30mm. Power: 12VDC.
Kit Order Code: 3140KT - **£39.95**
Assembled Order Code: AS3140 - **£59.95**

Serial Isolated I/O Module

PC controlled 8-Relay Board. 115/250V relay outputs and 4 isolated digital inputs. Useful in a variety of control and sensing applications. Uses PC serial port for programming (using our new Windows interface or batch files). Once programmed unit can operate without PC. Includes plastic case 130x100x30mm. Power: 12VDC/500mA.
Kit Order Code: 3108KT - **£54.95**
Assembled Order Code: AS3108 - **£64.95**

Infrared RC Relay Board

Individually control 12 on-board relays with included infrared remote control unit. Toggle or momentary. 15m+ range. 112x122mm. Supply: 12VDC/0.5A
Kit Order Code: 3142KT - **£41.95**
Assembled Order Code: AS3142 - **£59.95**

PIC & ATMEL Programmers

We have a wide range of low cost PIC and ATMEL Programmers. Complete range and documentation available from our web site.

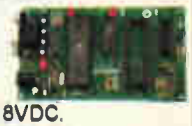
Programmer Accessories:
40-pin Wide ZIF socket (ZIF40W) £15.00
18V DC Power supply (PSU201) £5.95
Leads: Parallel (LEAD108) £4.95 / Serial (LEAD7B) £4.95 / USB (LEADUAA) £4.95

NEW! USB 'All-Flash' PIC Programmer

USB PIC programmer for all 'Flash' devices. No external power supply making it truly portable. Supplied complete with 40-pin wide-slot ZIF socket, box and Windows Software.
Kit Order Code: 3128KT - **£49.95**
Assembled Order Code: AS3128 - **£64.95**

Enhanced "PICALL" ISP PIC Programmer

Will program virtually ALL 8 to 40 pin PICs plus a range of ATMEL AVR, SCENIX SX and EEPROM 24C devices. Also supports In System Programming (ISP) for PIC and ATMEL AVRs. Free software. Blank chip auto detect for super fast bulk programming. Requires a 40-pin wide ZIF socket (not included).
Kit Order Code: 3144KT - **£54.95**
Assembled Order Code: AS3144 - **£59.95**

ATMEL 89xxxx Programmer

Uses serial port and any standard terminal comms program. 4 LED's display the status. ZIF sockets not included. Supply: 16-18VDC.
Kit Order Code: 3123KT - **£29.95**
Assembled Order Code: AS3123 - **£34.95**

NEW! USB & Serial Port PIC Programmer

USB/Serial connection. Ideal for field use. Header cable for ICSP. Free Windows software. See website for PICs supported. ZIF socket not incl. Supply: 18VDC.
Kit Order Code: 3149KT - **£29.95**
Assembled Order Code: AS3149 - **£44.95**

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peripherals. these are: IP68 sealed A type USB to standard B type USB. IP68 sealed B type USB to standard A type USB. and IP68 sealed A type USB to IP68 sealed B type. The mating panel mount connectors have either A or B type USB connection to the front of panel.

Bulgin
www.bulgin.co.uk
Tel: +44(0) 20 8594 5588

Surface mount device counter

DiagnoSYS has introduced a surface mount parts counter capable of counting all surface mount devices in embossed or punched carrier tape whether in clear or black material. Called the PC250+. the unit also allows for bi-directional part counting. subtracting when the direction of tape travel is reversed. Designed for counting SMD parts in tape, safely and quickly. The tape is handled without kinking or bending assuring maintained tape quality.

DiagnoSYS
www.diagnosys.com

LCD controller with dual video ports

An LCD controller from Averlogic Technologies is intended for applications such as progressive scan TV, DTV/HDTV, LCD and projection TV and LCD projector. Available from 2001 Electronic Components, the AL310 features dual video input ports which allows the input of graphical and video input data formats with functions such as picture-in-picture overlay and alpha blending. The chip has an additional video input port for

applications that require input switching. As well as the ability to directly display panels, the AL310 has triple embedded video DACs, which provides RGB or YPbPr progressive analogue output. Resolutions up to SXGA (1280 x 1024) are supported. The device is capable of accepting multiple digital formats (RGB/YCbCr/YPbPr).

Averlogic Technologies
www.averlogic.com

Controllers have 1Mbyte EEPROM

Cambridge Microprocessor Systems has a range of UK designed and manufactured 16/32-bit 68000 controllers with 1Mbyte of flash EEPROM and 512kbytes of battery backed static RAM. All models in the range have a clock speed of 33MHz, five times that of a standard 68000. A 66MHz option is also available. Measuring only 100x110mm, QuickFIRE has an RS-232 and an RS-232/RS-485 serial port, two serial peripheral interface buses, two 16-bit pulse width modulation outputs and two 16-bit timer/counters. The controller provides up to 76 general purpose user configurable. digital I/O channels. on-board power management and a graphic LCD interface with colour option.

CMS
www.cms.uk.com
Tel: +44(0) 1387 875644

Proximity sensors for OEMS

The X-Sensors range of proximity sensors from Switchtec includes inductive, thru-beam, area and contrast sensors. Devices are aimed at OEMs and end users in the packaging industry, for use as security devices, for product line management and in automotive applications. Thru beam devices offer easy alignment and high contamination immunity for use in 'dirty' applications. The receiver/emitter combination is able to operate at up to 4m apart. Inductive sensors comprise amplified sensors operating at a nominal 10 to 30Vdc over a switching distance of 2mm in shielded form, and 4mm unshielded. Cross beams comprise a nine-beam transmitter with microcontroller designed to operate with the matching receiver unit.

Switchtec
www.switchtec.co.uk
Tel: +44(0) 1785 818600

EMI filter for DC-DC conversion

Vicor has introduced an active EMI filter for 48V DC-DC converter applications. The QPI-1 delivers over 40dB of common-mode and more than 80dB of differential-mode noise attenuation at 500kHz from a 24.5x24.5x5.1mm surface-mount package. Active filtering eliminates ringing on the input of the converter in response to load and line transients. An active filter attenuates noise over the entire frequency range. There are no resonant elements that can amplifiy the noise. The QPI-1 meets the specifications of the international 36 to 76Vdc telecoms bus. including the 100V, 100ms surge. Rated at 12A, the unit supports single or multiple DC-DC converters up to 576W at nominal line voltage. Units can be placed in series for higher attenuation or paralleled for higher currents

Vicor
www.vicor-eurpoe.com

WLAN card slot conforms to MD1

If you are adding a compact flash Bluetooth connection to a PC card, then the P312 PCI-to-CF card slot from Elan is designed to conform to the MD1 profile specifications.It features a compact flash Type 11 ejector which means it can also accept WLAN 802.11g CF cards when they emerge.

Elan Digital Systems
www.elandigitalsystems.com
Tel: +44(0) 1489 579799 ■

£11.99

Available from

Electronics World

All tracks on this CD were recorded on DAT from cylinders produced in the early 1900s. Considering the age of the cylinders, and the recording techniques available at the time, these tracks are of remarkable quality, having been carefully replayed using modern electronic technology by historian Joe Pengelly.

21 tracks – 72 minutes of recordings made between 1900 and 1929. These electronically derived reproductions are no worse than – and in many cases better than – reproductions of early 78rev/min recordings – some are stunning...

Pandora's drums

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Track

1 Washington Post March, Band, 1909
2 Good Old Summertime, The American Quartet 1904
3 Marriage Bells, Bells & xylophone duet, Burckhardt & Daab with orchestra, 1913
4 The Volunteer Organist, Peter Dawson, 1913
5 Dialogue For Three, Flute, Oboe and Clarinet, 1913
6 The Toymaker's Dream, Foxtrot, vocal, B.A. Rolfe and his orchestra, 1929
7 As I Sat Upon My Dear Old Mother's Knee, Will Oakland, 1913
8 Light As A Feather, Bells solo, Charles Daab with orchestra, 1912
9 On Her Pic-Pic-Piccolo, Billy Williams, 1913
10 Polka Des English's, Artist unknown, 1900
11 Somebody's Coming To My House, Walter Van Brunt, 1913
12 Bonny Scotland Medley, Xylophone solo, Charles Daab with orchestra, 1914
13 Doin' the Raccoon, Billy Murray, 1929
14 Luce Mia! Francesco Daddi, 1913
15 The Olio Minstrel, 2nd part, 1913
16 Peg O' My Heart, Walter Van Brunt, 1913
17 Auf Dem Mississippi, Johann Strauss orchestra, 1913
18 I'm Looking For A Sweetheart And I Think You'll Do, Ada Jones & Billy Murray, 1913
19 Intermezzo, Violin solo, Stroud Haxton, 1910
20 A Juanita, Abrego and Picazo, 1913
21 All Alone, Ada Jones, 1911

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. Where software or files are available from us, please email Caroline Fisher with the circuit idea name as the subject.

Send your ideas to: Phil Reed, Highbury Business Communications, Nexus House, Azalea Drive, Swanley, Kent, BR8 8HU email ewcircuit@highburybiz.com

Mains voltage deviation meter revisited

The original mains voltage deviation meter published in *Electronics World* November 1997 gave 1% accuracy but the simple zener reference was susceptible to drift which compromised the otherwise excellent accuracy and simplicity.

The following update circuit is immune from the problems above and still offers 1% reading accuracy.

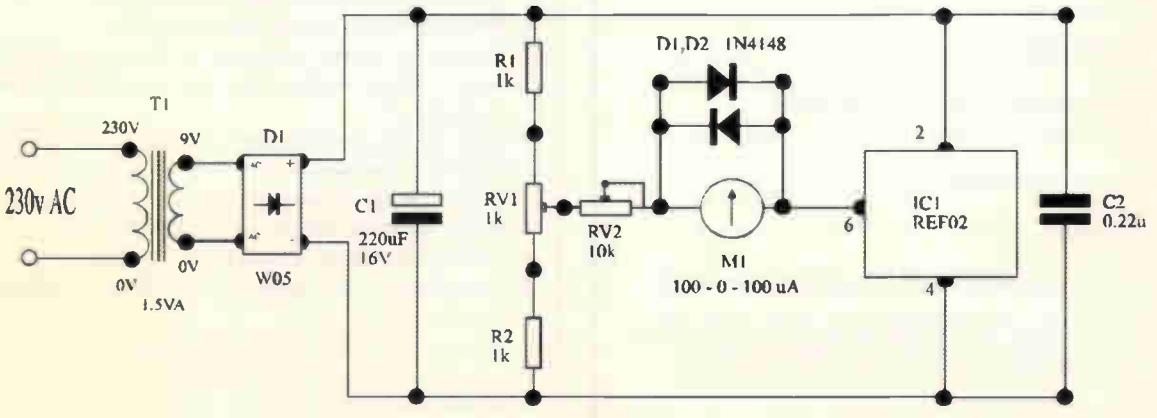
Mains is stepped down by transformer T1, rectified by D1 and smoothed by C1. This voltage varies in sympathy with the mains supply voltage. Circuit time constant is determined by the loading of the supply and the value of C1, hence the low value of C1, a small amount of ripple is of no consequence, as the moving coil meter will integrate this ripple to a large degree. A too higher value capacitor would cause excess damping and prevents transients being detected.

The reference side of the bridge is now formed by a REF 02 voltage reference, this gives a time and temperature stable reference of 5.00V. This device, though designed to source current, can sink up to 500µA without change in output voltage.

Bridge indication is provided by M1 and VR2. A moving coil meter was chosen as trends can be followed and transients can be observed. An LCD DVM type indicator was tried and while it gave a reliable steady indication, trends could not be followed and rapid fluctuation gave a confused jumble of digits. D1 & D2 offer protection to the meter under extreme conditions.

To calibrate the instrument set the input voltage to a steady 230V AC with a variac and adjust VR1 to give zero deflection. Now adjust the variac to give a mains input of 230V +10% = 253V AC and adjust VR2 to give the correct reading. The instrument is now calibrated. The two steps above may need to be repeated as a small amount of interaction may occur. A stabilised AC supply is helpful here.

Nigel. Goodman
St Leonards
Sussex



Simple voltage-to-frequency converter

An integrator followed by a schmitt trigger are at the heart of this simple voltage-to-frequency converter.

When an input voltage V_{in} is applied to IC_1, its output rises over the upper threshold V_T^+, causing the output of IC_2 to go low. This shuts down IC_1 and capacitor C discharges through the resistors labelled R.

As soon as the voltage over C falls to the lower threshold, V_T^-, the output of IC_2 goes high, activating IC_1 again and the cycle continues.

Components for the schmitt trigger can be calculated using the equations below with respect to voltages V_T^+ and V_T^-, which is a further advantage of the circuit. For R=1,

$$R_2 = \frac{\left(V_R^+ - V_t^-\right)\left(V_T^+ - V_{cut}^+\right) - \left(V_T^- - V_{out}^-\right)\left(V_R^+ - V_T^+\right)}{V_T^-\left(V_T^+ - V_{out}^+\right) - V_T^+\left(V_T^- - V_{out}^-\right)}$$

and,

$$R_3 = R_2 \frac{R_2\left(V_T^- - V_{out}^-\right)}{\left(V_R - V_T^-\right) - R_2 V_T^-}$$

In the circuit shown, the values used for the calculations were:

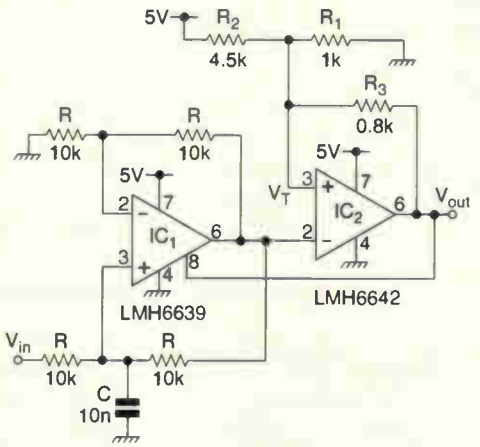
$V_T^+ = 3V$

$V_T^- = 0.5V$

$V_{out}^+ = 5V$

$V_{cut}^- = 0.1V$

$V_R = 5V$



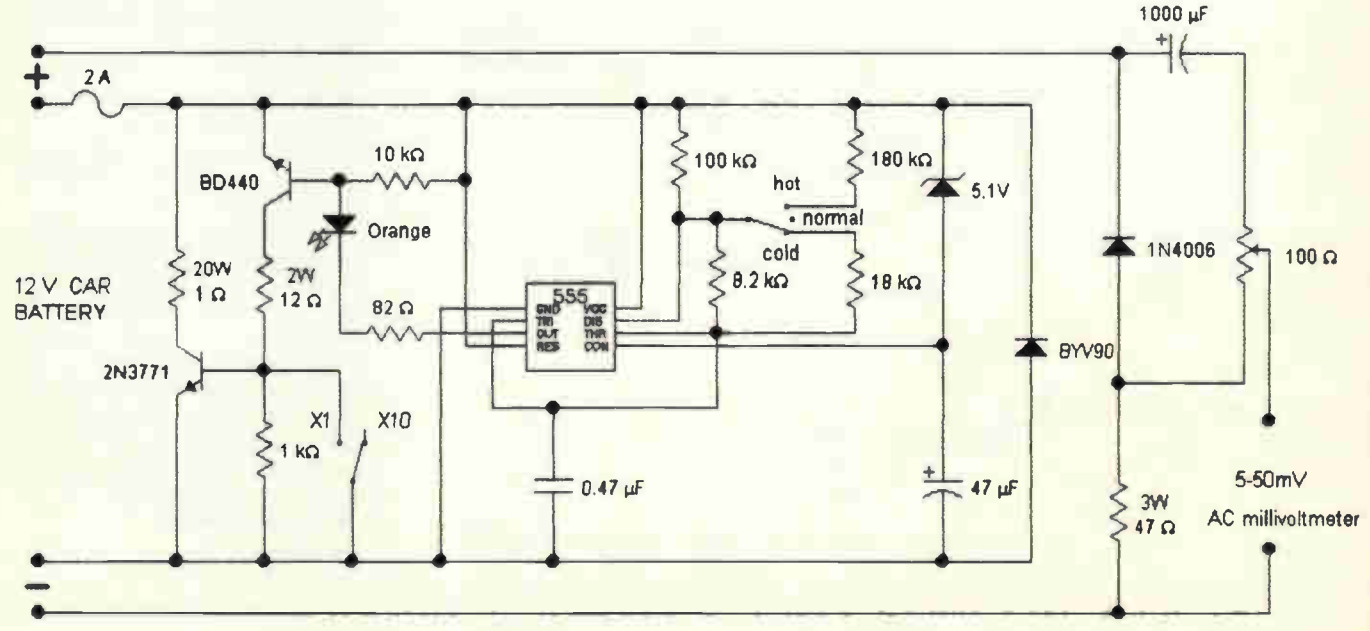
This voltage-to-frequency converter is simple yet convenient in that its component values are easy to calculate.

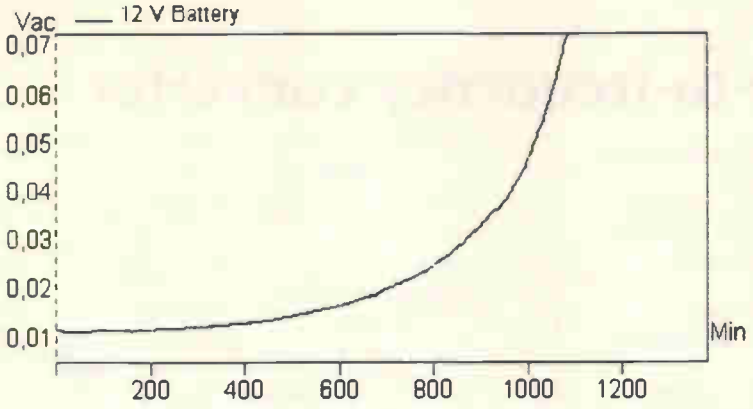
The LMH6639, IC_1, is configured as a high-speed positive-going integrator with a shut-down input on pin 8. Schmitt triggering is performed by IC_2, which is an LMH6649.

Kamil Kraus
Rokycany
Czech Republic

Car battery tester (32-100Ah)

Measuring the AC ripple on a car battery gives an indication of its internal impedance: the lower the ripple the smaller its impedance and the healthier the battery. The circuit is a controlled load connected to the battery 25 times a second and the resulting ripple is measured with a digital meter connected to the output. A fast flashing LED gives a n indication of its operation. The measuring leads must be kept separate from the wires connecting the load: they are both connected to the same battery lug but they should not touch each other. This will insure a reliable reading even if the electric contact is not perfect. The scale of the 100Ω pot is marked linearly from 2 to 100Ah, lower value towards negative. As I had the need to measure





AC voltage across a 12 V lead accumulator

batteries used in home alarm systems. I added a switch marked X1 - X10 which will change the measuring range to 3.2 - 10Ah.

An additional switch takes care of the battery temperature: hot is between 35° and 54°C (95°-129°F), normal is between 15° and 34° C (59°-94°F) and cold is between 4° and 14°C (25°–58°F). In order to get a reading set the range, the battery capacity with the 100Ω pot, then select the battery temperature, which is not necessarily the ambient temperature and connect the leads to the battery, there is no need to disconnect it from the car wiring but there should be no heavy load connected and the car must not be running. You should now read between 5 and 6mV for a healthy fully charged battery. The older the battery and the more the battery is discharged the higher will be the reading. A reading between 15 and 20mV should make you worry as the battery could be on its last legs or in dire need of a recharge. However it was found difficult to assess exactly when a battery is to be considered flat as it depends on the make, model, type and the car where it is used. The best practice is to experiment and see how the voltage changes during the year and you will find that the highest reading will be reached in winter.

The curve shows the AC ripple voltage of a freshly charged car battery: it is to be considered flat when the voltage is between 2 and 3 times the initial voltage.

D. Di Mario
Milan
Italy

Around the clock timer

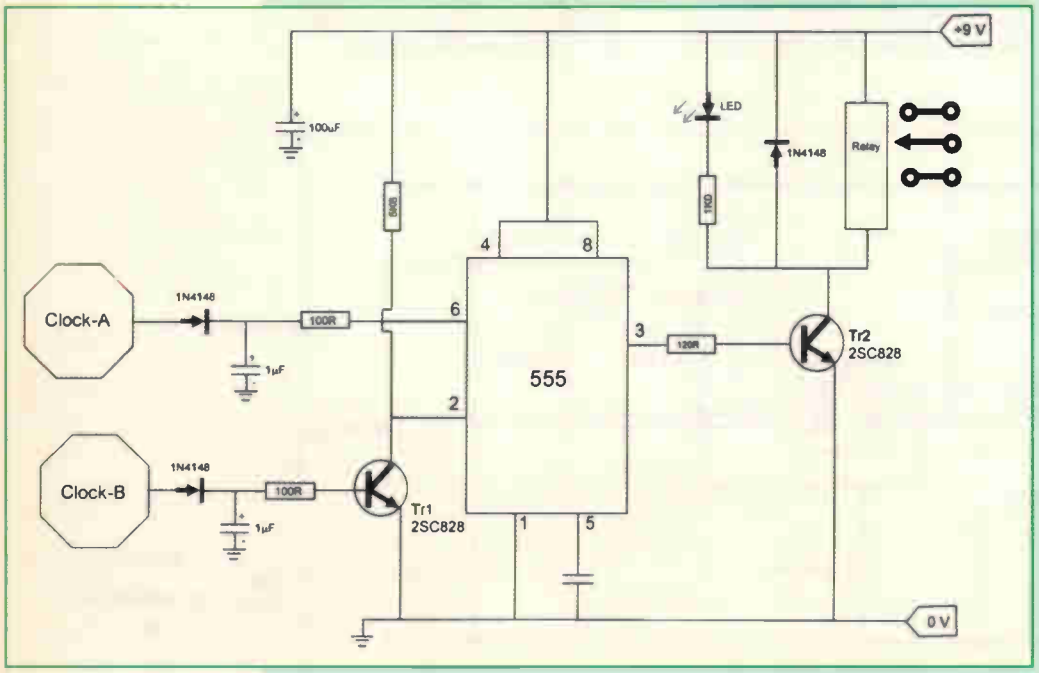
One of my friends started his career as a cable operator. He asked me to construct a timer circuit, which can shut down the whole of the cable network and restart it at a preset time. Many complicated and expensive circuits and designs are though already available in the market and circuit books, but I planned to address the problem in a different way.

The main idea was to use low price digital alarm clocks. The triggering pulses will be picked up from the wires going to the sounder of the clock when the alarm activates at a preset time. One clock will be used to trigger the relay circuit, such that the system will be shut down, and second clock will be used to restart it.

The famous timer IC 555 is used in the relay circuit. The IC consists of two comparators, a flip-flop, and an inverting amplifier. The lower comparator is triggered when pin 2 of the IC decreases to $^1/_3$ of the Vcc. This trigger pulse drives the flip-flop to a lower state and the inverting amplifier drives the pin 3 to a high state, which consequently energies the relay coil through Tr2. Vice versa, the upper comparator is triggered when pin-6 increases to $^2/_3$ of the Vcc. This trigger pulse drives the flip-flop to a higher state and the inverting amplifier drives pin 3 to the lower state which consequently de-energies the relay coil.

When the circuit is switched on, out put of the IC is at the lower state, so the relay coil is de-energized. Relay connections are arranged such that in de-energized state the load is in the ON condition. When a pulse from Clock A triggers the upper comparator at the preset time, the relay will energise, and the load goes to the OFF condition. Similarly, when Clock B triggers the lower comparator through transistor Trl, output pin 3 goes to the low state, which will consequently de-energize the relay coil and the load goes to the ON state.

M. Mansoor Malik
Rawalpindi
Pakistan



An unusual tuned circuit

It is known that an oscillatory circuit can be tuned to the wanted frequency by means of a variable resistor as it is shown in **Fig.1**. However the known tuned circuit is never used in practical circuits because the variable resistor abruptly changes the Q-factor as shown in **Fig.2**. The application of such tuned circuits in oscillators, for example, is very problematic because the reduction of the Q-factor can cause the break-down of oscillation.

The Q-factor of the proposed tuned circuit **(Fig. 3)** is stable, as shown in **Fig.4.**

The impedance of the proposed tuned circuit does not contain a reactive component at frequencies

$$f = \frac{1}{2\pi}\sqrt{\frac{1-\frac{C}{L}R^2k^2}{LC-C^2R^2(1-k)^2}}$$

where the coefficient represents the tuning of the variable resistor and changes from 0 to 1. This formula can be used for calculation of the oscillating frequency of the oscillators applying the proposed tuned circuit. The analysis of the formula shows that a single variable resistor can change the oscillating frequency of the oscillator from 0 to ∞ when

$$R = \sqrt{\frac{L}{C}}$$

Therefore a variable resistor can ensure wider frequency range of oscillator than a variable capacitor or a variable inductor.

S. Chekcheyev
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Russia

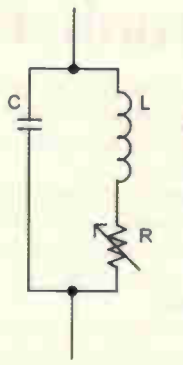


Fig.1. Known tuned circuit.

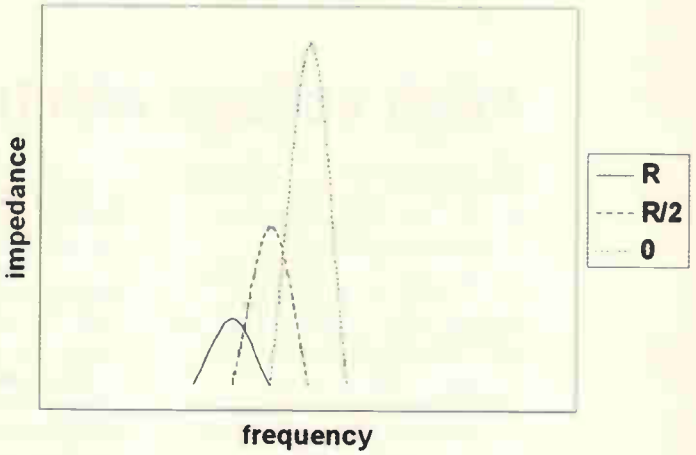


Fig.2. Impedance of known tuned circuit vs. frequency.

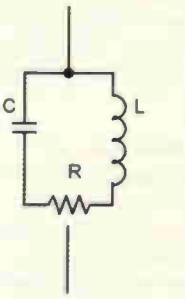


Fig.3. Proposed tuned circuit.

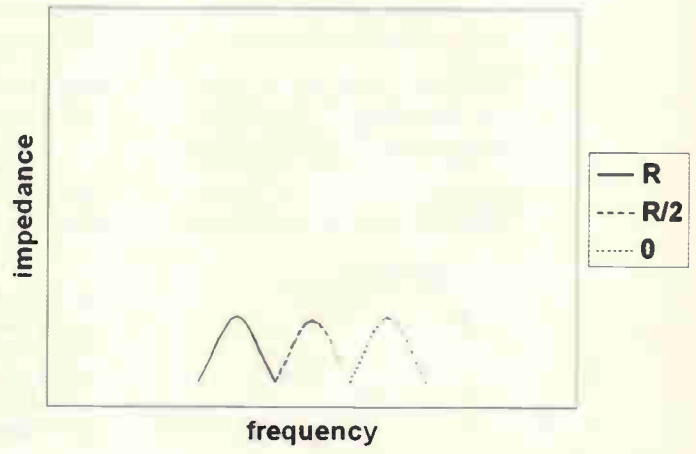


Fig.4. Impedance of proposed tuned circuit vs. frequency.

Manually controlled digital pot

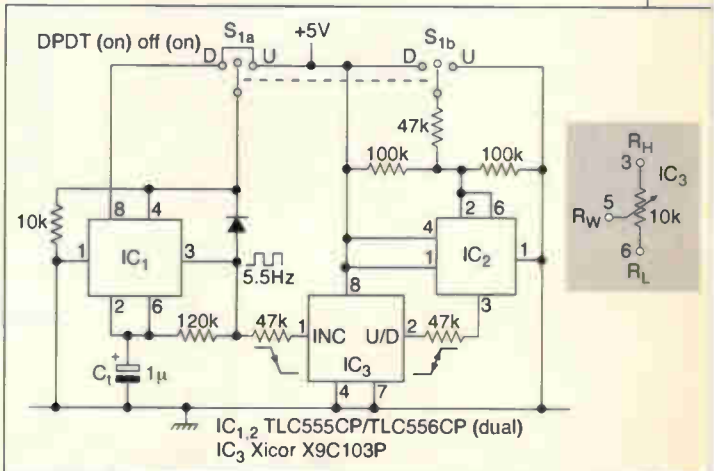
One panel switch controls single or stepwise change in wiper position of a resistive element with 1% resolution. This interface allows manual setting of amplifier gain or voltage, for example, without computer signals.

Gated oscillator IC_1 is latched in via diode feedback to the reset input, pin 4. This ensures a bounce-free turn-on when S_{1a} forces +5V at that node. It also gives coherent turn-off when pin 3 returns to ground reference and switch section 'a' has moved to centre off position. That has advantage for a one step change as the first cycle will be prolonged by C_t charging from zero to $^2/_3V_{cc}$. At 5.5Hz. A brief switch depression easily selects one clock cycle – a single step.

