Calibrating LF antennae

An electronic universe

Designing for EMC

Interfacing an AT keyboard to a PIC microcontroller

Circuit ideas:
Charger delay unit, Standalone button, latch Ozone generator, 5W Inverter
Radio Communication Test Sets

**Anritsu MT 8801C** 3000kHz - 3GHz (opt 1,4,7)
- Hewlett Packard 8920B (opt 1,4,7,11,12)
  - £3650
- Marconi 2955A
  - £1750
- Marconi 2955B/60B
  - £3500
- Marconi 2955S
  - £1995
- Ralac 6111 (GSM)
  - £6250
- Ralac 6115 (GSM)
  - £11250
- Rohde & Schwarz CMT 55 (2GHz)
  - £17950
- Rohde & Schwarz CMT 90 (2GHz) DECT
  - £3995
- Rohde & Schwarz CMTA 94 (GSM)
  - £29500
- Schlumberger Stabilock 4033
  - £2750
- Schlumberger Stabilock 4040
  - £1300
- Wavelet 4103 (GSM 900) Mobile phone tester
  - £1500
- Wavelet 4106 (GSM 900, 1800, 1900) Mobile phone tester
  - £2000

**MISCELLANEOUS**

Balcomm 1651A 100MHz Transconductance Amplifier
- Bias unit 5220 and 3225S, Cat.Call AVAILABLE if required.
  - £3400
- E500 Microwave Frequency Counter (16GHz)
  - £1800
- E54A4 B and 26.5GHz Frequency Counter
  - £1000
- E551 Source Locking Freq.Counter (1GHz)
  - £1500
- E556 Pulse Freq.Counter (1GHz)
  - £1200
- Gigahertz 8514C Power Meter + B300A Peak Power Sensor
  - £1750
- Gigahertz 8542C Dual Power Meter + 2 sensors 80041A
  - £3250
- Hewlett Packard 3395A Distortion measurement set
  - £750
- HP 3458A Dual power meter and sensor (various)
  - £750
- HP 3353A - synthesizer (200MHz-81MHz)
  - £1995
- HP 3457A 30MHz multi meter 0.5 digit
  - £850
- HP 3764A - Digital Transmission Analyser
  - £2750
- HP 37690D - Signalising test set
  - £2250
- HP 4278A LC2 Meter (100MHz-20GHz)
  - £1400
- HP 5324A Microwave Freq.Counter (16GHz)
  - £350
- HP 5326B 20GHz Microwave Freq.Counter
  - £2000
- HP 5326B (1 & 2) Microwave Freq.Counter (26.5GHz)
  - £350
- HP 5385A - 1 GHz Frequency counter
  - £485
- HP 6022A - Autonumbering System PSU (20v-30a)
  - £750
- HP 6032A - Dual OP system p.a
  - £1250
- HP 6593A - Dual Output Power Supply
  - £2750
- HP 6592B - Dual OP Power Supply
  - £1950
- HP 6592A - Quad OP Power Supply
  - £1450
- HP 6592B - System Power Supply (20v-35a)
  - £1950
- HP 8360B - Sweep Generator Mainframe
  - £1500
- HP 8660A, B and E - Distortion Analyser
  - £1000
- HP 8674A - High performance RF synthesizer (0.1-1050MHz)
  - £350
- HP 8656A - Synthesised signal generator
  - £750
- HP 8656B - Synthesised signal generator
  - £500
- HP 8657A - Synth. signal gen. (0.1-1040MHz)
  - £1500
- HP 8657B - 100MHz Syn. Gen. - 2000 MHz
  - £2500
- HP 8657D - XX-DOP/5k Sig Gen
  - £350
- HP 8691B - Modulation Generator
  - £1500
- HP 81175B/C Carrier Noise Test Set
  - £2500
- HP 83131A Universal Frequency counter (30GHz)
  - £350
- HP 83151B Microwave Freq.Counter (26.5GHz)
  - £350
- HP 83331 - Frequency Counter
  - £1000
- Keithley 237 High Voltage - Source Measure Unit
  - £1450
- Keithley 238 High Current - Source Measure Unit
  - £4500
- Keithley 6460/70 Preamplifier (volk source)
  - £13500
- Keithley 8008 Component test fixture
  - £1750
- Marconi 2804A 2 Mols Transmission Analyser
  - £1100
- Marconi 9905S/990/9990B Power Meters & Sensors
  - £3000
- Philco 895 - TN - Code 7: A9, 2m power meter generator
  - £1400
- Philips PM 515 - 50M Function Generator
  - £1500
- Leader 1311 Signal generator 100MHz - 1400MHz - AM/AM/AMCW with built in FM stereo modulator (as new) at £2500
- Rohde & Schwarz FAM 6 and 8 A & B Modulation Analysers
  - £850
- Rohde & Schwarz NVR dual channel power meter & NAV 22 Sensor
  - £1000
- Tektronix TDS 650G - Audio Signal Generator
  - £750
- Wavelet 176 Function generator (50MHz)
  - £750
- Wayne Kerr 2245 - Precision Inductance Analyser
  - £450
- Wayne Kerr 3250A - 3250A Precision Magnetics Analyser with Bias Unit
  - £2500
- Wayne Kerr 8245 - Precision Component Analyser
  - £2250
COMMENT
Reasons to be cheerful...

NEWS
- Half a million on broadband
- Carbon in missing link
- Green power gets go-ahead
- Zetex moves to p-channel
- Hot air sensor
- Power dressing Scots
- Jelly foils fingerprint checks
- New life for old filaments
- HD hits 300 Gb
- Flash dual bit memory
- Governemt pushes RF tags

CALIBRATING LF ANTENNAE USING DCF39
Paolo Antoniazzi and Marco Arecco give us various designs for LF antennae and show how to calibrate them by using a broadcast transmitter.

DESIGNING FOR EMC
Judging by some of our letters, Ian Darney is set to fuel another interesting discussion, this time about grounding in the context of EMC. I'm already looking forward to the mailbag.

KEY FACTORS IN RF POWER AMP DESIGN
Stephan weber thinks that there are some situationswhere a discrete component solution fits the bill. But this route is not without its pitfalls.

LETTERS
- Star grounding
- Super regen
- 500 MHz sampling front end
- More PCBs

CIRCUIT IDEAS
- Ozoniser
- Valve portable PSU
- 5W inverter
- Standalone button latch
- Battery charger timer

NEW PRODUCTS
The month's top new products.

AN ELECTRONIC UNIVERSE
Nigel Cook gives us his interesting standpoint on some well-established theories. Look out Mr. Ohm.

KEYBOARD INPUT FOR PIC PROJECTS
One of the problems with PIC projects is data input. Roger Thomas thinks he has a solution in the form of keyboard input.

WEB DIRECTIONS
Useful web addresses for electronics engineers.

September issue on sale 1 August
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- **Order Code**
  - 48V: FR710

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- **Order Code**
  - 48V: FR720
Changing times

Welcome to the August issue of Electronic World and let me introduce myself as your new editor. My name is Phil Reed and I'll tell you a bit about myself later on in this leader. But firstly, I'd like to thank Martin Eccles for many years of superb editorship of this respected journal and I can only hope that I come up to the high standards he has already set.

So, who on earth is Phil Reed? Well, I am an engineer by trade, having worked in the broadcast industry for the last 32 years. Whilst I have rarely had to pay the mortgage by designing electronics - I do understand most of what goes on in these pages - and have certainly had to fix some of the circuitry designed by some EW readers! And it was only a couple of weeks ago that my soldering prowess was earning me a crust (and a burnt thumb). My career has taken me to all corners of the broadcasting world, from acquisition to post production and even touching upon delivery technologies, stopping short of actually working on a transmitter station. I am not new to scribbling for a living, either. I have written regular columns in the broadcast trade press and my journalistic career reached new heights when I was editing the esteemed 'International Broadcast Engineer' magazine. But I have decided that I needed to get back to my roots and have do some proper engineering. In my spare time I'm engineering for a London based post production company, building and looking after many video editing suites and sorting out all manner of technical problems with a popular 'reality TV' series, based in Elstree film studios.

I used to be an avid reader of EW's predecessor, Wireless World, for many years and it has been an eye-opener to me to see how the design industry has moved on in the intervening 20 years or so! I am quite thrilled to be involved in this side of the business and look forward to be able to serve the readership with some ideas of my own. As with all things technical, the industry is changing rapidly – only a few years ago the things that you can do with PCs now would have seemed impossible. The same thing goes for DSP chips whose power to do ridiculously clever things in a cheap mass produced package is legendary and I hope to reflect some of these profound changes in these pages in the future.

As you can imagine, there are lots of boxes of article and circuit ideas that I've inherited – and it's going to take me some while to go through them all, so if you were expecting a reply about any submissions you've made – it might be an idea to send me an email to remind me. But do keep the circuit ideas and article submissions rolling in.

Over the next few months I will start the process of making some subtle changes to EW, nothing major you understand, just some small adjustments spurred on by feedback from you, which came from our 2002 reader survey. It appears that most of you (70%) are electronics professionals, 31% of you spend over £200 on components each month and 71% of you have a PC with internet access. So, armed with all this info, I'll be tweaking the content to suit. Suffice to say, though, that any comments are always welcome (even negative ones) and the best ones will be published. Editorial comments should be sent to me directly at p.reed@highburybiz.com.
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Half a million on broadband

Over half a million broadband connections have been set up in the UK, claims telecoms watchdog Oftel. “With over 20,000 broadband connections a week, the current level of growth outstrips the equivalent demand for mobile phones and dial-up Internet when they were first introduced,” said David Edmonds, Oftel’s director general of telecoms. The figures include all four main access technologies; cable modems, DSL technology, broadband fixed wireless and broadband satellite services.

The lure of broadband access will continue, Edmonds said: “Over 10 million homes use the traditional dial-up Internet access, including four million with unmetered packages. “I am confident that more Internet users will take up high speed broadband as the range of services increases and prices fall.”

Douglas Alexander, the Government’s e-commerce minister, said: “The milestone of half a million connections represents a 54 per cent increase since the beginning of 2002. Of course there is more to do, but the work of building Broadband Britain is under way.”

Carbon in missing link

The continued research into carbon nanotubes continues with IBM of the US and Infineon Technologies of Germany pushing the integration of nanotubes with silicon.

IBM has taken a major step towards transistors and ICs made from carbon nanotubes by proving that devices can outperform silicon transistors.

Researchers at the firm created a prototype nanotube transistors with twice the transconductance of the best prototype silicon devices, IBM said. “Proving that carbon nanotubes outperform silicon transistors opens the door for more research related to the commercial viability of nanotubes,” said Dr Phaedon Avouris, manager of nanoscale science at IBM Research.

Avouris’ team used single walled nanotubes (SWNTs) in a conventional Mosfet-like structure, with the nanotube forming the channel between the source and drain.

However, the gate dielectric was thicker than a Mosfet, at 10 to 15nm, even at gate voltages of 1V. Transconductance of 2,300µS/µm is more than double that of a 15nm length Mosfet with a 1.4nm gate oxide.

IBM was also able to make both p- and n-type nanotube Fets. Meanwhile Infineon has managed the controlled placement of nanotubes on standard 150nm silicon wafers.