Direction, 'up' or 'down' is determined by the voltage at pin 3 of IC_2, which is configured as an R/S bistable device. Switch S_{1b} triggers the upper or lower comparator at $^1/_3$, $^2/_3$ V_{cc} for a bounce free transition to +5V or 0V well before the negative-going clock increment.

Terminal voltages for the potentiometer at R_L and R_H can be plus or minus 5V making this simple interface a flexible, efficient control for stand alone projects.

John Haase
Colorado State University
USA



Pressing the switch increases or decreases the resistance of the digital potentiometer IC_2 depending on which way you push the switch. Resolution of this digital pot is 1%. Connections for the potentiometer part of the IC are shown in the inset.

High voltage electronic fuse

Electronic fuses like 'Polyswitches' from Raychem or 'Multifuses' from Bourns show a heavy positive temperature coefficient.

They turn off – almost – if heated by a current that exceeds their nominal value. Afterwards they stay 'off', remaining hot due to their leakage current.

Such fuses only return to their original state after cooling down when the supply is removed. Unfortunately they also only handle low voltages – usually 50V for devices handling under an ampere or 30V for those handling more current.

This new circuit, **Fig. 1**, works with high voltages – even 220V mains or a valve's plates voltages.

A 36V Zener diode clamps the supply appearing at the electronic fuse to 30V via a MOSFET source follower. A diode prevents the parasitic capacitance of the MOSFET from discharging when the voltage is near zero

A 9.1V Zener diode protects the MOSFET's gate against overvoltage and a full wave bridge rectifier allows use of the circuit from AC voltage sources.

When overcurrent occurs, the fuse resistance rises. When 30 volts is reached, the MOSFET clamps the voltage, reducing the current to the value needed to maintain the fuse's temperature.

We tested the RXE series from Raychem from 100 to 500mA. Results are shown in **Table 1**.

All these devices are included in the same package and need near a watt to remain in high impedance state. Working on 220V mains, we choose the MTP3N50 MOSFET which is rated at 35W, 500V and 4A. We used a 5°C/W radiator.

For example, with a current of 310mA in an RXE010, the fuse worked within 10s, remaining hot with less than 25mA, dissipating 750mW. At the same time, the MOSFET was consuming 4.5W. With 127mA the fuse didn't work and the voltage across the AC side of the bridge was 6.5 V.

This circuit doesn't work for higher currents. Case sizes of bigger fuses are larger so they remain hot longer. The biggest Polyfuse I know is the RUE500, which is a 4 to 10A device with a 14 x 24.9 mm package. It should need more than 4W to stay open and the FET would consume some 20W in this state, but the MTP3N50 would not suit because of its 2.5Ω R_{DSON}.

As dissipation becomes more of a problem it also becomes necessary to invest more money in the circuit – **Fig. 2**.

In the high-power circuit, an MOC3020 coupler cuts the power when the there's more than 30V across the electronic fuse, a 32V diac being the trigger.

The MOC3020 remaining on, there is no dissipation in the IGBT that replaces the MOSFET of Fig. 1. This is because the current is completely cut off. However, the transistor needs a heat sink in accordance with the working current.

Hold current of the MOC3020 is delivered by the 220nF capacitor. This circuit can also replace the first one, if you don't want to waste 25mA in the blocking state.

Jean-Marc Brassart
Saint Laurent Du Var
France

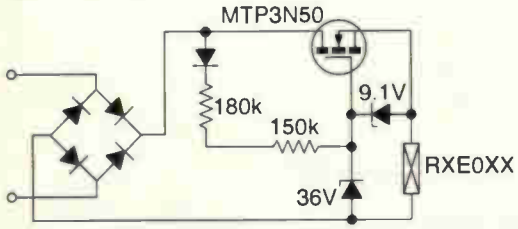


Fig. 1. Electronic fuses are usually only rated at 30 or 50V depending on their current handling capacity. This circuit ramps their usefulness up to much higher working voltages.

Table 1. Characteristics of a variety of electronic fuses.

Device	Hold current (mA)	Break current (mA)	Cold resistance (Ω)
RXE010	100 mA	200 mA	4.5
RXE017	170	340	5.21
RXE020	200	400	2.84
RXE025	250	500	1.95
TXE030	300	600	1.36

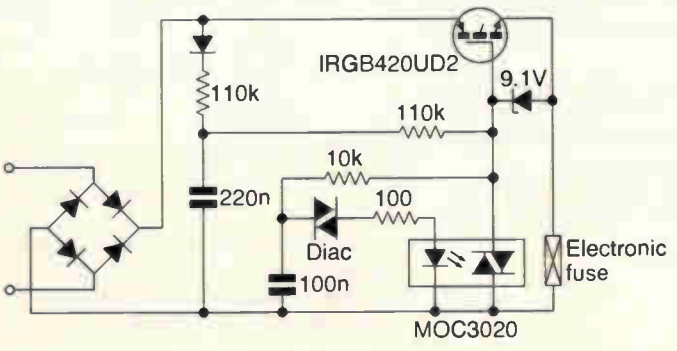


Fig. 2. If even higher voltages and currents are required from electronic fuses then this circuit may prove useful.

Novel battery charger

While building a portable instrument powered by a 12V 300mAh NiMH battery, I designed this simple charger circuit for it. I found – by accident I admit – that despite its extreme simplicity it has remarkably appropriate properties.

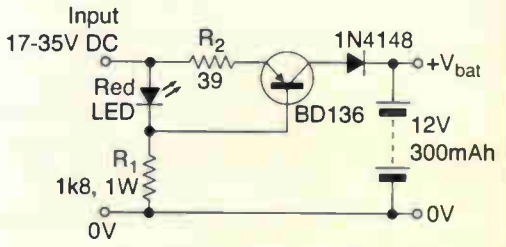
It is so simple it hardly needs description. Between 9 and 18mA flows through R_1 and the LED, producing about 1.8V across the LED. This voltage biases the current source R_2 and the transistor, which charges the battery. The diode is there to prevent battery discharge when there is no charging supply.

The remarkable thing is the behaviour of the LED. When there is no charging supply, it goes off. But if the charging supply is too low, the transistor saturates and the current through R_1 is diverted through the transistor and R_2, so the LED dims or goes off.

Further, if the battery is disconnected, the transistor again saturates, turning the LED off in the same way. The LED really is a 'charging' indicator. If charging current is not maintained, it goes off.

In my prototype, there is also a bridge rectifier and reservoir capacitor on the input so that the battery can be charged from various AC or DC sources. The choice of components was on the time-honoured basis of using bits I already had.

With the values shown, the charging current is about 30mA. For heavier charging currents, the transistor could be replaced by a super-α pair. The circuit can also be inverted, putting an n-p-n transistor in the negative side instead. The circuit will not work so well for low charging currents because R_2 then has too high a resistance to divert all the LED current, so it would only go dim rather than turn off.



Despite being very simple this charger, which was designed for a 12V 300mAh NiMH battery, has remarkable properties.

Alan Robinson
York

A digital frequency comparator

It is known that frequencies of two signals can be compared by means of watching Lissajous figures on an oscilloscope, but that method is not convenient. Besides, the Lissajous figures method is applicable only for sinusoidal signals.

The proposed circuit (**Fig. 1**) is more convenient. Its operation is very simple. LED2 illuminates when the frequency of the first input signal is larger than the frequency of the second input signal. Otherwise LED1 lights up.

The proposed frequency comparator can be used instead of the frequency meter when a reference oscillator is available.

More than that: the circuit of the frequency comparator can be built into the oscillator in order to make the oscillator work as a frequency meter. The oscillator can be connected to the frequency comparator as is shown in **Fig.2.** The indicators of the frequency comparator allow you to tune the

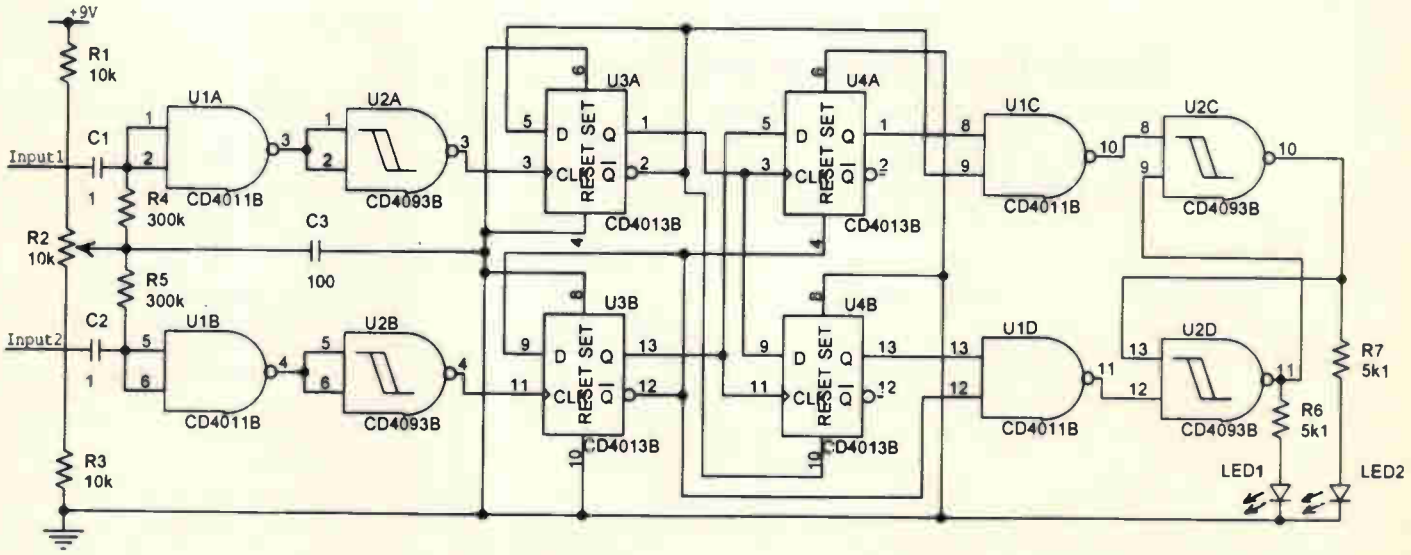


Fig. 1. Frequency compartor.

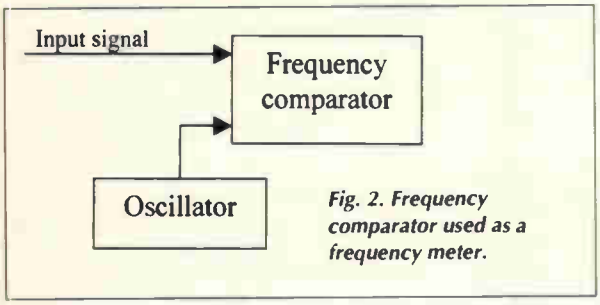


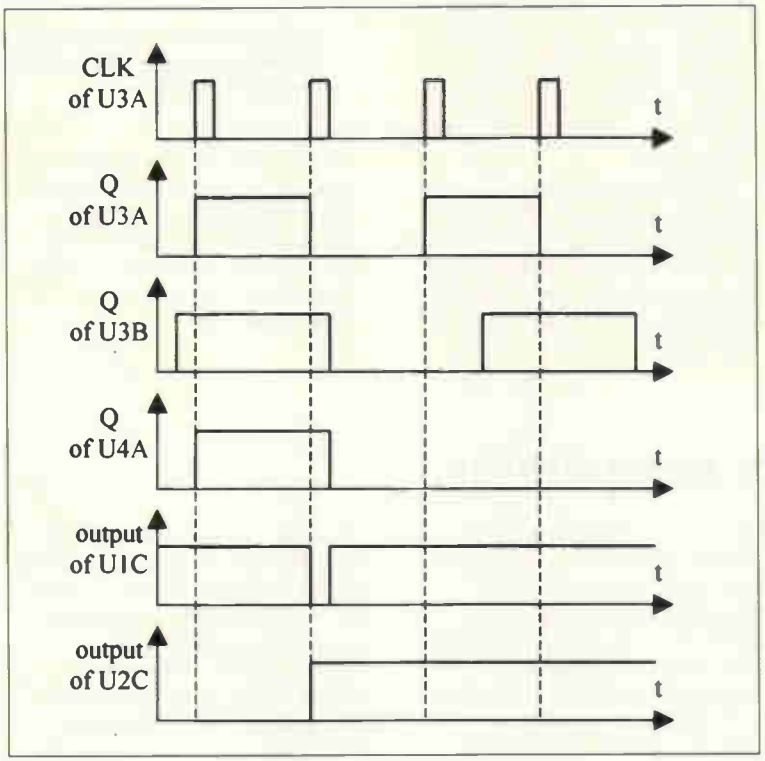
Fig. 2. Frequency comparator used as a frequency meter.

oscillator to the frequency of the input signal. After that the frequency of the input signal can be determined by the oscillator scale.

The device contains only four microcircuits. It can work with periodic signals of any shape. NAND gates U1A and U1B work as analogue amplifiers for the input signals. The adjustable resistor R2 ensures the necessary bias of the gates. Due to the Schmitt-trigger action, the NAND gates U2A and U2B transform input signals of any shape to the pulse signals. The relative pulse duration of that pulse signals depends on the shape of the input signals. D-type flip-flops U3A and U3B divide the frequencies of the input signals in order to remove that dependence. It is shown in Fig.3 that the pulse length of the output signals of that flip-flops does not depend on the pulse length of their input signals.

The circuit compares durations of output pulses of the flip-flops U3A and U3B. Let us assume that the Q-output pulses of U3A are shorter in comparison with the 'Q' output pulses of U3B (i.e. the frequency of the upper channel is higher than the frequency of the lower channel). In that case a Q output pulse of U3B sometimes overlaps a Q output pulse of U3A as shown in Fig.3. The positive edge of the Q output signal of U3A makes the Q output of U4A high, and the $\overline{Q}$ output signal of U3B makes it low. The output of the NAND gate U1C becomes low when the length of the output pulses of U3B exceeds the length of the output pulses of U3A. The low level of the output of U1C makes the output of U2C high and LED2 lights up. The RS-latch U2C-U21) remembers its state until the frequency of the first input signal becomes smaller than the frequency of the second input signal and the output signal of U1D will become low (the circuit of the frequency comparator is completely symmetric).

Fig. 3. Comparator signals.



The analysis has shown that the frequency comparator cannot change its output state when the frequency ratio $\frac{f2}{f1-f2}$ is ideally equal to an integer (1, 2, 3, etc) and the initial phase shift of the compared signals is ideally equal to zero. However the experiment with two real oscillators has shown that such an ideal situation never occurs and the indications of the frequency comparator are always correct.