The firm sees nanotubes replacing both the Fets and the interconnect in integrated circuits. Nanotubes allow current densities up to 10[super]10A/cm[super]2, three orders of magnitude higher than copper can manage. Interconnect in conventional silicon chips is expected to reach its thermal limits in around ten years’ time.

Finally, a group of researchers from the UK, France and the US have shown carbon nanotubes can ignite after exposure to a photographic flash.

A flash gun with more than 100mW/cm[super]2 of light power is enough to ignite SWNTs, which reach temperatures of at least 1,500°C, said the team.

The light leads to a photoacoustic effect caused by the expansion and contraction of trapped gasses. The high thermal conductivity of nanotubes helps propagate heat through a bundle.

The first US airborne laser missile-defence aircraft, a modified Boeing 747-400 freighter, is being prepared for flight testing later this summer. Flight-worthiness testing will be followed by a trip to Edwards Air Force Base in California where the laser and optics will be fitted.
Green power gets go-ahead

The Department of Trade and Industry has rubber stamped plans for the country's largest wind farm at Cefen Coes, near Aberystwyth.

With 39 turbines, the £35m project will be one of the largest of its type in Europe, said the Renewable Development Company, which is backing the project.

The scheme is part of the Government’s plan to supply ten per cent of the UK’s energy needs through renewable sources by 2010.

It is hoped that the wind farm will provide up to half of the local area’s demand for electricity, and a full one per cent of Wales’ total generation capacity.

However, the size of the scheme meant it bypassed the Welsh Assembly and went straight to the DTI in London for approval, a move that angered many activists in the West Wales area.

Energy Minister Brian Wilson has also unveiled a £2.3m plan for off-shore wave energy systems. The development and demonstration systems will be installed off the Western Isles.

Cash for this scheme comes from the £100m fund set up by the Government last year.

Three devices, located in shallow water, will generate power based on the oscillating water column principle. These techniques have already been used closer on-shore.

Zetex moves to p-channel

Analogue chip specialist Zetex has developed a p-channel Mosfet using its trench semiconductor process.

Zetex licensed techniques from an unnamed company that allow the Fets to be made without any critical alignment steps.

"P-channel Mosfets are tricky to make," said company product development manager Peter Blair.

Swapping materials in a existing n-channel design is not the answer, "there are additional challenges", he said.

The photo shows the device mid-process, with two and a bit recessed polysilicon gates in trenches. Oxide will back-fill the trenches to make a planar surface for metalisation after sources are implanted in the mesa sides.

The oxide layer on the mesa tops is sacrificial and will be removed before metal deposition.

The first devices made using the p-channel Fets will be a 40V, 70mΩ SOT223 for digital audio.
Dual bit memory is very flash

A new flash memory cell that stores two bits per cell without using multi-level techniques has been announced.

AMD calls the technology MirrorBit and partner Fujitsu calls it MirrorFlash.

There are two main differences between MirrorBit and normal flash: the transistor is symmetrical in MirrorBit and the floating gate in which data is stored is insulating silicon nitride, not the usual conductive polysilicon.

The new floating gate is the critical element as, being insulating, is can store regions of different charge.

In a normal floating gate injected electrons swim about as they wish. In an insulating gate electrons "are injected into traps in the nitride", said Joe Raushmayer, v-p of engineering at AMD.

Trapping allows electrons that make up one bit of data to be stored at one end of the gate while the second bit resides at the other end. Being symmetrical, the underlying transistor allows both ends of the floating gate to be treated equally. Reading and writing the bits involves manipulating the two transistor electrodes appropriately.

Erasure is performed like a normal flash memory. The main gate is set negative and the transistor electrodes and its substrate are set positive. This forces the trapped electrons out of the storage structure erasing both bits.

Government pushes RF tags

Major UK firms have signed up to a Home Office initiative to add radio frequency identification (RFID) tags to consumer goods.

Woolworths, Dell, EMI and Asda are part of the scheme, which aims to stamp out the trade in stolen and counterfeit goods. Items tagged will include CDs, laptop PCs and clothing.

Goods will be fitted with a unique tag that stores information such as their origin, current location and final retail destination.

"As criminals are using increasingly sophisticated methods so we must harness the latest technology available to us if we are to catch them," said Crime Reduction Minister John Denham.

The Government is putting £5.5m into its Chipping of Goods initiative. It has already tested the system on mobile phones, watches, alcohol and boats.

Made by Bedfordshire-based INSYS, this will be the last thing to touch UK satellite Beagle 2 before it rendezvous with the Red Planet. Called the spin-up and ejection mechanism (SUEM), it has just passed qualification testing at Astrium in Stevenage. The SUEM will hold the satellite in place on its rocket during launch and on the six month cruise to Mars.

US firm Digit Wireless has come up with a novel method of adding characters to a standard mobile phone keypad. Raised letters are placed inbetween the number pads while software comes with letters being pressed on the way to a number. The firm said the design should dramatically increase text entry speeds, and make it easier for partially sighted users.
Sensor is all hot air

US firm Memsic has developed a two-axis accelerometer that uses a bubble of hot gas as the proof mass. The Massachusetts-based firm is selling its hot gas accelerometers in 5x5x2mm surface mount packages. Using a bubble of gas brings two immediate benefits - high shock resistance and low noise.

“There are no moving parts except air. It will survive 50,000g,” claimed Mike Higgins, marketing and sales manager at Memsic, “where g is acceleration due to gravity (9.8ms-2), not grams.”

This seems like overkill for any imaginable application, but Higgins sees it as a safety margin above normal production processes. “Snapping a circuit board out can produce 3,000g,” he said.

Noise is particularly low and was recently halved by changing the working gas. “We can resolve very small g-forces: better than 1mg,” said Higgins. Over frequency he claims 0.2mg/√Hz on some variants.

Accuracy in the devices, which range from 1 to 10g full-scale with options to 100g, is 0.2 per cent typical, 0.4 per cent max. Due to the tiny amount of air involved, response time is small - 40ms and 120ms worst-case claims Higgins.

“So what are the disadvantages of thermal accelerometers?”

“Dependence on temperature. The sensitivity changes and this has to be compensated externally,” said Higgins. Although he points out that the compensation curve does not vary between devices as it derives from the gas law.

A datasheet and application note including compensation circuits is available from the company website.

Power consumption small - 3.6mA at 5V - and can be cut by pulsing, but may be enough to deter use in some battery powered applications.

As noise is so low, well under 1° of tilt can be measured, the accelerometers could be used to control cursors in portable devices where tilting the device moves the cursor or view. Car alarms, rollover detectors and navigation are all being considered as well.

www.memsic.com

How it works

In principle, the hot air accelerometer is simple. Hot air is less dense than cold air.

If they co-exist in a sealed environment and the environment is accelerated the hot air gets displaced in the direction of acceleration.

A similar effect can be seen if a toy helium balloon is let loose in a car. Accelerate the car and the balloon moves towards the windscreen. Brake and it moves towards the boot.

Memsic devices work in two-dimensions. The gas is held in a domed void with a flat silicon bottom within the chip packaging. In the centre of the silicon is a heater. This maintains the hot air “bubble” as Memsic’s Higgins describes it. “[Silicon] thermopile sensor under the bubble detect the way it moves,” he said.

The chip, which includes conditioning circuitry and is made by TSMC, is standard CMOS except that the heater trench is added post-foundry by Memsic in its own Chinese plant.
Scots go for power dressing

Practical power generating fabrics are possible, is the conclusion of a research project at Herriot-Watt University in Edinburgh, although the team has not actually made any yet.

"We can see several ways to put silicon photo-voltaics directly onto fabrics without a glass substrate," said Professor John Wilson of the university.

What the team has done is to make photo-sensitive cloth and prove that photo-coated cloth can be stable, flexible and reasonably durable.

Polymer and similar organic photo-semiconductors may in future be ideal for photo-cloth, but were rejected from the project as they are too immature. Instead thin-film silicon was chosen and has been coated onto both woven and non-woven (felt-like) materials.

To make a cloth photosensor, silicon layers and electrodes are plasma-coated onto the fabric over a sealing layer.

The result is a cell which follows the contours of the fabric strands and is flexible. "The cell is unlikely to be the problem," said Wilson. "Reliable connections between cells are more difficult."

Photo-clothing is far into the future. Wilson sees photo-voltaic tarpaulins and tentage as initial applications. "A roll-up canvas photo cell would be much easier to transport over rough roads than a glass one."

Finding a large-scale roll-to-roll plasma coating processes should not be a problem for production, as these are currently under development for a number of markets and, said Wilson, some carpets are currently being coated using a related high-tech process.

Herriot-Watt is seeking partners and funding for the next project phase.

Jelly foils fingerprint checks

A Japanese mathematician has broken the security on 11 fingerprint sensors by copying fingerprint patterns using cheap kitchen ingredients such as gelatine.

Tsutomu Matsumoto, from the graduate school of environment and information sciences at Yokohama National University, can fool fingerprint detectors 80 per cent of the time with his jelly-mould fingers.

His technique is to take an impression of a finger in a plastic mould, easily available in hobby shops, and then pour in liquid gelatine, which sets to form the fake finger. From start to finish the whole process takes less than one hour.

Fingerprint sensors can usually detect when a silicone prosthetic is used, but Matsumoto’s use of gelatine deceives the technology. He can also fool sensors that claim to detect only ‘live’ fingers, by moistening the gelatine before pressing onto the sensor.

In a presentation to the International Telecommunications Union’s workshop on security, Matsumoto said: "The experimental study on the dummy fingers will have considerable impact on security assessment of fingerprint systems."

More significantly, Matsumoto is able to copy prints made on surfaces such as glass. The process involves fixing and enhancing the print with cyano-acrylate (superglue) fumes and photographing it, exactly as forensic scientists would do. The image is enhanced in a software package such as Photoshop and then copied onto a blank copper PCB. The print is then etched and pressed into a mould ready for the gelatine.

Whether copying fingers direct, or reproducing them from prints on glass, Matsumoto was able to break 11 commercially available sensing systems. These included optical and capacitive systems.

In his conclusions, Matsumoto pointed out that manufacturers and users of biometric systems should carefully check their security against artificial clones.
New life for old filaments

Good old tungsten-filament bulbs, currently left behind in the efficiency stakes, could catch up through a development at Sandia.

So far the experiments have not been extended to visible light. Instead a filament below dull red heat that would normally emit mostly medium-wave infra-red has been made to emit much more short wavelength infra-red.

"Energy was being preferentially absorbed into a selected frequency band. Meanwhile periodic metallic-air boundaries led to an extraordinarily large transmission enhancement. Experimental results showed that a large photonic band gap for wavelengths from 8 to 20 microns proved ideally suited for suppressing broadband blackbody radiation in the infrared and has the potential to redirect thermal excitation energy into the visible spectrum," said Sandia.

Could it work at visible frequencies?

"The work was performed with a photonic crystal operating in the mid-infrared range," said the lab, "but no theoretical or practical difficulties are known to exist to downsizing the structure into the visible light range."