The input sensitivity of the device is 100mV.

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Simple touch-senstive switch

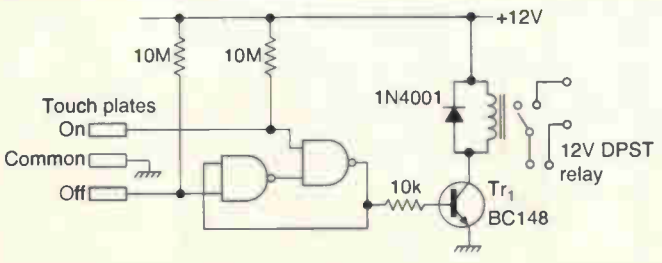
In this simple touch-sensitive switch, the relay shown is energised and de-energised by placing a finger on the touch contacts.

The RS bistable device, made up of the two NAND gates, is set when a finger bridges the 'On' and 'Common' touch contacts. It is reset when a finger bridges the 'Off' and 'Common' contacts.

Output from the top gate drives transistor to energise and the de-energise the relay in its collector circuit. The relay, in turn, controls the loads. A diode across the relay coil protects the transistor from back EMF induced in the relay coil during breaks.

Raj Gorkhali
Kathmandu
Nepal

With this simple touch switch, bridging two contacts with your finger turns a relay on or off depending on which contacts you touch.



802.11 Core technologies - the basics

Following from Ian Poole's article in the August *EW*, Eddy Insam discusses some of the low level technologies used in implementing Wireless Local Network standards.

There is something appealing about walking around the office or our house carrying a wireless enabled laptop. Slouching on a sofa, typing away, far from those open plan jungles. Not a filing cabinet in sight. Too hot? Walk out into the garden. Who says productivity cannot be improved by sitting by the pool with a glass of lemonade.

One thing we have certainly got used to over the years is the sophistication and the technology used to make these devices work. We simply take laptops, organisers and cellphones for granted. Many of us will not know exactly how the bits inside these things work, as their furious rate of development has meant that such information is simply non-existent. We may even have a suspicion that whatever is in there will be working on well-known techniques and on the same age old concepts we all learnt at college. But none of that will help when confronted with the simplest of technical problems, when we will feel completely demoralised and incompetent.

There is always the frustration of getting the things to work in the first place. We can buy a cellphone at the high street and it works more or less the minute we switch it on. WLANs are one of those technologies that just does not run into such pleasant black magic. The scenario is all too typical: we go to a computer superstore, buy all the bits, books and manuals, and as soon a we get home or to the office, we spend the best part of a day if not more, trying to get it all to work. So be warned, if you are intending to upgrade your office or house network for wireless. Make sure you allocate enough time and keep away from children and pets who may get in the way. On the plus side, and when it is all finished and working, you will have the satisfaction of being able to waste as much time boring colleagues and friends on explaining how we got it all to work (or not, as the case may be).

This article is not an installation or help guide. The writer is possibly one of the least qualified persons to do so as he hasn't even got his own system working properly yet. However, the article contains some useful basic pointers on how the basic technology works, and may help in deciding where problem areas are, and in planning for the future.

Basic Architecture and Security

A Wireless Local area Network (**WLAN**) is based on an architecture consisting of one or more radio cells. Each cell is called a Basic Service Set (**BSS**) and has a limited signal range depending on the power used, the environment, and most importantly, the presence of other sources of interfering radio energy in the area. In practice, range is limited to a few hundred square metres, but it can be a lot less in built up areas. Each cell is controlled by a base station called an Access Point (AP). The simplest WLAN system consists of a single cell with a single AP and one or more roaming wireless users, although it is possible to run cells without an AP with the roaming stations talking

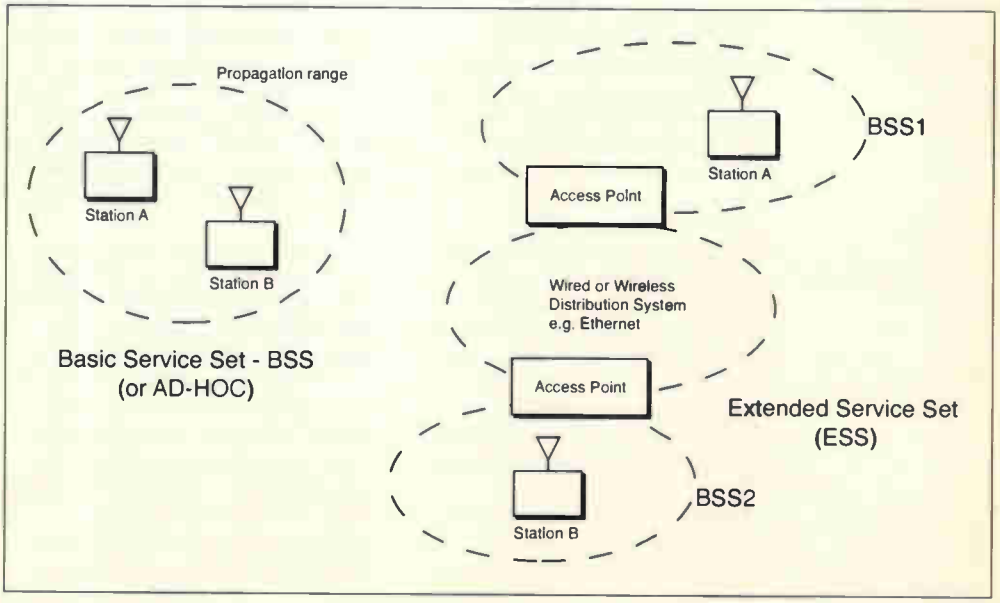


Fig. 1. WLAN cells have a limited range and multiple cells can co-exist. The base station software keeps a track of which remote belongs to which network.

Fig. 2. A sign you would not like to find on the pavement outside your office!

directly to each other. The AP is connected to the rest of the LAN network in the building via Ethernet or other existing wired infrastructure. Wireless stations can freely roam around a cell's radio area, walk into a cell controlled by a different AP, and enter cells allocated to other networks. Protocol software ensures that wandering wireless stations are recognised and enumerated as they fade in and out of reach, or in and out of the different Access Point's ranges.

The radio range of each of the cells can overlap. This does not matter too much as different radio channels can be assigned to each cell, This makes installation and planning easy, as there is no need to worry about overlaps, just ensure there are no 'no coverage' black spots. Because of the packet and error correcting nature of data communications, fade-outs and small black spots tend to be unnoticed, and the user sees a continuous transmission between their laptops and the base station.

This free wandering scheme also means that unauthorised stations could easily listen in or join into our networks. A WLAN enabled laptop can conceivably log in to any network if it is within radio range. After all, all the radios work in a similar way and use similar frequencies. Many readers may have seen news features on TV about 'hackers' sitting in the back of taxicabs or at coffee houses, logging into nearby office networks at will. This is because many network operators tend not to bother with enabling any kind of security leaving their networks open for snooping by anybody within radio range.

The IEEE 802 WLAN standards provide an encrypted security mechanism based on RSA's RC4 techniques called Wired Equivalent Privacy (WEP). In practice, even this scheme can be breached (see reference). Chasing unsecured networks in cities has given rise to the activity of 'Warchalking' where hackers with laptops mark the areas of coverage of local radio networks using chalk signs on the pavement outside offices. These marks indicate the frequencies and channels open for hacking in the locality. Wireless networks are not a secure system.

Why those frequency allocations?

The choice of frequencies for WLANs in the gigahertz range was dictated by the availability of large chunks of spectrum in the Instrumentation, Scientific, and Medical (ISM) radio bands. These allocations in the 900MHz, 2.4GHz and 5GHz bands are conveniently allocated around the globe, so a worldwide standard could be derived. At the same time, development of low cost, miniature microwave components has resulted in very cheap RF modules. The net result is a range of very low cost products using these frequencies. No wonder similar technologies such as Bluetooth, HomeRF, Ziggy, RF tags and video senders are sharing the benefits. (And contributing to the cross interference).

At present, the most popular frequency is the 2.4GHz band; with the 802.11b standard (also known as WiFi) the most popular scheme for networking. There may be a move in the future to the 5GHz band as soon as the present band becomes clogged with interference. Oddly enough, the end effect of interference pollution is range reduction, rather than making the band unusable (as it would be for AM broadcast), so expect the present band arrangement to stay for a long time.

The main designers of the current WiFi standards were the FCC in the USA and the IEEE via Subcommittee 802 (IEEE sub-committees follow a rather non sophisticated numbering scheme based on the date of creation: 802 is the 8th week of the second month, February). At the time, the IEEE wanted to ensure that the new standards were compatible with other local and wide area network standards in progress of definition. They also wanted to ensure the standards were more or less independent of medium of transmission, so the WLAN 802 standards apply not only to radio as the carrier, but also to infrared.

The first Wireless standard to be corroborated was 802.11, which defined methods for data transfer at 1Mbps and 2Mbps using either frequency hopping spread spectrum (FHSS) or direct sequence spread spectrum (DSSS) using radio, or pulse position modulation (PPM) using infrared. Commercial pressure to make the standards more compatible with fast wired Ethernet technology resulted in improved specifications and higher bandwidths (in exchange for fewer channels and reduced range). The enhanced 802.11b option offered 5.5Mbps and 11Mbps communications using more advanced modulation schemes. The penalty to pay was a reduction in the number of coexistent networks in the band (three as opposed to 79). Further enhancements defined in 802.11g increased the bandwidth to 54Mbps in the 2.4GHz band. Similarly, for the 5GHz band, 802.11a defined various data rates at up to 54Mbps. As these wider band systems were meant to operate over very short distances (e.g. within a room) problems of co-existence and interference with other networks became less of an issue.

The two radio based spread spectrum methods, FH and DSSS, were adapted by the standards to conform to the rather strict FCC regulations 15.247 which controls the use of the ISM bands. The FCC established the operating rules specifically to facilitate shared use of the band for the transmission of data and voice by multiple users in this unlicensed environment. The specific use of spread spectrum techniques where incorporated in order to minimise interference with these other services. For example, analogue video senders and microwave ovens operate at a rather constant carrier frequency. The use of packet modulation and spread spectrum techniques schemes could (at least in theory) go some way to avoid interference with these services. The extra complexity required of the radios is no problem with current miniaturisation techniques. However, efficient decoding of spread spectrum signals requires extra design qualities in radio receiver design, such as

extended dynamic range, which in many cases has not been achieved yet.

The task at hand - differences between wired and wireless networks

At first sight, it may appear that communication by radio is a simple task; one station transmits while the other receives. If the packets sent are not received properly after some checksum calculation, the receiver asks for a re-transmission. Is this all there is to it? Well, the collection of 802 standards easily fills a bookshelf. In particular, the standard relating to wireless systems is by itself quite a fat volume. So there is obviously a bit more to it. What needs to be realised is the number of tasks a wireless network has to deal with. In addition, wireless media is drastically different to standard Ethernet (a wired media system). This is a reason why the standard specifications covering Ethernet were not just enhanced to operate in a wireless environment. The main differences are:

Shared Boundaries: wireless has neither absolute nor observable boundaries. In geographical terms, the media can be shared with other similar wireless networks operating in different domains, including somebody else's computer networks. Stations can wander in and out of ours, and other people's domains while communicating.

Lack of full connectivity: In a wireless environment, we cannot assume that all stations hear each other (which is the basic assumption of a wired Ethernet system). The fact that a transmit station senses the medium around it as free does not necessarily mean that the medium is free around the receiver area.

Time varying propagation properties: signal levels may change drastically or fade out completely for relatively long times during a session.

Destination address does not necessarily mean destination location: In wired LANs an address is always related to a physical location. In wireless LANs, an address can be a message destination, which is not necessarily a fixed location.

Real time collision detection is impractical because this would require full duplex radio sets. Collisions can only be predicted/assumed rather than actively detected.

In wired systems, medium error rates are minimal and error management is usually implemented by the higher layers of a protocol. A typical radio link may need to tolerate much higher error rates, which implies that wireless systems must include some form of error correction capabilities at the local level.

The standards and encoding methods used

As with other IEEE 802 protocols, 802.11 contains a management (MAC) layer which deals with addressing and packet management, and a Physical Layer (PHY) which deals with interfacing with the medium. The current IEEE standards define a single MAC, which interacts with three optional PHY layers: FHSS, DSSS and infrared. Of these three, DSSS is used in most implementations today. FHSS and infrared are rarely used.

For FHSS, the FCC defined a minimum 1Mbps rate using a two level Gaussian frequency shift keying method (2GFSK) with an optional 2Mbps rate using four level keying (4GFSK). This is basically a frequency modulation scheme where binary ones and zeros are represented by two (or four) closely spaced frequencies. Further to this, the average carrier frequency is caused to slowly jump around the 79 allocated channels in the band in a pseudorandom (PN) fashion, with the receiver tracking the transmitter as both generate the same timed PN sequence. The FSK modulation scheme is preferred in FHSS systems as it is difficult for the hopping synthesiser to maintain phase coherence over the wide hopping bandwidth, FSK is also relatively easy to demodulate non-coherently. The centre frequencies for the 79 channels are defined in 1MHz steps beginning at 2.402GHz and ending at 2.480GHz. (Other similar allocations are defined for Europe and Japan). The actual hop rate is not defined by the specifications but is usually greater than 2.5 hops per second. In order to maintain the 1MHz channel spacing and keep with the strict FCC bandwidth requirements, the FSK modulation index is kept small (maximum of ±160kHz) resulting in a non-optimal modulation scheme. One side-effect of this is that attempts to increase channel capacity by addition of multi levels results in a degraded signal to noise ratio trade-off. In other words, the use of narrow FSK becomes impractical for higher capacity systems due to the prohibitively high signal-to-noise ratios required for the constrained bandwidth specified.