Hard drive hits 300Gbit/in [super2]

Fujitsu is claiming to be able to achieve a record hard disc drive density of 300Gbit/in² after developing a new read head and a new magnetic material.

"The new technologies are expected to lead to the commercial introduction within two to four years of 2.5 inch hard disc drives with capacities up to six times the recording density available today," said the company.

Current-perpendicular-to-plane mode is used in the new giant magneto-resistive (GMR) heads. These are credited with three times the playback output levels of existing hard drive heads which operate in current-in-plane mode are considered to have a limit of approximately 100Gbit², said Fujitsu.

Fujitsu engineers have developed a synthetic ferromagnetic media that can handle one million flux changes per inch to surface its proposed discs.

Within four years, Fujitsu claims it is likely to be making 360Gbyte hard drives.
Enhanced ‘PICALL’ ISP PIC Programmer

Kit will program virtually ALL 8 to 40 pin* serial and parallel programmed PIC microcontrollers. Connects to PC parallel port. Supplied with fully functional pre-registered PICALL DOS and WINDOWS AVR Software packages, all components and high quality DSPTH board. Also programs certain ATMEG AVR, SCENIX SX and EEPROM 24C devices. New devices can be added to the software as they are released. Blank chip auto detect feature for super-fast bulk programming. Hardware now supports ISP programming. *A 40 pin wide 2IF socket is required to program 0.3" devices (Order Code AZF40 @ £15.00).

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<td>A5314F2</td>
<td>Assembled Enhanced PICALL ISP Programmer</td>
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ATMEL 89xxxx Programmer

Powerful programmer for Atmel 8051 microcontroller family. All fuse and lock bits are programmable. Connects to serial port. Can be used with any computer and operating system. 4 LEDs indicate programming status and all programs 89C1051, 89C2051, 89C4051, 89C51, 89L051, 89L52, 89LV52, 89C55, 89LV55, 89S8252, 89LS8252, 89S53 & 89LS53 devices. NO software needed — uses any terminal emulator program (built into Windows).

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Atmel 89C051 and AVR programmers also available.

PC Data Acquisition & Control Unit

Use a PC parallel port as a real world interface. Unit can be connected to a mixture of analogue and digital inputs from pressure, temperature, humidity, sound, weight sensors, etc. (not supplied) to sensing switch and relay states. It can then process the input data and use the information to control up to 11 physical devices such as motors, sirens, other relays, servo motors & two-stepper motors.

FEATURES:
- 8 digital Outputs: Open collector, 500mA, 33V max
- 16 Digital Inputs: 0V max.Protect 1K in series, 5.1V Zener to ground.
- 11 analogue Inputs: 0-5V, 10 bits (5mV/step)
- 1 analogue Outputs: 2.05V or 0-10V. 8 bit (20mV/step.)

All components provided including a plastic case (140mm x 110mm x 35mm) with pre-punched and silk screened front/rear panels to give a professional and attractive finish (see photo). With screen printed front and rear panels supplied. Software utilities & programming examples supplied.

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<td>A53093</td>
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ABC Mini ‘Hotchip’ Board

Currently learning about microcontrollers? Need to do something more than flash a LED or sound buzzer? The ABC Mini ‘Hotchip’ Board is based on Atmel’s AVR 8535, RISC technology and will interest both the beginner and expert alike. Beginners will find that they can write and test a simple program, using the BASIC programming language, within an hour or two of connecting it up. Experts will like the power and flexibility of the Atmel microcontroller, as well as the ease with which the little Hot Chip board can be “designed-in” to a project. The ABC Mini Board ‘Starter Pack’ includes just about everything you need to get up and experimenting right away. On the hardware side, there’s a pre-assembled micro controller PC board with both parallel and serial cables for connection to your PC. Windows software included on CD-ROM features an Assembler, BASIC compiler and in-system program. The pre-assembled boards only are also available separately.

Order Ref | Description                        | Inc. VAT ea |
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<td>ABC MINI Starter Pack</td>
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<td>ABCMINIB</td>
<td>ABC MINI Board Only</td>
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Advanced 32-bit Schematic Capture and Simulation Design Studio

Advanced Schematic Capture & Simulation Software

- Streamline your design process with VisualSpice.
- Built-in 64 Channel Real-Time Virtual Oscilloscope.
- Built-in Digital Logic and PLC Trainer allow you to simulate ladder logic and PLCs.
- All different styles including essential Monte CarloNC
- See web site for full details and demo.

Serial Port Isolated I/O Controller

Kit provides eight relay outputs capable of switching 5Amp max and four optically isolated inputs. Can be used in a variety of control and sensing applications including load switching, external switch input sensing, contact closure and external voltage sensing. Programmed via a computer serial port, it is compatible with ANY computer & operating system. After programming, PC can be disconnected. Serial cables can be up to 35m long, allowing 'remote' control. User can easily write batch file programs to control the kit using simple text commands. NO special software required — uses any terminal emulator program (built into Windows). Screw terminal block connections. All components provided including a plastic case with pre-punched and silk screened front/rear panels to give a professional and attractive finish (see photo).

Order Ref | Description                        | Inc. VAT ea |
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Full details of these items and over 200 other projects can be found at www.QuasarElectronics.com
Calibrating LF antennae using DCF39

Our first attempt at making an LF loop antenna was disastrous. After months of study, measurements and discussion though, we are now true supporters of the loop antenna for receiving LF signals.

A simple loop with 38 turns at about 80cm diameter is a good competitor for a vertical rod and a 2m diameter loop will result in a superb antenna — the equivalent of 20 to 50m height at 136kHz.

An important question is how to make reliable measurements of the performance of loop antennas and other similar configurations. Here we propose a solution to the problem using the high-powered DCF39 station in Germany in conjunction with a small and simple reference loop.

Bear in mind that a loop antenna that performs wonderfully when receiving signals will not necessarily achieve the wonderful performance when transmitting.

Loop antennas for 136kHz

A loop antenna comprises a large coil wound on a suitable isolated support with an appropriate base. The main advantages of the loop used as an LF receiving antenna are

- directivity and narrow band if tuned
- less sensitivity to local electric noises
- smaller dimensions relative to an equivalent vertical rod
- easy to build.

The antenna works by taking energy from the incoming wave, due to the phase differences between the voltages induced in the two vertical opposite sides. When the plane of the loop is perpendicular to the direction of the propagation wave, no voltage results at the aerial terminals. In contrast, when the loop antenna's plane is parallel to the incoming wave, the voltage across the antenna reaches the maximum value.

The directivity of a loop is about 90° in the front and at the back (−3dB perpendicular to the antenna plane) This is certainly an advantage in comparison to a vertical rod because it prevents unwanted signals coming from different paths, Fig. 1.

The following relation describes the voltage across a loop receiving aerial submitted to an electric field.

\[
V = \frac{2\pi ENA \cos \theta}{\lambda} = Eh \cos \theta
\]

where:

- \( V \) = voltage at the ends of the loop (mV)
- \( E \) = electric field (nV/m)
- \( N \) = number of turns of the loop
- \( A \) = average turn area (m^2)
- \( \lambda \) = wavelength (m)
- \( \theta \) = angle between loop plane and the arriving wave: if the angle is 0°, \( \cos \theta = 1 \) and this term disappears
- \( h_a \) = antenna equivalent height (m)

This equation is applicable to any loop shape provided that the antenna's dimensions are small compared with the wavelength — i.e. less than approximately 0.1\( \lambda \). In the low-frequency range, it is very easy to satisfy this requirement.

You can tune the loop by placing a variable capacitor across the antenna terminals. This cause a larger voltage to appear at the balanced preamplifier inputs because of the Q of the parallel-resonant circuit.
Table 1. Tuned loops comparison at 136kHz.

<table>
<thead>
<tr>
<th>#</th>
<th>Turns (N)</th>
<th>Dia (m)</th>
<th>A (m²)</th>
<th>Total Wire Length (m)</th>
<th>N x A (m²)</th>
<th>Q Unloaded</th>
<th>Induct. (µH)</th>
<th>Tuning Cap. (pF)</th>
<th>Equiv. height h₀ (m)</th>
<th>Notes</th>
</tr>
</thead>
<tbody>
<tr>
<td>Loop 18M008</td>
<td>18</td>
<td>0.31</td>
<td>0.0754</td>
<td>17.5</td>
<td>1.36</td>
<td>200</td>
<td>148</td>
<td>8200</td>
<td>0.774</td>
<td>Plastic covered 1.8 mm diameter wires Moplen support</td>
</tr>
<tr>
<td>Loop 38M047</td>
<td>38</td>
<td>0.77</td>
<td>0.470</td>
<td>92</td>
<td>17.9</td>
<td>210</td>
<td>1700</td>
<td>806</td>
<td>9.17</td>
<td>Plastic covered 1.8 mm diameter wires Wood support</td>
</tr>
<tr>
<td>G3LNP (*)</td>
<td>54</td>
<td>0.90</td>
<td>0.640</td>
<td>153</td>
<td>34.6</td>
<td>70</td>
<td>4320</td>
<td>318</td>
<td>6.90</td>
<td>Litz Wires Wood support</td>
</tr>
<tr>
<td>Loop 18M2 (**)</td>
<td>18</td>
<td>1.60</td>
<td>2.00</td>
<td>91</td>
<td>36</td>
<td>200</td>
<td>1032</td>
<td>1327</td>
<td>20.5</td>
<td>Plastic covered 1.8 mm diameter wires Wood support</td>
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<tr>
<td>Loop 24M4 (**)</td>
<td>24</td>
<td>2.26</td>
<td>4.00</td>
<td>171</td>
<td>96</td>
<td>200</td>
<td>2631</td>
<td>520</td>
<td>54.7</td>
<td>Plastic covered 1.8 mm diameter wires Wood support</td>
</tr>
</tbody>
</table>

( * ) Tony Preedy, G3LNP (Ref. 16)
( ** ) Calculated only

We prefer to achieve the insensitiveness to local electric noises, generally man made, by fully-balancing the whole antenna circuit: the loop, the capacitances (a fixed capacitor plus varicap diodes for the fine tuning) and the preamplifier.

To match the high impedance of the resonant circuit with the LF receiver’s low impedances, we use an instrumentation amplifier comprising three op-amps. It provides high input impedance, high gain and bandwidth and a relatively low output impedance.

Considering the electrical characteristics of our 38-turn loop, in which \( L = 1.7mH \) and \( Q = 210 \) (Table 1), the parallel resistance of the resonating circuit, \( R_p \), is \( 2\pi fLQ \). At 136kHz, this is 305kΩ. Being in parallel with the 2MΩ input resistance of the operational amplifier, this resistance becomes 265kΩ.

Such a low resistance deteriorates the merit factor of the antenna circuit from 210 to 182. In other words, the load constituted by the input of the operational amplifier produces an insertion loss of 1.25dB.

This loss figure indicated that it was not possible to increase the loop antenna’s equivalent height as much as we would have liked. Equivalent height is limited by the impedance that can be connected at the input of the operational amplifier. Increasing this impedance also increases noise.