Where more bandwidth is required, a DSSS approach is used. Like FHSS, DSSS uses a PN code to spread the signal. The DSSS encoding method used in 802.11b is not new. Similar technology is being used in GPS satellite navigation systems and in CDMA cellular telephones. In the basic 1Mbps DSSS system, the information data stream at 1Mbps is combined via an exclusive or (XOR) function with a high-speed pseudo-random numerical sequence running at 11MHz. The PN specified by

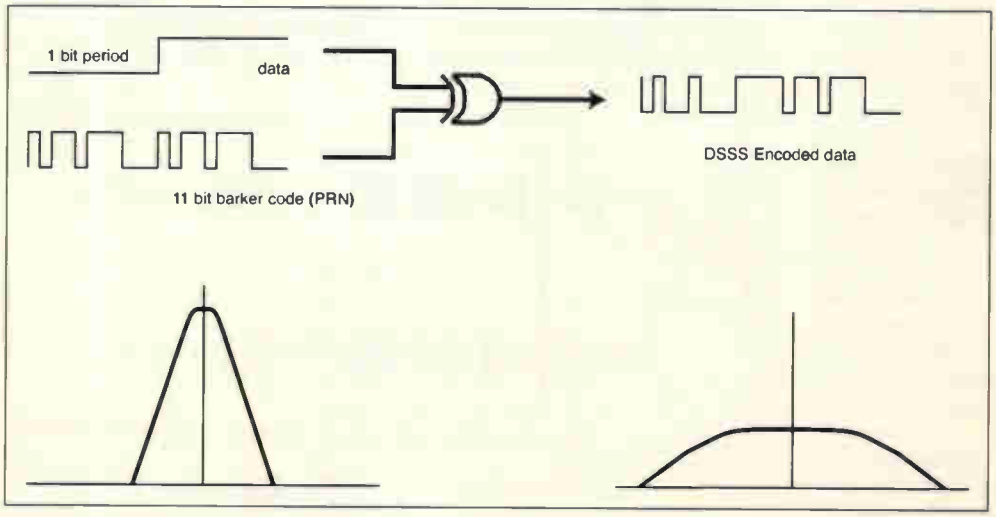


Fig. 3. Spread spectrum is simply obtained by XORing the data stream with a much faster, predictable "chip" sequence. This has the effect of spreading the transmitted power over a wider band. The receiver must be able to receive the whole bandwidth and recover the original data stream using an inverse process.

802.11b is an 11-chip Barker code. This particular sequence has well known autocorrelation and comma-free properties that makes it suitable for this application. The term chip is normally used to denote bit positions in a PN to denote the fact that the Barker code does not carry any binary information by itself. The result of the XOR operation is an 11Mbps digital stream, which is then modulated onto the 2.4GHz carrier using Differential Binary Phase Shift Keying (DBPSK), i.e. the carrier phase is inverted or not inverted depending on the incoming signal binary data transitions. The effect of the pseudo random modulation (or scrambling) is to spread the transmitted bandwidth of the resulting signal by a ratio of 11:1 (hence the term 'spread spectrum'). The total bandwidth required is just under 20MHz; the peak power of the signal is also reduced by a similar ratio. DSSS signals are nominally spaced 30MHz apart, so up to three DSSS networks can coexist in the 2.4GHz band (note how the use of DSSS reduces the original 79 channel band capacity to a maximum of three users). Upon reception, the signal is recovered using a correlation process with a locally generated version of the same PN chip sequence. The correlation process has a significant benefit; it reduces the level of narrow band interference, which falls in band by the same 11:1 ratio. This effect is known as processing gain.

In the 2Mbps DSSS option, the modulation system used is Differential Quadrature Phase Shift Keying (DQPSK), which effectively doubles the bit rate without increasing the radio bandwidth. This is done by modulating both the in-phase and quadrature versions of the RF carrier (known as I and Q modulation), the penalty to pay is a slightly lower signal to noise ratio.

The faster 5.5Mbps and 11Mbps DSSS options use Complementary Code Keying (CCK) to further compress the data rate, while still maintaining the same overall bandwidth. In CCK modulation, input data is XORed with spreading chip sequences much in the same way as in the 1Mbps system described above. However, the chip sequences have 8 chips each (as opposed to Barker 11 chips sequences). Each data input symbol is modulated with one of 256 chip sequence combinations to produce an 8-bit data message for every symbol to be transmitted. The chip codes are based on complementary codes, which are in turn related to Hadamard and Walsh functions. Complementary codes have the important property that the cross correlation between any two codewords is zero, so a data stream can be detected by the receiver implementing a number of parallel matched filters, with each 'tuned' to one codeword, a majority detector then selects the strongest output. On the 11Mbps system, each data byte to be transmitted is partitioned into a 6-bit selector, which his used to select one of 64 chip spread sequences, and the other 2 bits are used to phase invert modulate that symbol. Each of the 64 sequences contains terms in the I and Q phases. Thus, the total possible number of combinations of sequence and carrier phases is 256. The 5.5Mbps option operates in a similar way, but does not use quadrature modulation.

How do the radios work?

By radios, we really mean chipsets.

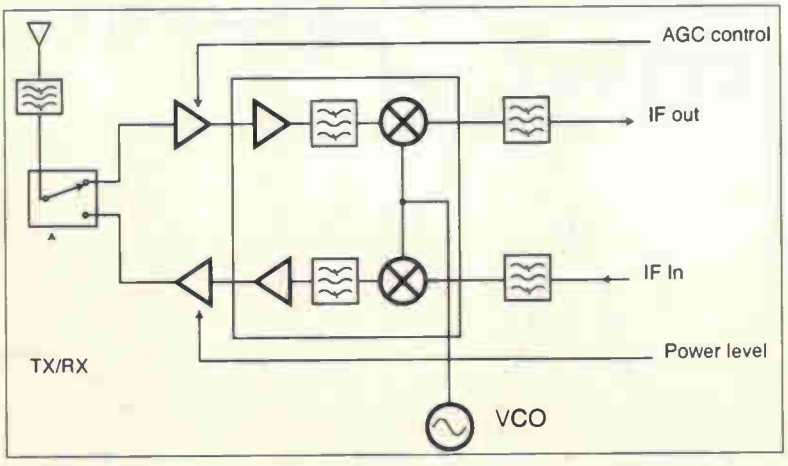


Fig. 4. Front end of a typical 802.11b radio. These simple designs lack sophistication in terms of signal handling, but are perfectly adequate for the short distances involved. Both transmit and receive gains are constantly controlled to ensure a fixed voltage at the detectors. The latest generation of radios do not use intermediate frequencies and convert directly to baseband from RF.

Most manufacturers offer IC chip combinations that can be assembled onto PCBs to form complete radios. A typical chipset is the Intersil PRISM family of devices. These are based around common modules such as RF amplifier, TX power driver, IF amplifier, mixer, baseband processor, and MAC logic interface to the host or network. A number of supporting ICs include PLLs, duplex switches, and passives such as coils, crystals and band-pass elements.

Signal from the built-in 2.4GHz aerial is connected via passive band-pass filters to the TX/RX switch, which may also contain a RF amplifier and a programmable power TX amplifier. The signal is now split into the TX and the RX channel. Both are fed into the RF/IF chip, which mixes the signal down to an intermediate frequency of around 280MHz. The local oscillator is a VCO or PLL controlled by serial signals from the control microprocessor. The oscillator may range between 2132MHz and 2204MHz, to give the required IF frequency. The use of such a relatively low IF frequency makes filtering much easier, although the latest generation of radio sets can now do mixing directly to baseband from 2.4GHz rate without the need for an IF stage.

Discrete band-pass IF filters are used to limit the bandwidth to just under 20MHz, enough to isolate one channel. The IF signal goes through two limiting amplifiers and further 280MHz SAW band pass filtering. The purpose of the limiters is to fix the amplitude of the signal to a relatively fixed value (around 200mV) under all input signal conditions. This levelling is in addition to the AGC provided by the variable gain RF and IF stages. The integrated receivers have limited intrinsic dynamic range, and their gain needs to be critically adjusted by the control microprocessor in order to ensure the radio outputs levels are relatively constant in amplitude. In practice, both RF and IF amplifiers are fed from an AGC signal derived from a D/A converter to provide this compensation

The signal is then demodulated to baseband using two quadrature multipliers operating at the IF frequency. The reference used is a locally generated VCO phase locked 560MHz oscillator, from which the

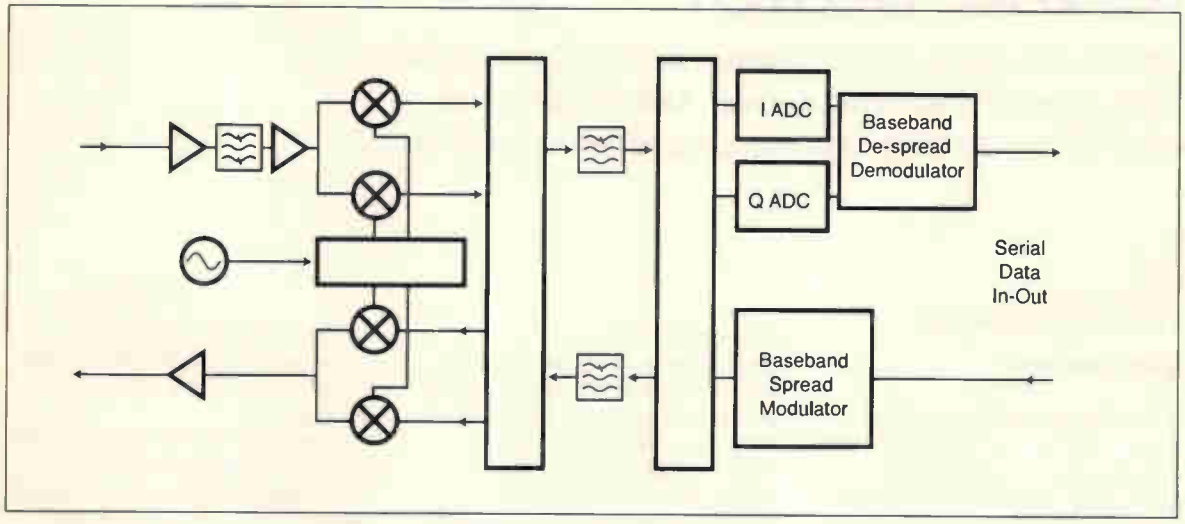


Fig. 5. Baseband detector uses quadrature demodulation and simple three-bit A/D converter sampling at about three times the chip rate to feed the all digital convolutional decoder.

two quadrature 90 degree out of phase signals are generated. The two resulting quadrature outputs are low pass filtered and fed to the baseband processor. Here, the two signals are analogue to digital converted in wideband 3-bit converters at a rate of 22Msps, which results in two 3-bit data samples per input chip. At this point, the signals are baseband spread spectrum and of a constant vector amplitude. In other words, each quadrature I and Q input may vary in amplitude, but their combined vector sum will be constant in amplitude. The baseband processor then correlates the signal with a locally generated PN spreading to remove it and to uncover the differential BPSK (or QPSK) data.

At this point the data packets. which now resemble MAC packets, are fed to the MAC processor (not shown) All packet signals have a preamble followed by a header containing a standard IEEE 802 frame including a start frame delimiter, headers and a cyclic redundancy check (CRC). The MAC processes the header data to locate the start of frame, determine the mode and length of the incoming message and check the CRC. The MAC then processes the packet data and sends it on through the bus interface to the host computer. The MAC also checks the CRC to determine the data purity. If corrupted data is received, a retransmission is requested locally by the MAC, as specified in the IEEE protocol specifications.

The MAC processor also includes methods for synchronising the link and establishing timing relationships. The system initially uses simple differential detection to identify and lock onto the signal. It then makes measurements of the carrier and symbol timing phase and frequency and uses these to initialise tracking loops for fast acquisition. Once demodulating and tracking, the processor uses coherent demodulation for best performance.

For transmission, the baseband processor scrambles the packet and differentially encodes it before applying the spread spectrum modulation. The data can be either DBPSK or DQPSK modulated and is a baseband quadrature signal with I and Q components. The BPSK spreading is a chip sequence that is modulated with the I and Q data components. Transmit quadrature single-bit digital inputs are low passed and applied to the quadrature IF modulator/demodulator from the baseband processor. The IF signal is bandpass filtered and applied to the up mixer. A variable gain RF amplifier feeds the transmit aerial with a controlled signal. This is to ensure no more RF signal than necessary is transmitted.

A word on spread spectrum

Spread spectrum techniques can be used to improve the performance of a communications channel. However, it is important to realise that for a receiver to be able to realise this potential, they must be designed to a higher specification than their normal counterparts. Specifically, they must be designed to detect small wanted signals in the presence of large amounts of background noise. This implies the receivers must possess much improved levels of linearity and dynamic range. Many simple WLAN radios do not posses such characteristics, with a corresponding degradation in performance. Remembering that WLAN systems are designed to work within a very local environment, this is not much of a problem.

The future

The fast growth of the use of these bands will ensure they will become saturated sooner than later. With the proliferation of 2.4GHz equipment and interference, the bands are bound to get congested very quickly. The net result will be reductions in range to feet rather than yards. Any forward planning for the use of equipment in this band must take this into consideration.

The Author

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For More Information

To obtain the full set of relevant IEEE standards, visit the IEEE website, or enter "802 IEEE" from any online search engine.

AT&T Lab Technical Report TD-4ZCPZZ. "Using the Fluhrer, Mantin and Shamir Attack to Break WEP" August 2001.

Also, Eddie Insam's book *TCP/IP Embedded Internet Applications* (published August 2003 by Newnes) gives more information about LAN and WLAN technology and principles.

LETTERS
to the editor

Letters to "Electronics World" Highbury Business Communications, Nexus House, Azalea Drive, Swanley, Kent, BR8 8HU
e-mail EWletters@highburybiz.com using subject heading 'Letters'.

Throwing stones in glass houses

In the forward to *EW*, Aug 2003, Nigel Cook, implied the general view that physicists such as Einstein, Plank, Dirac, Hiesenburg. Bohr, etc. etc. were rather short on their mental facilities. He made a claim that "the reason why the electric speed is that of light ...remains disingenuously unexplained".

On the contrary, the best current and fully accepted theory of EM is Quantum Electrodynamics (QED) for which Richard Feynmann got the Nobel Prize. This theory explains EM by the momentum exchange of virtual photons between charges, hence why EM phenomena must travel at the speed of light. It has been proven accurate to 13 odd decimal places, and has never failed in millions of experiments, by 1000's of physicists, over 60 years. Regarding Nigel's further comment, and others, on the so-called 'Catt anomaly', this whole subject matter is really a bit of a red herring. Maxwell's equations are simply wrong. They cannot be used to explain all the results of EM. This was decisively proved in the early 1900s when it was realised that accelerating electrons in an atom do not radiate their energy away and therefore spiral into the atom, in addition to other such effects as the photoelectric effect.

Maxwell's equations are only an approximation to the correct quantum theory, so it's not surprising that issues can arise in using them to understand all EM phenomena. Indeed, the technical details of exactly how momentum exchange in QED, via the Hiesenburg uncertainty principle, can correctly account for an attractive force is usually beyond the normal training of electronic engineers such as Ivor Catt.

Kevin Aylward
Peterborough
Cambridgeshire
UK

Metal detector

I would like to thank Ezio Rizzo for his additional information to my article on Metal Detection. It appears from his work that the change of inductance of a coil is dependent on both the frequency of measurement and the presentation of the sample to the coil. I am totally convinced by his observations of inductance change but would like to stress that the BFO metal detector has been largely replaced by more sophisticated detectors which can distinguish between ferrous and non-ferrous objects.

Frank Thompson
By Email

Book request

Why don't you get somebody to collate all the articles written in the past 60 years or so by Cathode Ray and publish them as a book? Whoever he was (probably Marcus Scroggie) always wrote with such clarity and horse sense that if you were to do it, you would probably find that the demand was a lot greater than you might have imagined. Or have you already got something like this on the go and am I out of date?

Frank van Vloten
South Africa

An excellent idea Frank. But the last time something like this was mooted – in fact an article on Scroggie – I got two offers of help – not really enough to justify such a project. But lets throw down the gauntlet again – anybody out there feel up to the task? – Ed.

Degrees devalued

I would like to take issue with Dr. Les May's problems with people calling certain degrees "Mickey Mouse". He used the example of a "Mickey Mouse" degree being a Media Studies degree. and it is a good example, because watching TV in 'lectures' is hardly as difficult as, say, wrestling with advanced mathematical concepts that you find in maths, engineering and science degrees.

As someone who has studied at four universities, ranging from the very bad (ex-poly) to the very good (as the Sunday Times University League Table ranks them), then I can at least speak with some level of experience on this matter, and I have to say that the difference between the good and the bad universities is enormous. Most students at the ex-poly I attended simply would not have been able to cope with the pace of lectures or the difficulty of the material that was presented at the best university I attended, yet students from both universities come out with a degree in the same subject. So, to me at least, it seems obvious that because of the conversion of polytechnics to universities, ownership of a degree has been devalued.

Although the UK Government's target of getting 50% of 18-year-olds into higher education is good from the point of view of improving the overall education of the workforce, now that all the polytechnics have been converted to universities we will have about 50% of people with degrees. As little as 15 years ago, those with degrees stood out from the crowd, now it just means that you're part of the crowd, because there's too little differentiation between easy and difficult degrees, and between good and bad higher education establishments.

And if Dr. May is under any illusions as to whether there are easy and difficult degrees I propose a test: Ask 100 Media Studies students to solve a 2nd order homogeneous differential equation (a relatively simple task for any 1st year maths, engineering or physics student), and set any Media Studies task for 100 maths, engineering or physics students to complete, and compare the results.

I also disagree with Dr. May about his implication that engineering, electronics, computer science or applied anything as being one step up from your average ex-polytechnic vocational degree course. My experience was that students look upon such courses as being relatively difficult, and I somehow doubt that you could compare them to Pop Music or Surf Science (i.e. surfboarding to you and me) degree courses, both of which are now available at UK ex-polytechnics.

Steve Green
Manchester
UK

CE marks

Andrew Denham (Letters, *EW* November), like many other people, has misunderstood the meaning of the CE mark. It is supposed to mean that the item or equipment meets the requirements of ALL APPLICABLE directives. Thus a mains lead must comply with the Low Voltage Directive (I think we would all agree that it should not give electric shocks amongst other requirements), but does not come under the scope of the EMC directive (except sometimes as part of another piece of equipment, and covered by that equipment's CE mark). A lot of equipment has to comply with several directives, some of which stipulate additional markings. If there is sufficient interest, perhaps the editor can find an expert who is prepared to explain the complexities of the CE-marking rules in just a page or two!

Arthur Prent
Brentwood
Essex
UK

Another gauntlet – I would certainly like an explanation of CE marking and how it applies to our industry. Any takers? - Ed

Cover photo mystery solved

I won't be the first, I'm sure, to comment on your response to the gentleman inquiring as to the location of the hurricane picture; "...no reference on the picture CD..." indeed.

Let's see. Lake Okeechobee easily visible in the very prominent Florida panhandle, and the distinctive Great Lakes at the top of the picture?

Your required 'reference' was as close as the nearest Public School underachiever carrying an outdated 1/2" thick classroom-issue atlas of the world.

O.K. I'm done. Thanks for letting me rant.

Newt
By email

If I may be of service, the view is one quite recognisable to pilots and meteorologists: The eye of the storm is at approximately 30° 50' 4" Lat. N (North Latitude), 78° 10' 1" Long. W (West Longitude) which is just above the Tropic of Cancer in the West Longitude just off the southeast coast of the United States. (The storm is rotating counter-clockwise, which coincides with a northern hemisphere location.)

Prof.Terry Haynes
Westchester NY
USA

The Michigan peninsula is visible at the top left of the photo with Lake Michigan to its west and Lake Huron to its east. Lake Erie is obscured by the storm, but Lake Ontario is visible just north of the storm mass. At the eastern extremity of the landmass, the coasts of North Carolina and Virginia are visible immediately to the north of the storm's main body.

As to why an American weather satellite is positioned over Cuba, I have complete confidence that it is for scientific purposes only.

Herman Respess
Shelby Twp
MI
USA

Looks to me like the eastern united states - florida at the bottom to the great lakes at the top, the easten seaboard on the right.

Martyn Preston G0THY
Kidlington Oxon UK

The coastline under the storm is fairly distinctive. The shapes of the great lakes on the top of the picture are more distinctive still.

Jussi Suuronen
By email

I'm an engineer and consider my geography awful, but even to me it seemed obvious the location of the storm pictured was off the east coast of Florida. The land mass is clearly visible. A quick check with my pocket atlas confirmed it. Lake Okeechobee is clearly visible.

Robert Atkinson, MRAeS, G8RPI.
By email

I am surprised it isn't clear that the storm is off the US east coast. Hidden behind the top left of the storm is Georgia and South Carolina. Below and to the left of the storm is Florida – the dark spot is Lake Okeechobee.

Above the storm the coast shows the Chesapeake and Delaware bays quite clearly. New York and Long Island are under the horizontal yellow cloud towards the top right. In the extreme top right of the photo, you have the Gulf of Maine just clear of the cloud.

South of the storm and partially hidden by the broken cloud is the Bahamas. The island of Andros can be seen along the bottom of the picture an directly below the storm. You can see at least twenty two American states, some of Ontario and a little of Quebec.

Mick Walker.
Stevenage
Herts
UK

Further to the letter and your reply regarding the storm location on the July cover. A quick look in my atlas, conformed that it is in the Atlantic. The Miami peninsula can be seen to the south west and Andros island in the Bahamas to the south. To the coastline visible north of the storm, is just south of Norfolk, Virginia.

Mark Buja
By email

In answer to the question from A. Walting in the letters section of the November issue, it looks like

a hurricane or tropical storm just east of Florida.

Harbanse Deogan
Groby
Leicester
UK

Well it is not just a storm it is a full hurricane with a well defined eye at the centre of the cloud complex. The Florida peninsula is clearly outlined under the clouds in the lower left corner of the picture.

After a quick search in the hurricane archive it appears to be the photo of hurricane Floyd, see the attached picture, one of the deadliest and costliest to hit the USA coast on September 1999.

Francesco Bracchi
Milano
Italy

It looks to me like the South East United States with Florida at the bottom and the coastline up to Washington DC can also be seen on the right hand side of the photo. Going inland Lake Michigan can be seen at the top of the photo together with Lake Erie also further North is the Canadian Border, which in parts can also be seen. I hope this helps to clear up the mystery.

Richard J C Reynolds
Guildford
Surrey
UK

Well, that about settles it. We all thought that the picture was of Florida, but rather than print an erroneous caption, decided to leave it up to you to decide! Ed.

The story of 'O'

I hesitate to cross tinned leads with the gurus, but the 'O' in OC71 really *was* an alphabetic O. I was in the lab long years ago when the first ones came through the door. They were for a time packed in valve boxes which made them quite difficult to find.
More to the point, I am looking at the Reference Manual of Transistor Circuits (Mullard, 1960) and all references to transistor types say OC in a typeface which clearly distinguishes figures. The same applies in contemporary radio service sheets as published in *Radio & Television Servicing*, carefully edited by the excellent Pat Hawker. Even, if I am allowed to mention it in these pages, *Wireless World Radio Valve Data* 1958!
The reason is not far to seek and it lies in the fact that they were Mullard devices. It is true that in the RCA notation used in the US, the first number is the heater voltage with the following letters and figures arbitrarily defining the valve's function. There are a few listed types starting with 0, cold cathode rectifiers or neon stabilisers but the system was never applied to semiconductors in the US. However, Mullard were part of the Philips group whose valve notation - widely adopted in Europe - is distinctive. The first character is a letter which defines the heater voltage or current, D for 1.4V, E for 6.3V, U for 100mA series chain etc. The use of letter O to represent zero heater volts was not inevitable but it probably seemed to be a nice touch. The second letter is the valve structure. A for a diode, B a double diode, C for a triode (but CC for a double triode) an so on, followed by figures which define the actual valve within the category. All of the available semiconductors were either diodes or triodes, hence OA70 and OC71. The figures were apparently arbitrary but the first digit gave some indication of where the device went in the circuit - OC42 RF, OC29 o/p etc.
I think that this wraps it up.

Michael Hawkins
Farnborough
Hampshire
UK

Oh not Zero!

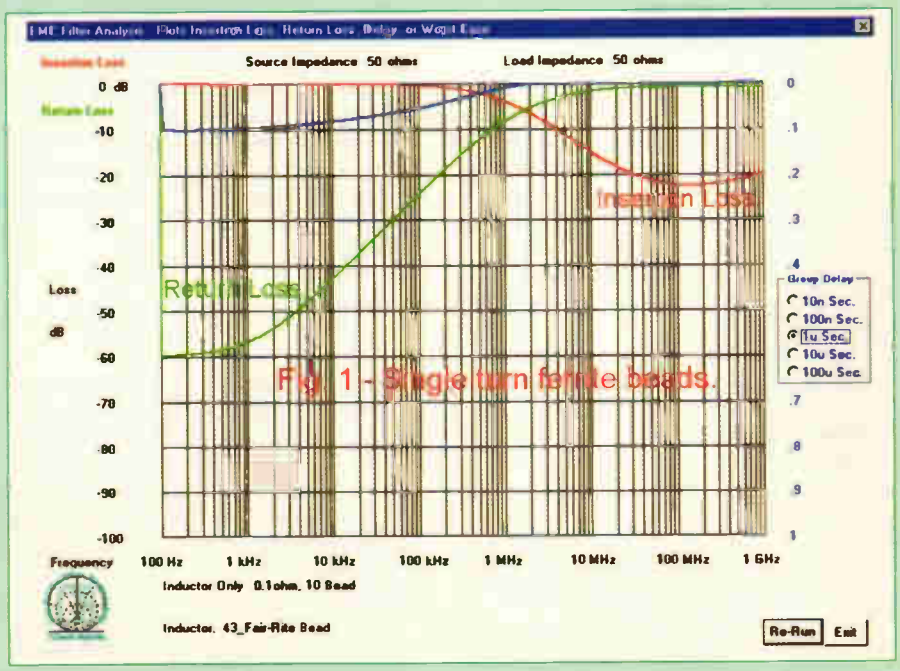
Regarding the naming of early transistors in your December issue letters, the 'Mullard Reference Manual of Transistor Circuits'.

EMC filters

The Robertson letter with the editor's heading (*EW* November) shows that both fail to properly understand EMC Filters and basic transmission line theory. Regardless of whether designed 'properly' or not, in order to provide attenuation every filter must provide a mismatch, so presents a high VSWR to the system. Let me explain. A perfectly matched attenuator having a VSWR of 1:1 and used in a matching system, dissipates energy to produce its attenuation, no energy being returned. However an attenuator having a non-ideal VSWR dissipates energy but also reflects some energy back to the source. That is exactly what its VSWR indicates. A VSWR of 2:1, often considered a reasonable match, whether for an EMC filter or an attenuator, indicates a return loss of 9.542dB and 33.33333% of the incident energy is reflected back to the source. In like fashion a VSWR of 3:1 represents a return loss of 6dB or 50.1% reflection. These VSWR results apply equally to resistive attenuators and filters, however in its stop band a filter has very much larger VSWR. A single turn winding, a ferrite bead threaded over a wire, behaves rather different over frequency, compared to a conventional inductor, in that while the material chosen exhibits inductance at lower frequencies, it becomes essentially resistive at high frequencies.

However as with the mis-matched attenuator, this resistive effect still represents a high VSWR so most of the incident energy continues to be reflected back to the source. It does of course also dissipate some energy, but it is the proportion of energy reflected to that dissipated which matters. Perhaps a quick simulation will clarify matters. In 1983 I wrote an EMC filter design simulation software, which was subsequently used to design EMC filters prior to manufacture, many thousands in all, while I was responsible for design and quality for the largest UK maker of professional or weapons systems EMC filters. These were 'type approved' filters, either under the BSI or UK MOD requirements.

Now aged 68 years, I no longer have access to that specific software or the Hewlett Packard workstation which ran it, however in 1996 for my own use, using Visual Basic I wrote a version for a PC with capacitor and inductor models more suited to my then needs. These models, which reflect the components measured characteristics by frequency, were based on R+jX values measured at a number of frequencies, using a MIL STD220 test rig. These 'spot' values are interpolated to provide the values used for each of the 140 frequencies shown in the plot. I append two plots: **Fig. 1.** is based on Fairite type 43 ferrite shield beads threaded onto a single wire turn, with DC resistance of 0.1Ω. The **Fig. 2.** plot uses a number of wire turns on a small MPP toroid to make a 15µH coil having the same DC resistance and similar 10MHz attenuation, for direct comparison. Both plots assume an exact 50Ω source and load and the calculation conforms with the measurement method set down in MIL-STD 220A. Each plot shows three traces,



Second Edition 1961 and 'Radio Valve Data', Sixth Edition c1959, show without doubt that the famous OC71 transistor was typeset with the letter 'O' and not the numeral '0'.

As the first publication was by the manufacturer and the second by *Wireless World*, I very much doubt that the first character was never a numeral even if a zero heater-voltage was to be indicated. Has someone been reinventing history?

Alan C Pickwick
Sale
Cheshire
UK

Oh, 'O'

If you stand to be corrected, Ed, sorry I mean Phil, regarding the transistor numbering typo referred to by J I Anderson in the December issue, then I'm sure you are not alone. Indeed, even Mullard were confused because their Maintenance Manual, 2nd edition of 1961, uses both the letter 'O' and the figure '0'; and their Reference Manual of Transistor Circuits, also 2nd edition of 1961, appears to use the letter 'O' exclusively.

Whilst we were familiar with valve numbering systems, probably most of us in those days were not aware of any meaning behind the '0C' numbering of transistors, and even as recently as 1977, T D Towers in his Towers' International Transistor Selector placed them immediately following the NKTs, rather than at the start of the numerical section. I think maybe that we were confused by the fact that many transistors of that era had a type number beginning with a variety of seemingly meaningless letters; other examples being AC, GET, XB and TS. Does anyone remember when 'red spot' and 'white spot' transistors were all you could buy at a reasonable price, or am I really showing my age now?