At this point, it is useful to consider the equation for calculating the thermal noise at the preamplifier input:

\[
\eta_p = \sqrt{4kTRB} = 0.29 \mu V
\]

considering a bandwidth of 20Hz and a room ambient temperature of 25°C. Here:

- \( \eta_p \) = noise voltage (V)
- \( k \) = Boltzman’s constant, which is \( 1.374 \times 10^{-23} J/K \)
- \( T \) = absolute temperature in kelvin
- \( R \) = resistance across which thermal agitation is produced (Ω)
- \( B \) = bandwidth (Hz)
Another limit on how much antenna equivalent height can be obtained is the stray capacitance of the loop. To try to define a limit for the antenna equivalent height, we measured the stray capacitance of our 38-turn loop, Fig. 2. It turned out to be 70pF. This seems to be a good trade-off between the physical dimensions and the electrical performance. The disadvantage relative to an optimized antenna is only 10nV more thermal noise and about 1dB lower gain.

**Key parameter for loop antennas**
The product $NxA$, where $N$ is the number of turns and $A$ the area of the loop, is the key parameter for loop antennas. However, two antennas with the same $NxA$ product may be very different in terms of inductance. Comparing a loop 'A', which has 54 turns and 0.64 area, against a loop 'B' with 18 turns and 2 area, you can see that there's a 4-to-1 inductance ratio. Higher loop inductance means higher parallel input resistance – and hence amplifier noise.

**Component choice**
To underline the electrical performances of the operational amplifier to be used: the input noise of the circuit, see Fig. 3, is 0.4pA/VHz. This equates to 0.48μV considering an input resistance of 265kΩ and a receiver bandwidth of 20Hz using high quality OP37. The figure increases if TL081 op-amps are used in the first stage.

Gain of the input stage is set at 20dB and gain of the output stage is 6 to 12dB according to your design needs.

You can use a 600Ω direct output or coaxial cable matching with a 300/75Ω output transformer. Full power bandwidth for a 20V pk-pk output is 250kHz.

As you can see from Table 1, our 38-turn loop has an equivalent height of 9.17m – even though its diameter is only 0.77m.
Magnetic-cored loop

Loop antennas can be made using a magnetic core, for instance ferrite, instead of air.

If an air-cored loop is placed in a field, it cuts the lines of the flux without disturbing them. On the other hand, when a ferrite aerial is placed in the field, the nearby field lines are redirected into the loop. This is because the reluctance of the ferrite material is less than that of the air. The reluctance is inversely proportional to the relative permeability of the rod core (µr).

In this case the equation of the equivalent height becomes:

\[ h_s = \frac{2\mu_r A N Q}{\lambda} \]

Using this kind of antenna, it is not possible to reach the equivalent height of a loop wound on wood and air. For this reason, the best use for ferrite aerials is in compact portable instrumentation.

Applying this criterion, we used the ferrite antenna to perform magnetic field measurement from five metres to five kilometres away from the transmitting antenna.

Magnetic or electric field

The field's nearness to the transmitting antenna, whether it is a vertical rod or loop type, can be calculated using the following equations. They assume that the wave path is parallel to the Earth's surface:

\[ E = \frac{30h_s \lambda I}{\pi d^2} \]

Where:
- \( E \) = near electric field (V/m)
- \( h_s \) = antenna equivalent height (m)
- \( \lambda \) = wavelength (m)
- \( I \) = effective value of antenna current (A)
- \( d \) = distance from transmitting antenna (m)

The vector of electric field is perpendicular to the Earth's surface and with the positive direction upwards.

\[ H = \frac{h_s I}{4\pi d^2} \]

Here, \( H \) is the near magnetic field (A/m).

The relevant vector is parallel to the Earth surface and in quadrature with the electric field with the positive direction.
rotated rightwards looking at the transmitting antenna.
These relationships are applicable when the $h_t$ is less than
0.1A. That is, of course, easily verified because at 136kHz the
wavelength is 2206m.

![Fig. 7. Wave Impedance in the LF Near Field and Far Field.](image)

**Fig. 7. Wave Impedance in the LF Near Field and Far Field.**

**Fig. 8. Ground and skywave propagation at 136KHz.**

<table>
<thead>
<tr>
<th>Distance (Km)</th>
<th>100</th>
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<th>300</th>
<th>500</th>
<th>700</th>
<th>1000</th>
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<tbody>
<tr>
<td>Groundwave Good Ground</td>
<td>38.6</td>
<td>30.4</td>
<td>25.9</td>
<td>17.9</td>
<td>12.5</td>
<td>1.9</td>
</tr>
<tr>
<td>Groundwave Poor Ground</td>
<td>34.7</td>
<td>17.4</td>
<td>8.9</td>
<td>-4.6</td>
<td>-20.4</td>
<td>-31.5</td>
</tr>
<tr>
<td>Skywave Night</td>
<td>-2.6</td>
<td>-10</td>
<td>2.1</td>
<td>5.4</td>
<td>8.2</td>
<td></td>
</tr>
<tr>
<td>Skywave Day (**)</td>
<td>-25.1</td>
<td>-14.7</td>
<td>-8.2</td>
<td></td>
<td></td>
<td></td>
</tr>
</tbody>
</table>

(\* ) Low solar angle ( winter or late afternoon )

Table 2. Calculated Field Strength vs. Distance in dBuV/m, for radiated power 10 of 1W.

Analysing the above equations it becomes clear that the
electric field $E$ near the transmitting antenna decreases with
a slope of 18dB each time the distance from the radiating ele-
ment doubles. This is 60dB for each tenfold increase in dis-
tance.

Likewise the magnetic field $H$ decreases with a slope of
12dB for each distance doubling, or 40dB for each order of
magnitude increment of the distance.

**Figure 4** shows experimental confirmation of this attenu-
ation rule. The experiment was performed using both a bal-
anced dipole and a loop antenna to measure the electric/mag-
netic field at different distances from the radiating element.

Figs 5 and 6.

These types of measurements are not so easy.
Remembering that at 10 metres from the transmitting source
the difference between the electric field and the magnetic
field is about 30dB, great attention needs to be paid to the
balance and shielding of the antennas involved and also to the
operating levels.

Since the input impedance of an electrically short dipole is
predominantly a capacitive reactance, broadband frequency
response can be achieved with a high-impedance load. This
is not so important for the 136kHz tests using single-fre-
quency tuning and calibration.

A 40-40cm short balanced and tuned dipole has about a
20cm electrical height, but an accurate calibration is realised
by comparison with a reference antenna in the far field zone.

By trimming the gain of the high-input impedance dipole
amplifier we measure exactly a 1mV out on the precision
receiver for a known field of 1mV/m.

At this moment it is probably necessary to better define
what you mean by Near and Far Field.

In the technical literature there are many definitions of
the boundary between near and far field:\

We prefer to assume the edge of the near field at the dis-
tance which the wave impedance $Z_0$ becomes:

$$Z_0 = \frac{E}{H} = \frac{\mu_0}{\varepsilon_0} = 120\pi = 377\Omega$$

where:

$\mu_0$ = absolute magnetic permeability of the air = 4\pi x 10^{-7} H/m
$\varepsilon_0$ = absolute dielectric constant of the air = 8.85 x 10^{-12} F/m

This occurs at a distance from the transmitting antenna, given
by the following equation:

$$d = \frac{\lambda}{2\pi} = \frac{351m}{136kHz}$$

In the near field (d<351m) condition a vertical rod will gen-
erate mainly a high impedance electric field, while a loop
aerial will produce mainly a low impedance electric field.

This kind of behaviour is well shown in the Fig. 7 in which
is also displayed a transition region, about one sixth wave-
length wide, between near and far field regions.

This point of view is in accordance with the CCIR 368-7
recommendation that establishes to measure the effective
radiated power, through a field measurement, at a distance of
1km from the transmitting antenna because at this distance
the plane wave condition is also satisfied.

$$P = \frac{E^2}{90}$$

where:

$P$ = effective radiated power (W)
$E$ = electric field (mV/m)
In the far field condition, the electric field is given by the following equation:

\[ E = \frac{60\pi I}{d\lambda} \]

And consequently the magnetic field becomes:

\[ H = \frac{51 I}{d\lambda} \]

At a distance greater than far field condition \((d > 35 \text{ m})\) the slope of both the magnetic and electric fields versus the distance become 6dB each doubling or, if you prefer, 20dB each decade.

This kind of trend is valid until 300–500km, for the frequency of 136kHz, even if the fall is influenced by the imperfect ground conductivity \((\sigma\text{m})\) that worsens the slope as reported in the Fig. 8 where 

\[ \sigma = 10^2 \text{S/m} \quad \text{and} \quad \sigma_2 = 10^3 \text{S/m} \]

Until now the ground wave has been described. It concerns the electromagnetic fields travelling along the earth surface induced and being induced by the current flowing on and slightly below the earth surface. Sometimes those fields are defined as Surface Waves.

At distances greater than 300–500km the Ground Wave drops down faster and becomes significant compared to the wave reflected by the ionosphere.

The model performs some assumptions to simplify the geometric computation of the Sky Wave:

- the ionosphere is a zero thickness layer having a height of 70km daily and 90km nightly
- the Sky Wave path is a straight line
- the Earth is considered a perfect sphere
- the coefficients (ionosphere reflection and focusing factors, RX/TX antenna ground pattern factors) have been introduced in order to meet practical measurements with the theory.
- the ground conductivity \(\sigma=2\times10^2\text{S/m}\) and the ground rel-

![Image](image-url)

**Table 3. Typical DCF39 received signals (dB[V/m]) in Europe.**

<table>
<thead>
<tr>
<th>Km.</th>
<th>dB[V/m]</th>
</tr>
</thead>
<tbody>
<tr>
<td>100</td>
<td>85</td>
</tr>
<tr>
<td>200</td>
<td>79</td>
</tr>
<tr>
<td>300</td>
<td>74</td>
</tr>
<tr>
<td>500</td>
<td>60-66</td>
</tr>
<tr>
<td>750</td>
<td>45-50</td>
</tr>
<tr>
<td>1000</td>
<td>34-51</td>
</tr>
</tbody>
</table>

**Table 4. Loop antennas: calculated and measured output voltages (using DCF39 signal at 750km).**

<table>
<thead>
<tr>
<th>Measure</th>
<th>DCF39 Field</th>
<th>Reference Loop 18 (*)</th>
<th>Ferrite Aerial Length=60/mm</th>
<th>Tuned 135-139kHz Super Loop 38</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>N=18, A=0.0754m²</td>
<td>N=100</td>
<td>Q=120</td>
<td>N=38, A=0.47m²</td>
</tr>
<tr>
<td>(\mu\text{W/m)}</td>
<td>(\text{dB}\mu\text{W/m)}</td>
<td>(\mu\text{V)}</td>
<td>(\mu\text{W)}</td>
<td>(\mu\text{V)}</td>
</tr>
<tr>
<td>Calculated</td>
<td>800</td>
<td>58.1</td>
<td>3.09</td>
<td>0.00386</td>
</tr>
<tr>
<td>Night h. 22.00</td>
<td>800</td>
<td>58.1</td>
<td>3.09</td>
<td>0.00386</td>
</tr>
<tr>
<td>Night h. 21.00</td>
<td>737</td>
<td>57.3</td>
<td>2.84</td>
<td>0.00386</td>
</tr>
<tr>
<td>Day h. 15.00</td>
<td>300</td>
<td>49.5</td>
<td>1.16</td>
<td>0.00386</td>
</tr>
<tr>
<td>Day h. 17.00</td>
<td>580</td>
<td>55.3</td>
<td>2.23</td>
<td>0.00386</td>
</tr>
</tbody>
</table>

(*) Output voltage measured with RL=100Kohm and BW=20Hz

---

**Fig. 9. The Vertical Antenna of the DCF39 station in Magdeburg (324m high).**

**Fig. 10. Measurements of the Far Field Signal from DCF39 (by DK8KW and OH2LX).**
EIRP (Emitted Power referred to an isotropical antenna) is about 40kW omnidirectional, confirmed by many measurements\(^{12,13}\) taken by DK8KW and OH2LX, Fig. 10, in April 2000.