Different subject: were a diagram and a chunk of text omitted from Cyril Bateman's article in this issue? It doesn't seem to hang together as well as his articles do usually.

Mike Hall
Langport
Somerset
UK

Cyril has not come around and beaten us – so I can only assume all was as intended. And thanks for the OC support.

insertion loss dB (red), group delay µS (blue) and return loss dB (green). Naturally at low frequencies insertion loss starts at 0dB, while return loss starts at -60dB indicating 0.1% reflection due to the 0.1Ω DC resistance.

At 10MHz we see the toroidal inductor has low losses, so reflects 92% of the signal back to the source. The single turn ferrite beads are more resistive and more lossy, so reflect only 82% of the incident signal, which explains their temperature rise. Even more notable differences can be seen at the 3dB insertion loss frequency, the toroid insertion loss and return loss in dB are almost identical while that for the beads shows a larger difference with proportionally less energy being reflected.

However, while ferrite beads may reflect a smaller portion of the incident signal, they still reflect much more energy than they dissipate, thus confirming my original statement and hopefully clarifying Robertson's experience. Robertson also seems to assume a 'proper filter' uses only ferrite and no capacitors. Not so, while I too used an inductor to remove switching noise from my TanDelta meter design, in my professional life, every filter I designed used at least one capacitor, with or without a toroid or bead inductance. Domestic mains filters use large inductance values with small value capacitors, to reduce a shock hazard from capacitor current at the AC voltage. Wound inductors cost more than capacitors, so all lower voltage filters and especially those used in weapons systems, use relatively larger value capacitors and smaller value inductors. By weapons systems I mean filters used in anything which flies, aircraft both military and commercial, missiles or bombs and in land based vehicles. Anywhere size and weight is important.

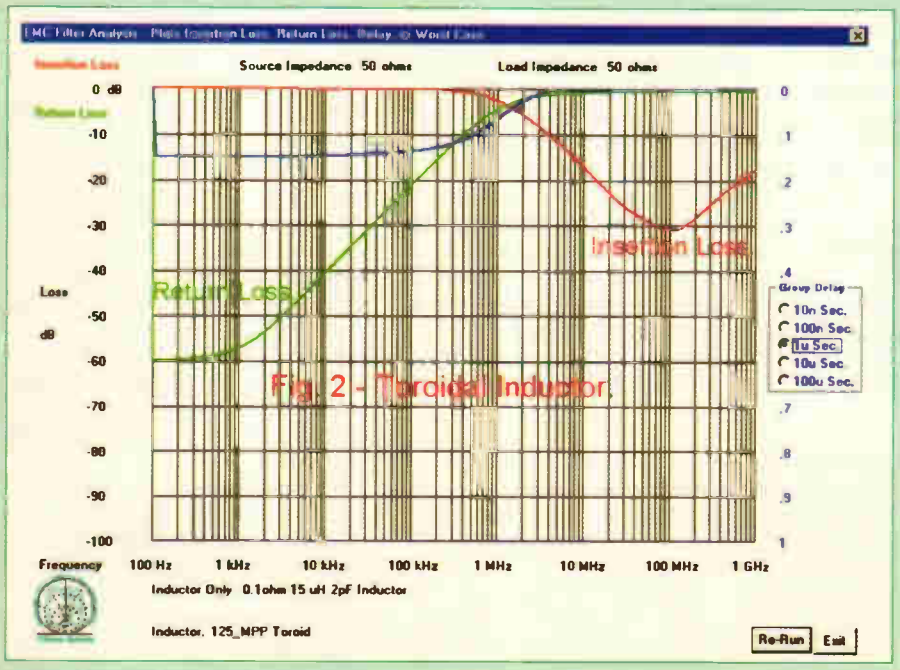


Fig. 2 - Toroidal Inductor

Not all such applications are aggressive, many such as used for ejector seats, self inflating dinghies, explosive bolts, cockpit canopies or drop tanks, are used to prevent accidental operation when an aircraft flies close to say a TV transmitter aerial. Other readers seem alarmed that the filter's reflected signal creates interference. Not so the filter doesn't create interference, if the source was free from interference then no interference would be reflected by the filter. Such EMC filter suppression systems do work because it is assumed the filter and the unwanted interference source are both within a 'dirty' box with a filter usually mounted in the box wall. Thus the area outside this 'dirty' box is kept clean and free from unwanted signals. This system of course also works in reverse, preventing outside interference from entering the 'box'.

Such methods work for signal lines as well as power lines. When measuring his washing machine, presumably Robertson opened the box to gain access, so broke this rule. I hope this lengthy response has served to answer other readers' queries, however should more information be required I can draft a suitable article.

Cyril Bateman
Acle
Norfolk
UK

Editor correct all along – horror!

I was very surprised to read J. I. Anderson's assertion (*EW* letters, December issue) that the 'O' in OA81, OC70, etc., should in fact be the figure zero. As far as I am aware, the first character in early (pre- 'Pro-Electron') Mullard semiconductor nomenclature has always been the capital letter 'O'.

Early transistors and diodes may well have been regarded as 'zero heater-voltage' valves. Indeed, the second letter of the early system of numbering seems to reflect Mullard/Philips (European) valve nomenclature: 'A' for a single diode (e.g. OA81) and 'C' for a (semiconductor) triode (e.g. OC70). Even the letter 'P' - used only as a third letter, and indicating a secondary emission valve - appears in the OCP71 phototransistor. The first digit seems related to the encapsulation and lead-out arrangement, just as in valve nomenclature.

However, in the Mullard 'Valve and Service Reference Manual' (Second edition, 1951) there is no 'first letter' given for cold cathode (zero heater-voltage) valves. Had there been, it presumably would have been a letter and not a digit. Clearly, 'O' would have been an appropriate choice, although in general, the letters 'O' and 'I' seem to have been avoided to prevent confusion with the figures '0' and '1'.

In the first edition of the 'Mullard Maintenance Manual' (published in the mid-1950s) the OA60, OA61 and OA70 follow the KLL32 and precede the PCC84. Supplements covering additional semiconductors, which were issued later, are numbered to fit into this part of the manual. Similarly, in the second edition (1961) of the Manual, 'OA' and 'OC' semiconductors follow 'N' and precede 'P'. And throughout the text of the Mullard 'Reference Manual of Transistor Circuits' (first edition, 1960), 'OA' and 'OC' devices are always identified as that, and never with a leading zero.

The fact that the first character is indeed a letter, is explicitly stated in the little Mullard 'Semiconductor Data Book 1966/67'. The 'old' system of numbering is described as: "The type nomenclature consists of two or three letters followed by a group of one, two or three figures".

As to why the first character might be interpreted as a zero, I can only speculate that it could have something to do with the numbering scheme devised by the Radio Manufacturers Association (RMA). Here, cold cathode tubes begin with the figure zero. I say 'tubes', as the scheme was associated with American valves. However, I believe it most unlikely that Mullard would have been influenced by this.

Philip Cadman
Dudley
West Midlands
UK

More 'O'

Yes, it is a business making sure there are no errors, but as your correspondent this month says, it is important. You will not thank me for pointing out that in the good old *WW* days there were few if any errors, but it is true. But there is no need to publish letters correcting errors once, let alone twice, even if they do go on to make complimentary remarks. A brief note mentioning the corrections should suffice.

I have been waiting to see if there is a correction to the suggested circuit in Fig.3 on page 34, August 2003. The circuitry around the second op-amp does not look to my untutored other transistors then in current production. If your correspondent looks carefully at the website page he refers to, he will see that it too uses the letter O in the prefixes OA and OC. If you are guilty of committing a typo, it is a typo hallowed by time, and a great deal of time relative to the short history of the transistor. It really was the Oh See series, not the Zero See series.

John Aysgarth
Durley
Southampton
UK

I quite realise that in the "good old WW days' there were less errors. Those 'good old days' are long gone, I'm afraid. There are all sorts of well-documented reasons why this magazine does not enjoy the circulation it once had. With a drop in circulation comes a drop in revenue and a consequent drop in staffing and alas, standards. We do our best with the electronic labour savers we have to hand, but nothing is infallible. – Ed.

Cyril Answer

I didn't see Cyril Bateman's problem in September but looking at the answer, my solution is to split it down a plane of symmetry - i.e. R3 & R5 - double them up for each half circuit and you can work out in you head, remembering to divide by two to bring the two halves back together. It appeared in Communications Quarterly several years ago to which I sent the above solution.

Stan Brown
Oswestry
Shropshire
UK

Over regulated?

I read your leader in the December edition of *EW* with interest. Some years ago I had a business manufacturing protocol converters. I've no idea what they radiated in the RF region but it was nothing compared to mobile phones. The main reason I ceased my business was the regulations coming out of Europe. The pages of *EW* told me about the legal action that would ensue if I sold equipment that had not been verified by a battery of expensive tests. I made bespoke modifications for customers and this would have required EMC tests each time, so I gracefully wound the whole thing up.

I compared your power feeder in Rhodes with the new French law about not smoking in public places. It seems that French restaurants have removed their no smoking areas because they are redundant. When I asked why the chap on the next table to me was smoking, I just got a shrug with hands held out palms upwards,

Wheatstone tolerances

Referring to the letter 'Meaningless Algebra' in the November issue of *EW*, I would query the statement that using 1% resistors in a Wheatstone bridge that you get 1% overall accuracy. Remembering that the balance conditions for a bridge are R1/R2 = R3/R4, then putting limit tolerances of 1% in the least favourable directions means that the overall accuracy tolerance is actually 3%, 1% being contributed by each of the three bridge resistors. Hence for a bridge to be accurate to 1% we need resistors in the bridge to have accuracies of 0.03% or better.

Arthur Bailey
Ilkley
West Yorkshire
UK

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the implication was that people will smoke wherever they wish. So this 'law' is useless too.
You are right, something is not making sense. And it won't be put right without some of us taking some sort of action.

Charles Coultas
Wokingham
Berkshire
UK

Living in London, as I do, I see laws broken every few seconds. Cars jumping red lights, cyclists ignoring any traffic regulations, littering etc. etc. I suspect though, that the cost of policing is deemed to be not worth it. It is much easier to police sitting ducks like electronic equipment manufacturers. Rather akin to shooting fish in a barrel. – Ed.

Skin effect

In regard to Mr Leslie Green's article on *Formula for Skin Effect* he notes the mistaken theory about skin effect in wires that it is widely taught in engineering schools. The incorrect explanation he claims is that it is due to internal inductance in wires. I checked my text books and found that I was one of the many who didn't get the proper understanding of the subject.

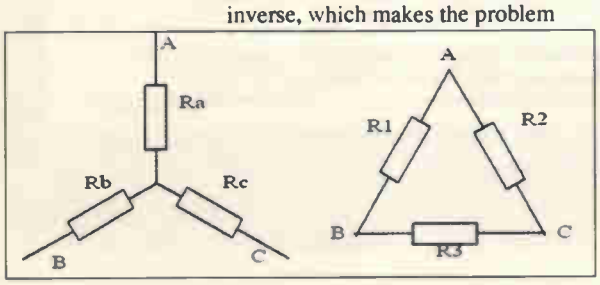
In light of the prevalent confusion would *Electronics World* and Mr Green consider publishing an article on the subject to clarify this important subject. And please don't spare the details. I for one want to know the whole truth about it.

George Woolcott
By email

Over to you, Mr. Green – Ed.

Cyril Bateman's cube of resistors problem.

I can also remember this problem first appearing in *Practical Television* many years ago, and at the time did not know how to solve it. Then when studying for my electronics degree in the late 1960s, was introduced to a method of network analysis called star to delta transform, and its inverse, which makes the problem possible even if the resistor values are different.



Cube of Resistors problem

When I saw this I thought there must be a way to solve it using the star to delta transform method. This would enable the problem to be solved no matter what value each resistor was, although the general solution gives rise to enormously complicated algebra.

For readers who don't know the star delta transform it is as shown. Taking external nodes as A, B, and C,

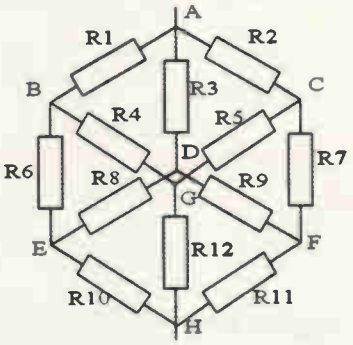
$$R1 = \frac{RaRc + RbRc + RaRb}{Rc}$$

R2 and R3 are similarly calculated. Going the other way from Delta to Star.

$$Ra = \frac{R1R2}{R1+R2+R3}$$

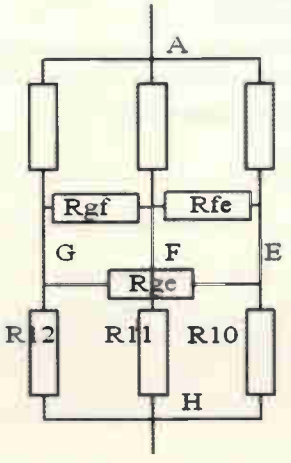
Working with the resistor cube it is possible to eliminate nodes B, C, and D by doing star delta transform on resistors R1,R4, R6 for node B; R2, R5, R7 for node C; R3, R8, R9, for node D.

This gives a pair of delta resistors



from A to each of the nodes E, F, and G which can be combined according to parallel resistors, and three resistors in a delta network between nodes E, F, and G.

If a Delta to star transform is done on these values to a new temporary node X, the remaining nodes E, F,

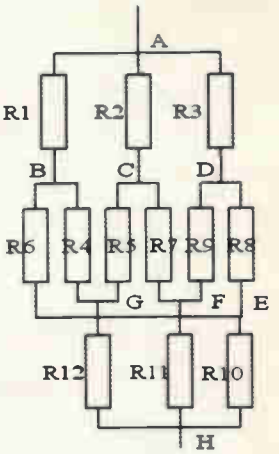


and G, can also be eliminated with a star to delta transform. This leaves three resistors in parallel between nodes A and X, A and H, X and H. which can easily be combined with series and parallel rules. However the general solution equations by this stage are horrendous.

If anyone knows a simpler way for the general solution I would be interested to hear, but I suspect that the ways 12 resistors in this topology combine is always going to lead to a complex general solution.

Looking at the flattened cube topology, just imagine that R1, R4,,R5, R7, R9, R8, R10 are all quite low values and the rest are quite high values, The predominant impedance would then be the former all in series, with small corrections due to high parallel impedances. The variations are endless.