The DCF39 station is intended for long wave teleswitching which is a new way in load management technology. It replaces the ripple-control technology, which is widely used in the utility industry worldwide. It is used for tariff-switching applications and load management as well as for the control of street lighting (the management of modern power supply systems requires the transmission of commands to control the consumption of electricity at any time). The newly offered LF teleswitching system is using the DCF39 radio channel to transmit the information.

**Antenna calibration**

With the availability of a suitable radio signal (as DCF39) the calibration of unknown loop or ferrite antennas is not so difficult. The first step consists of realization of a simple reference antenna or sensor (a magnetic-field probe or reference loop consists of an electrically small, balanced antenna) which is obtained by winding \(N\) turns of wire on a support of known area (A).

The complete original formula shows the \(h_s\) (equivalent height) of a corresponding vertical aerial.

The product of the equivalent height (metres) multiplied by the local field \((E=1\text{mV/m})\) is the received signal. At 136kHz we can use a simplified formula to show the unloaded output voltage of the simple but accurate reference sensor:

\[
v = h_s \times E = 0.00386 \times 10^{-3} = 3.86\text{mV}
\]

In our tests a plastic basin with a diameter of 31cm was used as support, Fig. 11. With \(N = 18\) turns (of 2.5mm\(^2\) copper wire) and \(A = 0.0754\text{m}^2\) loop aerial in a received field \((E)\) of 1mV/m the measured open circuit output voltage is 3.86mV (with S/N>20dB).

This type of reference loop was been tested\(^{12,13}\) by PA0SE and SM6PXJ with measured and calculated values within better than 0.5dB. The standard method used is that of the Helmholtz coils, but other people proposed a more simple test using the ANSI/IEEE standard\(^{15}\) 644-1987 normally suggested for 50/60Hz EMF probe calibration. The Helmholtz coil can provide a uniform, known magnetic field (H); the test object (ferrite aerial or small air loop) is centered equidistantly between each side of the coils. The accuracy of the coil was checked with a small calibration loop (5 turns, diameter 76mm) connected to the selective level meter.

### Table: Multi rods ferrite aerials: Calculations and Measurements

<table>
<thead>
<tr>
<th>Ferrite Rods</th>
<th>#</th>
<th>Length (mm)</th>
<th>Equivalent Diameter (mm)</th>
<th>L/D</th>
<th>Area (mm(^2))</th>
<th>(\mu\text{rod} \times A)</th>
<th>L ((\mu\text{H}))</th>
<th>(h_s) (m)</th>
<th>Q</th>
</tr>
</thead>
<tbody>
<tr>
<td>Single</td>
<td>1</td>
<td>200</td>
<td>10</td>
<td>20</td>
<td>78.5</td>
<td>118</td>
<td>9267</td>
<td>0.31</td>
<td></td>
</tr>
<tr>
<td>Two in series</td>
<td>2</td>
<td>400</td>
<td>10</td>
<td>40</td>
<td>78.5</td>
<td>210</td>
<td>16493</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Three in series</td>
<td>3</td>
<td>600</td>
<td>10</td>
<td>60</td>
<td>78.5</td>
<td>260</td>
<td>20420</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Two series and two in parallel</td>
<td>4</td>
<td>400</td>
<td>14</td>
<td>29</td>
<td>154</td>
<td>166</td>
<td>25554</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Three series and two parallel</td>
<td>6</td>
<td>600</td>
<td>14</td>
<td>43</td>
<td>154</td>
<td>220</td>
<td>33866</td>
<td>1100</td>
<td>1.16</td>
</tr>
<tr>
<td>Three series and three parallel</td>
<td>9</td>
<td>600</td>
<td>17</td>
<td>35</td>
<td>227</td>
<td>185</td>
<td>41991</td>
<td></td>
<td></td>
</tr>
<tr>
<td>Three series and four parallel</td>
<td>12</td>
<td>600</td>
<td>20</td>
<td>30</td>
<td>314</td>
<td>170</td>
<td>53407</td>
<td>1.82</td>
<td></td>
</tr>
</tbody>
</table>

or calibrated receiver. The maximum error was found to be within 0.1dB.

For the maximum accuracy of the tests it is very important to avoid resonating frequencies and parasitic capacities. In our 18 turn coil we have: L = 155 μH (XL = 132 Ω at 136kHz) and an autoresonating frequency of 1.2MHz. With a 100kΩ input impedance of the test setup we can measure exactly the open circuit voltage generated by the reference loop and also with a 600Ω input impedance we have a load error of only about 1dB.

One secret: all the tests with the DCF39 (at 750kHz from the transmitter) are made using high selectivity receivers with very narrow bandwidth (example: BW=20Hz).

Starting from a calibrated Reference Antenna we have measured three other interesting aerials: an untuned 38 turn 77cm diameter loop, the same with a tuned and loaded by the preamplifier input impedance (Q = 182) and a very portable Ferrite antenna.

These and other results are shown in Table 4. For people interested in details and phase of loop antennas the articles in references 16, 17 and 18 are advisable. For the Ferrite Aerials the calculated values for a number of ferrite rods are shown in Table 5. Such antennas mainly utilize the magnetic field component of the signal to be received, and the directional characteristics of the antenna correspond to that of a short dipole, which is an "S" with a flat maximum and a sharp null. 100 turns of Litz wire (many thin wires) may be wound on a single core (basic permeability = 500), or to increase the output, the core may be two or more rods taped together. Best performance is obtained with groups of rods glued end to end contained in a U-shaped electrostatic shield.

As shown in the table, the maximum suggested number of ferrite rods is about six. The calculated improvement with nine or 12 rods is not impressive. The equivalent height (h_e) of our realization (three rods in series x 2 rods in parallel = 6) is about 1 metre (calculated 1.16m). This is a good solution for portable use as secondary reference antenna. For more info on the ferrite aerials see also references 19 and 20.

Conclusions

The Loop Aerials are extremely interesting for receiving in the 136kHz band because of their specific characteristics: high gain, high selectivity, directivity and low interference noise. The possible limits for an optimised big loop at 136kHz (Q = 200, BW = 680Hz) are about: area (A) =8 - 10m². N = 30 turns, h_h -> 50m. This antenna has a good rejection to the local electric noise and an equivalent height not obtainable with any "practical" vertical Marconi antennas.

The more important parameters of a few loop aerials have been tested and the theoretical equivalent heights (h_h) confirmed using the DCF39 comparison method. Our record in the experimental tested antennas was h_h=30m. Other experiments and statistics are necessary to have a more complete knowledge of Signal to Noise optimization of loops.

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(*) Reference Data for Engineers is now published by Newnes.

August 2002 ELECTRONICS WORLD 19
During the first half of the last century, interference problems began to manifest themselves in valve equipment. One attempt to solve this problem was to wire all the components to a single point on the chassis. The 'star point ground' was conceived.

The desired effect was noise reduction. The opposite effect was achieved, in fact interference problems were created and these problems persisted for the lifetime of the equipment. In spite of this the idea gained widespread acceptance and some influential engineers still recommend it. As a guideline for circuit designers wishing to achieve Electromagnetic Compatibility (EMC) for their products, it has long passed its use-by date.

The fact that it retains wide acceptance identifies an even more deep-seated problem: too great a reliance is being placed on guidelines, tips, fixes and on the pronouncements of EMC gurus. This is a hit or miss approach. Guidelines become outdated as technology progresses, tips and fixes that work beautifully in one application are disastrous in others and gurus distance themselves from the project before problems appear. This note identifies the fallacy in the star point ground concept and points to a systematic approach to those aspects of design that achieve EMC of the product.

The star ground concept

Star point grounding is a method of wiring circuits that minimises the resistive coupling between two separate circuits. Fig. 1. illustrates the idea. The boxes A, B, C, and D can be thought of as printed circuit boards containing interface circuits.

The wiring is organised to carry signal 1 from A to B, and signal 2 from C to D. Return conductors are all routed via the star point, S. Since there is no resistive component common to both circuits, there can be no resistive coupling between them. The reasoning is that, if there is no common coupling, there can be no interference.

The fallacy

The fallacy in this reasoning is that it limits its consideration to resistive coupling. Magnetic and electric field effects are ignored. If inductive coupling is considered, the picture changes completely. In Fig. 1, the current I1 flows in a loop enclosing a wide area. A great deal of magnetic flux threads through this area. Inevitably, a significant proportion of this flux also threads through the second loop. Transformer action ensures that a relatively high voltage is developed in series with the second loop. This appears as an interference source; an unwelcome addition to the desired signal.

Where there are magnetic fields, you will find electric fields. These manifest themselves as capacitance coupling between the conductors and add their own contribution to the interference. Signal 2 will interfere with signal 1 in exactly the same way. Star point grounding creates a system in which every signal interferes noticeably with every other. If the system interferes with itself, of what use is it when subjected to an environment where the external field is greater than that of the signals being processed?

Alternative approach

If star point grounding is to be abandoned, what should replace it? Perhaps the best approach is to start with an overview of the system and then to implement the lessons learnt from theory. The initial objective can be formulated: to transmit one signal from A to B and another from C to D, with minimal interference between the two signals. It is assumed that there are a number of other circuits in the overall system and that cable conductors will be used to carry the signals.

Transmission line concept

Some fundamental concepts of electromagnetic theory and
Circuit theory are combined in the picture of the transmission line shown in Fig. 3. Current in the upper conductor is matched by an equal current in the lower conductor, flowing in the opposite direction. Illustrated are the electric and magnetic field vectors, E and H, at the midpoint between the conductors. There is a flow of electromagnetic power from left to right, identified by the 'P' vector. Some simple points can be made, namely, the currents in the supply and return conductors are equal and opposite at every cross-section of the transmission line and the vector sum of the current at any section of the line is zero. The action of the electromagnetic field tends to provide this equalisation. Don't fight it. Use it.

The most efficient way to transmit electric power between two points is to use a transmission line. Minimal power is transmitted to the environment and minimal power is received from the environment. A logical decision is to use transmission lines to carry the signals defined in the block diagram. Although the vector sum of the currents is zero in Fig. 3, the power vector clearly indicates which way the signal is going. This allows a very useful correlation to be made between the transmission line and the block diagram.

Wiring Diagram
If the block diagram is modified to include the conductors of the transmission line, the natural result is a wiring diagram and the components of Fig. 4. begin to emerge.