I am a little surprised that no other



reader has suggested this, as it features in two of my textbooks from that time, but maybe the advent of high tech computer network analysis and simulation has meant that this is no longer taught as a method of network reduction. The algebra to create a general solution is rather complex, but if the values are known the intermediate stages can be calculated, and used in the next stage of calculation quite easily. It is unnecessarily complicated for the special case where the values are all the same, but does still work, and does not involve and joining of equipotential points to obtain a result. It simply relies on the fact that from outside a boundary, there is no way of telling any difference between different networks if the current and voltages at the accessible nodes are identical.

Raymond G. Lee
Gateshead
UK

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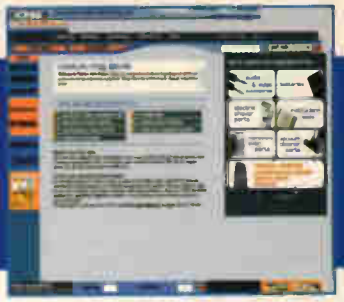

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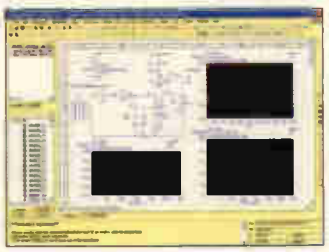

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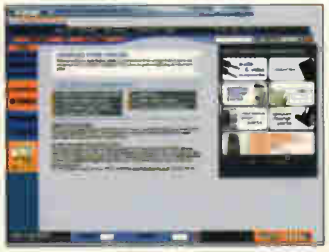

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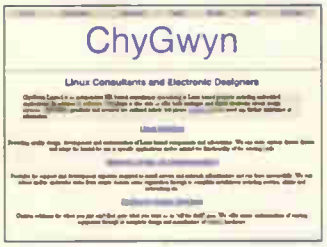

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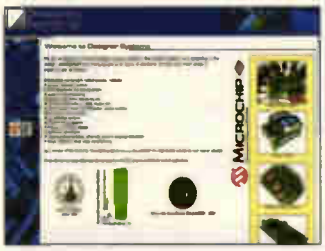

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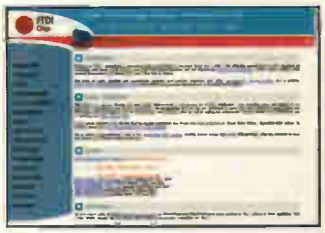

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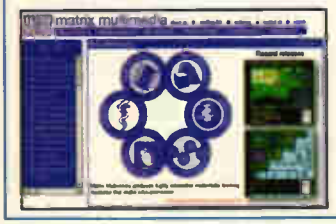

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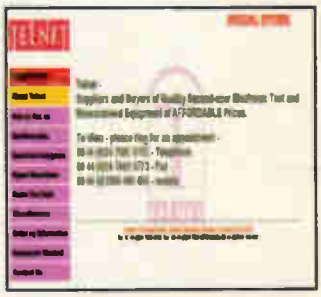

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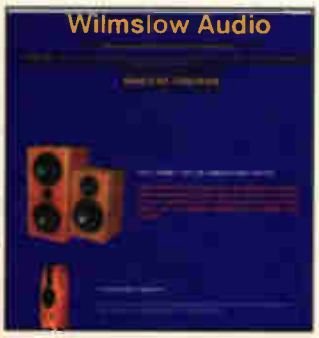

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RACAL 1772 COMMUNICATION RX – £400-£500.
TEK MODULE MAINFRAMES – TM501-503-504-506 TM5003-5006.
TEK P5010-M1 Logic Multimeter – 350. PG Prog 20MC/S Function Gen – £400. GPI Prog Scanner – £250 – DM Prog DMM – £400.
TEK 7000 OSCILLOSCOPE MAINFRAMES – 7603-7623-7633-7834-7854-7904-7904A-7104 – £150-£1,000.
TEK 7000 PI's 7A11-7A12-7A13-7A15-7A19-7A22-7A26-7A29-7A42-7B10-7B15-7B50A-7B80-7B85-7B92A-7D15-7D20.
TEK 7000 – 11[illegible]12[illegible] M+S1-S2-S3A-S4-S5-S6-S7-S53-S54.
HP POWER SUPPLIES – 6621A-6623A-6624A-6632A-6652A Qtys available. Also 6000 types CP[illegible]
WE KEEP IN STOCK HP and other makes of RF Frequency doublers which when fitted to the RF output socket of a Sweep generator doubles the output frequency EG.50-1300MC/S to 50-2600MC/S price from £250 – £450 each.

SPECTRUM ANALYZERS

HP 3580A 5Hz-50kHz – £250.
HP 3582A Dual 5Hz-25.5kHz – £1,500.
HP 3585A 20Hz-40MC/S – £3,500.
HP 3588A 10Hz-150MC/S – £7,500.
HP 8568A 100Hz-1.5GHz – £3,500.
HP 8569B 10Hz-22GHz – £6,500.
HP 8590B 9KC/S-1.8GHz – £4,500.
HP 8594E 10MC/S (0.01-22GHz) – £3,500.
HP 3581A Signal Analyzer 15Hz-50kHz – £400.
TEK492 50Hz-21GHz OPT 2 – £2,500.
TEK492P 50kHz-21GHz OPT 1-2-3 – £3,500.
TEK492AP 50kHz-21GHz OPT 1-2-3 – £4,000.
TEK492BP 50kHz-21GHz – £3,000-£4,000.
TEK495 100kHz-1.8GHz – £2,000.
HP 8557A 0.01MC/S-350MC/S – £500 + **MF180T** or **180C** – £150 –**182T** – £500.
HP 8558B 0.01-1500MC/S –£750 – **MF180T** or **180C** – £150 - **182T** – £500.
HP 8559A 0.01-21GHz – £1,000 – **MF18T** or **180C** – £150 –**182T** – £500.
HP 8901A AM FM Moduattion ANZ Meter – £800.
HP 8901B AM FM Modulation ANZ Meter – £1,750.
HP 8903A Audio Analyzer – £1,000.
HP 8903B Audio Analyzer £1,500.

MARCONI 2370 SPECTUM ANALYZERS – HIGH QUALITY – DIGITAL STORAGE – 30Hz-110MC/S Large qty to clear as received from Gov – all sold as is from pile complete or add £100 for basic testing and adjustment – callers preferred – pick your own from over sixty units – discount on qtys of five.

A EARLY MODEL GREY – horizontal alloy cooling fins – £200.
B LATE MODEL GREY – vertical alloy cooling fins – £300.
C LATE MODEL BROWN – as above (few only) – £500.

OSCILLOSCOPES

TEK 465-465B 100MC/S + 2 probes £250-£300.
TEK 466 100MC/S storage + 2 probes – £200.
TEK 475-475A 200MC/S 250MC/S + 2 probes – £300-£350.
TEK 2213-2213A-2215-2215A-2224-2225-2235-2236-2245- 60-100MC/S – £250-£400.
TEK 2445 4ch 150MC/S + 2 probes – £450.
TEK 2445A 4ch 150MC/S + 2 probes – £600.
TEK 2445B 4ch 150MC/S + 2 probes – £750.
TEK 468 D.S.O. 100MC/S + 2 probes – £500.
TEK 485 350MC/S 2 probes – £500.
TEK 2465 4ch - 300MC/S [illegible]50.
TEK 2465A 4ch 350MC/S [illegible]550
TEK D.S.O. 2430A 150MC/S + 2 probes – £1,750.
TEK D.S.O. 2440 300MC/S + 2 probes – £2,000.
TEK TAS 475-485 -100MC/S-20MC/S-4ch + 2 probes –£900-£1.1k.
HP1740A – 100MC/S + 2 probes – £[illegible]
HP1741A – 100MC/S storage + probes – £[illegible]
HP1720A – 1722A - 1724A - 1725A - 275MC/S + 2 probes – £[illegible]
HP1744A – 100MC/S storage large screen – £250.
HP1745A – 174[illegible] 100MC/S large screen – £350.
HP54100A – 1GHz digitizing – £500.
HP54230A – 500MC/S digitizing – £500.
HP1451A – 100MC/S digitizing – £500.
HP54100D – 1GHz digitizing – £1,000.

MICROWAVE COUNTERS – ALL LED READOUT

EIP 351D Counter 20Hz-18GHz – £750.
EIP 371 Micro Source Locking – 20Hz-18GHz – £850.
EIP 451 Micro Pulse Counter – 300MC/S-18GHz – £700.
EIP 545 Microwave Frequency Counter – 10Hz-18GHz – £1k.
EIP 548A Microwave Frequency Counter – 10Hz-26.5TGHz – £1.5k.
EIP 575 Microwave Source Locking – 10Hz-1.8GHz £1.2K.
EIP 588 Microwave Pulse Counter – 300MC/S-26.5GHz – £1.4K.
SD 6054B Micro Counter 20Hz-[illegible] SMA socket – £800.
SD 6054B Micro Counter 20Hz-18GHz N Socket – £700.
SD 6054D Micro Counter 20MC/S-18GHz – £600.
SD 6840A Micro Counter 20Hz-2.6GHz – £1.2K.
SD 6244A Micro Counter 20Hz-4.5GHz – £400.
HP5352B Micro Counter OPT 010-005-46GHz – new in box – £5K.
HP5340A Micro Counter 10Hz-18GHz – Nixey – £500.
HP5342A Micro Counter 10Hz-18GHz – £800-£1K – OPTS 001-002-003-[illegible]011 available.
HP5342A + 5344S Source Synchronizer – £1.5K.
HP5345A 500MC/S 11 Digit LED Readout – £400.
HP5345A +5354A Plugin – 4GHz – £700.
HP5345A + 5355A Plugin with 5356A 18GHz Head – £1K.
HP5385A 1GHz 5386A-5386A 3GHz Counter – £1K-£2K.
Racal/Dana Counter 1991-160MC/I – £200.
Racal/Dana Counter 1992-1.3GHz – £600.
Racal/Dana Counter 9921-3GHz – £300.

SIGNAL GENERATORS

HP8640A – AM-FM 0.5-512-102MC/S – £200-£400.
HP8640B – Phase locked – AM-FM-0.5-512-1024MC/S – £500-£1.2K. Opts 1-2-3 available.
HP8654A – B AM-FM 10MC/S-520MC/S – £300.
HP8656A SYN AM-FM 0.1-990MC/S – £900.
HP8656B SYN AM-FM 0.1-990MC/S – £1.5K.
HP8657A SYN AM-FM 0.1-1040MC/S – £2K.
HP8660C SYN AM-FM-PM-0.01-1300MC/S-2600MC/S – £2K.
HP8660D SYN AM-FM-PM-0.01-1300MC/S-2600MC/S– £3K.
HP8673D SYN AM-FM-PM-0.01-26.5 GHz – £12K.
HP3312A Function Generator AM-FM 13MC/S-Dual – £300.
HP3314A Function Generator AM-FM-VCO-20MC/S – £600.
HP3325A SYN Function Generator 21MC/S – £800
HP3326A SYN 2CH Function Generator 13MC/S-IEEE – £1.4K.
HP3336A-B-C SYN Func/Level Gen 21MC/S – £400-£300-£500.
Racal/Dana 9081 SYN S/G AM-FM-PH-5-520MC/S – £300.
Racal/Dana 9082 SYN S/G AM-FM-PH-1.5-520MC/S – £400.
Racal/Dana 9084 SYN S/G AM-FM-PH-.001-104MC/S – £300.

SPECIAL OFFERS

MARCONI 2019A SYNTHESIZED SIGNAL GENERATORS - 80KC/S-1040MC.S-AM-FM-£400 inc instruction book-tested.
MARCONI 2022E SYNTHESIZED SIGNAL GENERATOR - 10KC.S-1.01GHz AM-FM - £500 inc. instruction book - tested.
R&S APN 62 LF Sig Gen 0.1Hz-260 kHz c/w book - £250
MARCONI 2383 S.ANZ 100Hz - 4.2GHz. £2k
H.P. RF AMP 8349A 2.20 GHz microwave. £2k.
H.P. RF AMP 8347A 100 kHz - 3GHz £1.500.
H.P. 8922 radio communication test sets.
G - H - M. options various. £2,000 - £3,000 each.
H.P.4193A VECTOR IMPEDANCE METER + probe kit. 400 kHz. to 110 ML/S. £3,500.
H.P.83220A - E GMS UNITS for above. £1,000 - £1,500.
WAVETECK SCLUMBERGER 4031 RADIO COMMUNICATION TEST SET. Internal Spectrum ANZ. £1,800 - £2,000.
ANRITSU MS555A2 RADIO COMM ANZ. to 1000MC/S. No C.R. tube in this model. £450.
TEK 2445A - 4CH - 150MLS SCOPE. + New X1 + X10 - probe. Instruction book. £500 each.

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Amitsu – ME3620A SDH/SONET Analyser (Tx and Rx)	£7500
Amitsu – MD6430A Network Data Analyser	£9995
Amitsu – Emiscan EMCT Test Solution (Consists: MS2601B/ANT1/4401/MA8611)	£3500
Agilent – J2300E Advisor WAN Protocol Analyser	£5250
Agilent – 8753D (Opt 006) Network Analyser – Colour – 6GHz	£12000
Agilent – 54810A Infinium 500MHz – 2 channel	£3000
Agilent – 54610B 500MHz – 2 channel	£1250
Agilent – 3325B Synthesised Function Gen. (21 MHz)	£2000
Agilent – 3326A (opt 001) Synth. Function Gen. (13MHz)	£1500
Agilent – 35665A Dual Channel Dynamic Signal Analyser	£4000
Agilent – 71500A – 40GHz Microwave Transition Analyser (Inc 70004A/70820A)	£8000
Agilent – 37717C – Various options available – phone for details	from £1000
Agilent – 37718B – Comms. Perf. An. opts 1, 12, 106, 601, 602	£9750
Agilent – 8153A Lightwave Multimeter (less I/P Module) Modules available separately	£650
Agilent – 8163A Lightweight Multimeter	£500
Agilent – 8156A Dig. Optical Attenuator – various options	from £750
Agilent – 8164A Lightwave Measurement Mainframe	£2250
Agilent – 86120C Optical Wavelength Meter	from £1500
Agilent – 83480A Sampling Oscilloscope Mainframe (plus ins avail)	£2950
Burster – 1424 IEEE High Precision Decade Resistor	£1250
IFR 2031 2.7 GHz Signal Generator (opt 05)	£3500
IFR 2850 Digital Transmission Analyser	£900
IFR 2853 Digital Transmission Analyser (opts 14, 15, 25)	£1475
Tektronix CSA 8000 Comms. Sig. An. (ring for plug-ins available)	from £11000
Tektronix CSA 803C Communications Signal Analyser Mainframe	£3500
Tektronix MTS 100 MPEG Test System	£3000

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