In any practical system, there are a fair number of other conductors. These include the supply conductors necessary to distribute power to the various printed circuit boards. Signals at the individual boards are processed with respect to a common conductor, usually designated as the 'ground' reference. There is also some form of shielding, provided in part by the equipment structure. The inclusion of the conductor marked 'structure' in the diagram allows the existence of the grounding and shielding conductors to be recognised. In the illustration of Fig. 4, the return conductors are all grounded to local points on the structure.

Culprit Circuit
Any interference must have a source, a coupling mechanism, and a receptor. The term 'culprit' can be used to identify a network generating unwanted emissions, whilst a network which could be susceptible to interference is a potential 'victim'.

In the case under consideration, both culprit and victim are part of the same system, and the coupling mechanism is associated with current in the structure. If the culprit is assumed to be the wiring associated with signal 1, then it is logical to focus first on this segment of the system.

A circuit model can be created of the culprit, by treating it as a three-conductor transmission line. Fig. 5, is a simplified model, where each conductor is represented by an inductor. Each conductor also possesses the properties of resistance and capacitance, but there is no need to show these in an initial illustration. It is always possible to assign a value to each inductor. Any basic textbook that introduces three-phase power lines will provide equations relating physical dimensions to inductance values. If necessary, tests can be made on a representative assembly to measure the values. From a system point of view, the spurious output of the culprit is transient current in the structure, I3.

Common-mode rejection
There are two loops involved: the differential loop carrying signal current, and the common-mode loop carrying a portion of the signal current via the structure. A wire pair is usually constructed with identical conductors and these are held as close together as is physically possible. The separation between supply and return conductors is usually greater than that between cable and structure. This means that inductors L1 and L2 of Fig. 5, are equal, and have as low a value as is possible. Conversely, L3 has a relatively high value.

If the signal source is located on printed circuit board A, and the supply current I1 flows in L1, then the return current will be shared between L2 and L3. Since L2 is less than L3, a greater proportion of the return current will flow in L2. This means that I3 is less than I2. The ratio between I2 and I3 is even greater. That is, there is a useful amount of common-mode rejection, due to magnetic effects.

**Coupling Mechanism**
Common-mode current flowing in the structure will generate a voltage across L3, and the amplitude of this voltage can be calculated. Interference created by signal 1
will manifest itself as a voltage along the structure - 'Vthreat'. Invoking the Norton-Thevenin relationship of Fig. 6, allows the action of the culprit loop to be represented as a voltage source. Vthreat, in series with the structure.

From the point of view of the culprit, interference can be defined as the current, I3, in the structure. From the point of view of the victim, interference can be defined as the voltage, Vthreat, in the loop formed by structure and cable.

**Victim Circuit**
This interference source can then be included in the circuit model for the second signal, as shown on Fig. 7. In this model, common-mode current flows in the cable/structure loop and creates a voltage across L5. Since L4 and L5 act as an inductive potentiometer, the voltage induced in the differential loop will be significantly less than Vthreat. Again, there is a useful amount of common-mode rejection, also due to magnetic effects.

**Ground loops**
One feature of this approach is that it has introduced two extra loops into the configuration - the common-mode loops of the culprit and victim circuits. It has been shown that the action of the magnetic field in these loops reduces the level of coupling between culprit and victim. Another name can be given to these loops - 'ground loops.' In fact, the terms 'ground loop' and 'common-mode loop' are synonymous.

This means that the Breaded ground loop, which many individuals believe should be avoided if at all possible, actually helps to improve EMC.

**Improving performance**
Current in the ground loop is the prime cause of interference. To improve performance, the objective should be to reduce the amplitude of this current. Increasing the impedance of the loop can do this. The most obvious way to increase loop impedance is to open-circuit it. This leads to the familiar concept of the floating termination. From an examination of Fig. 7 it could be assumed that a floating termination would reduce common-mode current to zero, and solve the problem. Alas, it is not to be.

Up till now, attention has been focussed on magnetic effects. The action of the electric field has been ignored. There have been no capacitors in the circuit models. If the victim circuit of Figure 7 is modified to show the existence of these capacitors, to 'float' the receiver interface, and to replace the load Zd with an optocoupler, then the picture becomes as shown in Fig. 8.

The capacitors now provide a path for common-mode current. At low frequencies, this current has negligible amplitude, and common-mode rejection can be as high as 60 dB. However, as the frequency of Vthreat increases, common-mode current increases. The common-mode rejection is a function of frequency, and reduces at 20 dB per decade. The combined existence of capacitance and inductance means that, inevitably, there is resonance. At the resonant frequency, the differential voltage can be 10 dB higher than Vthreat. Of even more concern is the fact that the common-mode voltage at the optocoupler (between 'return 2' and structure) can be more than 40 dB higher than Vthreat. This raises more problems.

**Implications**
These problems can be solved. However and there is no need to describe the solutions here. The point that can now be made is that circuit modelling will provide a clear picture of the coupling mechanisms. When the problem is clearly defined, a solution can always be found.

As well as providing a clear picture, circuit modelling allows actual numbers to be assigned to component values, and for the frequency response of the system to be analysed. Circuit analysis software makes the calculations a simple task. Simple bench tests 2 can be devised to measure the response during product development. If necessary, the circuit can be modified and the analysis repeated, until the system is shown to meet its EMC requirements. The finished product can be submitted for formal EMC Tests with a high degree of confidence.

**Conclusion**
There are many guidelines, tips, and fixes to be found in the literature on EMC, and there is much advice provided by experts on the subject. Some of it is of dubious value. Using circuit models of the system under review, it is possible to identify the hidden assumptions, the limitations, and the errors in any particular recommendation. Circuit modelling allows the electromagnetic coupling mechanisms to be understood and analysed. The systematic use of circuit models will enable any system to be designed to meet its EMC requirements.

**References**
Key Factors in RF Power Amplifier Design

Although radio and amateur radio are a bit old-fashioned, today a lot of engineers have to deal with RF, e.g. on topics like cordless telephones, mobile phones or wireless LAN.

For some main-stream systems like GSM or AMPS, RF power amplifier modules are available from manufacturers like Hitachi, Fujitsu, Alps. etc. That eases the application, because they normally have 50Ω RF IOs. But such modules are quite expensive, MMIC’s are often cheaper. For some systems even a discrete solution might be competitive. In these cases - or for module or chip design - a much more detailed know-how is needed.

On a system level, such things like RF TX power at the antenna, power-time template and spurious signals are specified. So the best way to design an RF PA is starting with a level diagram. From this you get the output power of the PA. After designing the final output stage with its matching networks you get the input power needed to drive the last stage. Step-by-step you can go backwards to the fist PA stage which is normally connected to a modulator. VGA or VCO. S-parameters are only a good characterisation for small signal circuits. Power amplifiers are often very non-linear and the S-parameters will depend on power level. Despite this, S-parameters measured at the input port at the power level also used in the application are a very good starting point for the design of the input matching network. Even more critical is the output of an RF power amplifier. Power match based on small-signal S-parameters will result in highest small-signal power gain, but for RF power amplifiers the output power itself and the efficiency (normally specified by the so-called power added efficiency PAE=(Pout−P0)/PDC) are much more important. So the question is: What impedance

\[ Z_{\text{Opt}} \]

should be applied at the amplifier output to get a given output power with best efficiency? Many people are using an impedance tuner to search for the best match in the lab by hand. This will lead to a completely different design procedure than typically used in small-signal amplifiers! A faster way is possible here with some theory.

Let us consider a concrete design problem: Design a matching network for an ISM 2400MHz power amplifier (free band for industrial-scientific-medicine applications). In the USA, up to 1W (corresponding to 30dBm) antenna power is allowed for this frequency band. In reality, some loss occurs in the TX low-pass or band-pass filter and the antenna switch, so the PA is allowed to deliver approx. 31dBm. Because you need some safety margin for component tolerances, temperature drift, changes of supply voltage and RF input power, a PA with a nominal output power of 29dBm will be well-suited. On the market there are not many low-cost PAs which are able to deliver such high output power at 2.4GHz. For instance, Infineon has a Silicon PA family starting from a 22dBm Bluetooth PA up to the largest 29dBm device. All devices are balanced PAs with push-pull input and output stage. The balanced input eases the connection to the often also balanced transceiver output. To save board space and external components many system functions are included in these PA devices, such as power ramping and antenna switch drivers. A nice feature is the power select function. With two digital pins you can select four different output power levels, e.g. according the distance between handset and base station. For the balanced output we need a balun (balanced-to-unbalanced) to convert

Fig. 1. Two passive RF tuners used to sweep impedances.

Fig. 2. Our 2-stage PA system topology.

**Supply & Bias**

Driver
Stage

Interstage
Match

Output
Stage

Pre-Match

Supply Voltage

Balan

Filter & RX-
TX Switch

August 2002 ELECTRONICS WORLD
Fig. 3. Calculating $R_{\text{Logi}}$ via ANPASS

Fig. 4. PA output modelling in CS411TH and the L-type pre-matching network to 35L. Note: The end capacitor has a series inductances of 0.5-0.6nH as a typical 0603 SMD component.

Fig. 5. LC balun design using ANPASS.

the push-pull signal to the normally used single-ended signal (e.g. for filters, PIN diode switches and antenna).

The output power depends not only on the PA device but also on supply voltage $V_{\text{CC}}$ (due to $P=V_{\text{CC}}^2/2R_{\text{out}}$) and best efficiency PAE can be expected if the PA is deep in the compression (in this case app. 40%). This operation is allowed for systems like DECT (digital enhanced cordless telephone), HomeRF or Bluetooth (both new standards for general-purpose RF interfaces, WLANs, etc.), because they use modulation schemes (in these cases frequency shift keying) with constant RF envelope. For non-constant envelope modulation schemes like QPSK or 8PSK (e.g. IEEE801.11b or UMTS), you have to look at the peak power, not the average power. This is needed in these cases because a PA in compression would create too much adjacent channel leakage power. The Infineon device is fabricated in a 4V-25GHz silicon process, so for 29dBm the recommended supply voltage is 3.1V. Direct operation at two NiCd/NiMH cells is possible, because the supply voltage range starts at 1.9V. With this information we can calculate the optimum load impedance $Z_{\text{Logi}}$. A nice program to do this is the AdLab tool ANPASS [1]. It uses the formula $R_{\text{Logi}}=V_{\text{CC}}^2/2P_{\text{wanted}}=V_{\text{CC}}^2/2P_{\text{wanted}}$, which is pretty accurate for class-A operation (hints available on bubble help). There are some problems: Firstly we can only guess the saturation voltage, which should be close to app. 0.2V, because it’s a low-voltage bipolar design. Secondly we operate in deep compression, so the class-A approximation is not valid. For instance for class-E [2] the voltage swing is not $2(V_{\text{CC}}-V_{\text{sat}})$ but app. 3.5 ($V_{\text{CC}}-V_{\text{sat}}$). For the class-A approximation and $V_{\text{sat}}=0.2V$ ANPASS delivers $R_{\text{Logi}}=4.9\Omega$ for a single-ended PA and 19.6$\Omega$ for the balanced topology. This shows a clear advantage of the push-pull output, its impedance is already closer to 50$\Omega$.

The result is a real value for the impedance (19$\Omega$, so 9.5$\Omega$ for each side) which is not truly realistic with real world transistors and finite package inductances. So ANPASS delivers the correct value for an idealised PA. For compressed class-B operation a higher value of $R_{\text{Logi}}$ is a bit better for higher efficiency (say 11$\Omega$, for class-E operation ANPASS delivers 5.64$\Omega$ for single-ended...
configuration). Using another AdLab tool called CSMITH we can start with the corrected value as the generator impedance and we can add the transistor output capacitance (approx. 3pF with some series resistance representing losses in the silicon substrate) and the bond-wire inductance (app. 0.4-0.5nH and a small package capacitance) by hand. Note that CSMITH is able to use real elements with all their major parasitics like series resistors or inductances, also a frequency sweep with graphical output for gain, MAG, return loss, etc. is available.

What we need now is a match from the transistor output to the balun. Because we need a DC-feed, a L-type low-pass structure (high-impedance transmission line acting as a series-L followed by a shunt-C) is the easiest solution. In other situations a high-pass is a better choice, e.g. in the interstage match where a DC-break is needed or some compensation of the drop of the transistor gain at higher frequencies is needed.

A balun generally transforms a differential signal to a single-ended one (which is normally 50W) and vice versa. A standard LC balun can be designed using ANPASS. One open question is the intermediate balun input impedance. It's a good idea to take an intermediate impedance value (say 35W), so that the match is distributed over the first prematching network and the balun. This often gives the largest bandwidth and low tolerances. Other types of baluns are well-known (e.g. with transformers or L-transmission lines), but the LC all-pass is preferred here because it is very compact. Note, one balun capacitor could be merged with the shunt-C of the prematch.

The resulting circuit is very close to what we have achieved in the lab. Of course in reality some tweaking is always needed in 2GHz circuits due to component parasitics and modelling inaccuracies. Also the impedances at the harmonic frequencies are not unimportant due to large signal operation. This behaviour is known as harmonic matching, but it is not easy to get an advantage from this behaviour at a GHz power amplifier.

For higher output power levels the impedances become very low (e.g. typically 2Ω at GSM levels) and a single-step matching network would result in a small bandwidth, but more importantly in tolerance problems. In these cases you need a multi-step match. In principle such a matching network can be designed in the same manner using the Smith chart, although it is not easy to optimise both losses and bandwidth. The main problem is that in the Smith chart you normally calculate at one frequency, so you often don't get the bandwidth advantage of more complex circuit structures like Chebyshev filters. In CSMITH you can do such a design, because Monte-Carlo analysis, frequency sweeps and also optimisation (in conjunction with the general-purpose simulator APLAC [4]) are available.

Currently, we are only looking very roughly at the transistor. In fact, so far we only look at its saturation voltage, its current and voltage capabilities and its output capacitance. Of course other

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**Figure 6.** Measurement results for the 29dMm Si PA

**Figure 7.** The 2.4GHz PA board with the Infineon 2.4GHz-PA in VQFN20 package

**Figure 8.** CSMITH results of a 1.9 GHz 3-step matching network optimised for wide bandwidth. the MAG (upper curve) shows that the element losses increases at higher frequencies, so it's not easy to get a true flat response.
parameters such as feedback capacitance, transition frequency $f_T$, maximum frequency of oscillation $f_{max}$, maximum available gain MAG, stability factor $k$, current gain $B$, etc. are important - but not so much for the output match. Often a carefully chosen compromise is needed. For instance transistors with high $f_T$ and $f_{max}$ (like the new Silicon-Germanium technologies) have a high power gain $G$, which is advantageous for high PAE and getting a low number of RF stages. But these transistors tend to have lower breakdown voltages and might be less stable. As a rule of thumb the supply voltage should not exceed the transistors $V_{CEO}$, although breakdown behaviour also depends on the impedance at the transistor base ($V_{CEO} < V_{CER} < V_{CEO}$). Your transistors should be stable at the operating frequency ($k$=1), so the MAG is a good indicator of the possible gain. If the device is not stable you need damping elements (e.g. series resistor at the base) or feedback (series or shunt feedback).

Not only the transistor is important but also all layout parasitics, like emitter-ground inductance, parasitics of SMD components and also on-chip parasitics [3]. Many chip designers think only the parasitic capacitances and series resistances are critical for their layout, but this is completely wrong for low-impedance RF circuits, such as PAs. Even small metal traces within the interstage match are critical. A typical 300μm metal trace will have an inductance of approx. 0.3nH and a series resistance of 0.5Ω. Note that at 2.4GHz the inductance corresponds to $j4.5\Omega$, so the reactive part might influence the match and the frequency response seriously.

Most important is the ground inductance of the emitters (or sources for field-effect transistors) and in some cases (>2GHz & P>2W) only chip vias (available in many GaAs or LDMOS technologies) or a balanced concept will help. For a GSM PA the AC peak-to-peak current is in the range of 4A, so even 100pH will cause a ripple of 2.26$V_{pp}$ at 900MHz. This is a non-negligible part of the supply voltage and will reduce power gain dramatically and influences also PAE and stability. On the other hand some emitter inductance can help if the input impedances become low (e.g. <<1Ω), which will cause matching problems. The bipolar transistor input impedance is app.

$$Z_{in} = Z_{eg} \cdot \beta (f) \cdot Z_{g} = \frac{V_{T}}{I_L} + R_L + j\omega L_L$$

For high power amplifiers this will become $Z_{in} = 2\pi f \cdot L_L$. This is a nice result, because it is a real value which can be adjusted easily. Due to PAxPR the input power is proportional to $L_L / f_T$. Hence the power gain increases linearly with $f_T/L_L$. The other parameters are less important.

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**Table 1: Summary of key factors in modelling for RF power amplifiers.**

<table>
<thead>
<tr>
<th>Topics</th>
<th>Influence</th>
<th>Comments</th>
</tr>
</thead>
<tbody>
<tr>
<td>Transistor models</td>
<td>May have a large influence, especially on interstage matching!</td>
<td>Gummel-Poon may be sufficient for Si, but not in all cases. High current/low voltage region is critical, also quasi-saturation and breakdown.</td>
</tr>
<tr>
<td>Capacitances to substrate</td>
<td>Often a low influence (not for transistor or MOS-C capacitances)</td>
<td>This is different to low power/high impedance designs.</td>
</tr>
<tr>
<td>Series resistors</td>
<td>Medium influence. Look also at the on-chip MOS capacitances</td>
<td>Reduces gain</td>
</tr>
<tr>
<td>Series inductances</td>
<td>Large influence Not only as feedback in BJT emitters stages</td>
<td>Changes frequency response</td>
</tr>
<tr>
<td>On-chip coils</td>
<td>Medium influence. A peak Q of 5...10 is realistic for Si technologies. Include the lines to the coil.</td>
<td>Modelling is not too difficult, but Q is limited for typical Si technologies</td>
</tr>
<tr>
<td>Package model</td>
<td>Strong influence due to series inductances</td>
<td>Not easy to model</td>
</tr>
<tr>
<td>Substrate model</td>
<td>Medium influence on bias and RF performance</td>
<td>Difficult to model, important for mixed mode designs</td>
</tr>
<tr>
<td>PCB and external components</td>
<td>Large influence</td>
<td>Grounding and crosstalk are difficult to model</td>
</tr>
<tr>
<td>Bypassing and biasing</td>
<td>Large influence on stability and linearity</td>
<td>Don’t optimise only at the operation frequency</td>
</tr>
</tbody>
</table>

![Fig. 9: Effect of parasitic inductance on a 2.2pF SMD capacitor (LCFILTER from ELEKTA Professional [5]). At 2.4 GHz the component acts as a 2.9pF cap, because we operate not so far from the self resonance frequency.](image-url)
but base resistance rBB and feedback capacitance C_{PC} still have a strong influence, especially on stability factor k and isolation.

Careful biasing and supply bypassing is needed because any RF PA will not create trouble only at the operating frequency as especially at lower frequencies they often become unstable. In practice the transistor should ‘see’ no too extreme impedances at all its three terminals and over its entire active frequency range. Often damping resistances are necessary and can be part of the bias network. It is very interesting to see that bypassing with high-Q capacitors is in many frequency regions much worse compared to caps with lower Q, hence larger series resistors. The minimisation of any series inductance is very important and sometimes you need three or four capacitors with well-chosen values.

Some people say simulating RF power amps is nearly impossible, but this is not true. With careful modelling you can increase accuracy step-by-step. The remaining errors should by finally smaller then 1dB in output power and gain. To not overlook any aspect you should always ask yourself is what you calculate really close enough to reality.

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Star grounding
I cannot agree with Ian Darney "The star point concept is a thoroughly bad idea, and is based on a needless concern" (Letters, April 2002).
Indeed, having worked in a Department of Arics and Sparcs (Dept of Plasma Physics, Uni of Sydney) for a couple of decades I can report that not eliminating earth loops and other multiple earth paths (i.e. not adopting a star or tree like topology for earth) will most certainly result in some intractable interference problems in many circumstances.
Therein is part of the issue. Not everyone is wrangling small signals is such a hostile environment, but the intelligent approach to interference problems requires that one assess the nature and cause of the interference and deal with it appropriately.
Generally speaking a branched/star/tree topology will result in far fewer interference problems that one with multiple electrically parallel earth paths.
Judging by Ian's description of the problems with valve radios I suggest that the problem was not caused by the star earth topology but primarily by other, bad wiring practices. The electrical topology of the star and the routing of sensitive wiring (away from hostile parts of the circuit) must take precedence over the physical shape of the star. My feeling is that the position of the star point could have been better chosen, although this can only be confirmed by proper measurement and experiment.
The physical layout of the wiring is important. The area enclosed by a signal wire and its earth should be as small as possible, otherwise the wires will act like a loop antenna and pick up all sorts of junk. They are prone to radiate as well, and in the RF bands the extra inductance will ruin your matching. For this reason some situations mandate the signal and earth wires (power and return, for that matter) be twisted together as much as possible, or screened cable be used. This may even apply for some DC feeds, if the load current is pulsed for example. DC wiring can also act as a receiving antenna that funnels interference into shielded parts of a system.
Ian's description of "a set-up where there are several items of equipment" is altogether too sketchy to draw any but the vaguest conclusions. The nature of the set-up, how the equipment is wired internally and interconnected, the range of frequencies, voltages, and currents concerned, all impact on whether there is likely to be a problem with interference. Consider the following example of conducted interference in a real, well-shielded set-up.
In the Tokamak Lab in Plasma Physics there was a screened room (approximately 3 metres square) to shield the data taking and control equipment from the tokamak, which included several big sources of interference (like the main field current of about 20kA fed from a 2.5kV cap bank, and three 20kW RF sources). On a particular occasion a student was trying to view a signal of about 50mV by 50s on a CRO (scope), which was triggered from a 50V (nom) 3us trigger pulse. The signal and the trigger pulse were routed diagonally across the screened room from the bulkhead where they entered the room, earths connected, to the CRO (in the opposite corner) through terminated RG58 co-axes. The signal showed a spurious pedestals of about 20mV amplitude and 20us duration, caused by some of the earth current of the trigger pulse running down the signal's co-ax shield.
Fitting an isolating pulse transformer at the CRO's trigger input cured the problem. We had electrically gone from a loop topology for the earths to a branched (star, tree) topology. Far from causing any intractable problems it cured one. Generally it was found that equipment used in the screened room had to have its earth through the power disconnected to kill earth loops, although that had some unpleasant effects on people who touched a chassis that was not otherwise earthed while it was plugged in. There have been numerous other instances such as a zapped PC, a magnetron that apparently consumed more power.

Super Regen
I have been following the recent articles on super-regeneration by Eddie Insam with interest and his suggestion for an electronic tape measure using a Doppler module may well be a perfectly practical proposition. Indeed, under Recent Inventions, the November 1947 Wireless World describes something very similar.
A super-regenerative circuit is used both as a transmitter and receiver for short-range radar, the quench frequency being manually adjusted until it coincides with the time interval between an outgoing pulse and the incoming echo from a distant target, coincidence occurring when the normal 'hiss' ceases in the valve circuit. A patent application, No. 581982, was filed by A.C. Cossor Ltd. and F.R.W. Stafford on December 30th, 1942 but there are is no information regarding actual performance.
Browsing through Wireless World, I have found a further article, namely, 'Super-Regenerative Receivers - a reassessment in the light of recent developments' by 'Cathode Ray' in the June 1946 edition that examines principles in some depth.
Certainly, the field of super-regeneration would seem to offer much scope for experimentation and applications are by no means confined to antiquity as some modern car alarm remotes utilise super-regenerative receivers.

J. Bubez
West Sussex U.K.
than it was fed, all caused by earth loops. I recall from my days as an appliance repairman that there were frequently problems with mains hum in cassette decks whose signal and mains earths were connected creating loops with the rest of the stereo system. Earth loops pose serious threats to signal integrity and even to equipment sometimes.

While Ian is not incorrect in his description of skin effect and its role in interference it is irrelevant for most audio applications because the skin depth in copper, for example, is greater than the diameter of most shielded cables over the audio band. This means that the interference punches right through the shield, through the inner signal conductor, and out the other side. The reason shielding still works in these circumstances is that the interfering signal induces (virtually) identical currents and voltages in both the earth and signal conductors, and these currents and voltages cancel. Both the interfering and signal currents will be distributed across the entire cross sectional area of the relevant conductors. Skin depth does not define a sharp cut-off anyway, it is an arbitrarily chosen depth at which the current density has dropped to a particular fraction of that near the surface, and in fact some portion of the current will flow through all areas of the conductor. Superconductors are another matter.

Ian may care to ponder the nature and purpose of the lump in the signal cable most PC monitors these days. It is a ferrite sleeve acting as a one turn common mode choke, and it is intended to suppress the effects of earth loops at high frequencies. I expect this is to reduce radiated interference rather than protect the monitor from interference. Whatever, the purpose is to break high frequency earth loops because they are a major cause of interference. Even in shielded systems earth loops can cause problems, partly because no shield is 100% effective at all frequencies, and the necessity to provide access for assembly and repair, and holes for connectors and feed-throughs, means that most shields have breaks in them.

It should be remembered that most power transformers provide very little isolation at high frequencies due to their high interwinding capacitance. Most mains filters and switched mode supplies include a common mode choke. It is possible a high frequency earth loop will be created through interwinding and other parasitic capacitances even when no hard earth connection exists, and the purpose of the common mode choke is the same, break the loop (or at least increase its impedance). Such chokes have little or no effect on differential mode interference (i.e. between Active and Neutral).

500Mhz sampling front end

You have probably received a fair number of comments from other readers concerning Mr Hickman's interesting and informative article in the June issue. Nevertheless, I thought I would write to you with an observation of my own.

With regard to producing shorter Gate 1 sampling pulses, I suspect that the avalanche pulse generator employed has already reached the limit of its capability in this direction. Some improvement might be indeed be achieved by using a shorter delay line, L1, and a transistor having a higher transition frequency than the BFR91. Unfortunately, this will inevitably be at the expense of pulse amplitude, since most commonly available low cost transistors with higher transition frequencies also tend to have lower avalanche voltages. Therefore, it might be worth considering an alternative method of generating shorter sampling pulses. The method I have in mind is a variation on the theme of the classic step recovery diode (SRD) impulse generator. However, instead of employing an SRD - which is an unusual device that readers are unlikely to find in the majority of mainstream electronic component distributors' catalogues - try using an inexpensive and readily available PIN switching diode. With suitable biasing some short lifetime epitaxial PIN diodes exhibit behaviour very like that of SRD's. For example, Agilent Technologies' HSMP-3820 PIN diode or similar would probably make a suitable candidate for experimentation. In principle it should be possible to generate sampling pulses, having step deficient amplitude, of around 300ps or less using this method.

Douglas R Taylor
By email

More PCBs

Regarding the excellent article on ‘Making single sided PCBs’ by Cyril Bateman in the May issue, you can make a fairly decent prototype by printing the artwork out on standard paper, on a laser printer, reverse image, and then laying the artwork to the copper and using a normal household clothes iron to transfer the image (Hottest setting). Make sure you go over the whole artwork with the iron. Peel off the paper while its still hot, and you will have a reasonable quality artwork that needs to be gone over with a Dalo etch resist pen (This can take a while). Then just etch. I also have used Ruby automask and that ‘Press’n’Peel’ stuff. The Ruby was ok, but as all artworks had to be at 10 times size and the non editing quality of the process, I didn’t think it was used anymore. As for the ‘Press’n’Peel’ ...

Never again.

The easiest and quickest way is to use a laser printer with standard overhead transparency film, specifically for laser printers and use that in the UV exposure box. Double sided PCBs are made simply by taping the two sides of the artwork together, taping the UV sensitive PCB into this sandwich and then exposing both halves separately. Before I bought my UV box, I had kludged together one out of a standard light fitting and a UV tube.

B. Teleki
Newcastle-under-Lyme, U.K.
Ozoniser

In hot and damp climates, fungus and mould can develop in all places but mainly in books. Ozone (O₃) is a powerful oxidizer that kills microorganisms and bad smells in the air.

The quartz bulb inside any mercury vapour lamp emits strong ultra-violet light which energy is enough for the reaction 3O₂ → 2O₃. This circuit is a driver and timer for a 220VAC 125W mercury lamp powered by a 120VAC mains source. The ballast in series with the lamp operates as a current source so the output is not reduced appreciably when operating at 120VAC but a voltage doubler is needed to start the plasma inside the bulb.

Pressing S1 starts the lamp and powers the timer that sends pulses generated by Q2 to trigger Q1 continuously. After the time selected by P1, IC1 output goes high and Q3 shuts down the pulses, cutting the power.

I use this ioniser when leaving for work to avoid any exposure of UV rays, harmful to eyes and skin, and ozone that burns the lungs.

The outside glass bulb must be broken to expose the quartz bulb and trigger connection inside (a wire with a 80kΩ resistor, R2). I installed the bulb inside a piece of plastic tube with tinfoil glued in the inner wall and put a fan in the bottom to disperse the ozone in the room like a fountain.

Be sure to open the windows when you came home again.

_Tiaraju Vasconcellos Wagner_
_Brazil_

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Operation of valve portable radios from the mains

There have been many circuits published for the operation of "picnic case portables" from either battery inverters or from mains power supplies. Most of these suffer from either safety problems if mains operated, or interference problems if of the battery driven inverter type.

This circuit overcomes these problems and is intended to operate any battery set from a mains supply in safety and without interference.

The heart of the circuit is a mains transformer from a video recorder. These transformers are double insulated as most recorders are not connected to mains earth and have several secondary windings ranging from about 4 to 40 volts.

The voltages of the various windings should be measured and one giving about 4 to 6 volts selected for the valve heaters. The higher voltage windings should be series connected to make up about 1/3rd the required HT voltage. This is then voltage doubled and smoothed for the HT supply. The HT current being low (10-15mA) means that this rectifier connection will give nearly 2.8 times the secondary RMS voltage. This voltage is not critical as in practice it varied widely as the HT battery discharged. Simple capacitive smoothing is all that is required.

Voltages up to 150V can be obtained in this manner, The smoothing resistor R2 can also be adjusted for voltage setting.

The LT circuit requires much more care, in terms of voltage control and hum level; an LM317 adjustable regulator is used. This device will only operate down to 1.3 volts, so a diode is connected in series with the output so it operates at 2.1 volts out, 1.4 volts at the filaments. For older sets with 2-volt filaments the diode is omitted. The reverse diode across the regulator is to discharge the output capacitor at switch off. A string of four 1N4002 diodes can be fitted across the output to act as a crowbar in the event of the regulator going short. The unit should have the filament voltage pot set using a dummy load. With good valves this should be 1.3 to 1.35 volts for long life; with older valves that have possibly been overun it is permissible to increase the voltage to 1.45V to achieve acceptable performance. R4 can be adjusted to give the required setting range on the pot R1.

For very old battery receivers that require a grid bias supply one of the spare windings is shown utilised for this purpose. The resistor chain R6,7,8 is adjusted for the required output voltage tappings.

Ed Dinning
Newcastle
The HS801: the first 100 Mega samples per second measuring instrument that consists of a MOST (Multimeter, Oscilloscope, Spectrum analyzer and Transient recorder) and an AWG (Arbitrary Waveform Generator). This new MOST portable and compact measuring instrument can solve almost every measurement problem. With the integrated AWG you can generate every signal you want.

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

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

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

- Measured signals and instrument settings can be saved on disk. This enables the creation of a library of measured signals. Text balloons can be added to a signal, for special comments.

- The (colour) print outs can be supplied with three common text lines (e.g. company info) and three lines with measurement specific information.

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

- The minimum system requirement is a PC with a 486 processor and 8 Mbyte RAM available. The software runs in Windows 3.xx / 95 / 98 or Windows NT / 2000 / XP and DOS 3.3 or higher.

- TiePie engineering (UK), 28 Stephenson Road, Industrial Estate, St Ives, Cambridgeshire, PE17 3WJ, UK Tel: 01480-460028; Fax: 01480-460340

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- Web: http://www.tiepie.nl
5W Inverter

This inverter has been designed with readily available components. The transformer is a standard 10VA mains transformer with two 6V windings connected as shown in the schematic. Its purpose is to provide a suitable voltage for all those mains battery chargers that surround us: mobile phones, electric razors, generic battery chargers and even for a 5W electronic neon lamp. Frequency of operation is between 70 and 190Hz depending on the load. The frequency is not quite the mains frequency but is good enough to supply the intended loads. A small neon light at the output gives an indication of the presence of a dangerous voltage. The circuit will withstand temporary shorts and battery reversals. Some switching chargers require an initial peak current that might look like a short to the inverter. In this case it is necessary to disconnect and reconnect the load until it works. A fuse rated at 2.5A is a useful addition. Reverse one of the windings if the circuit does not oscillate.

D. Di Mario
Milan

Standalone button latch

The conventional way of entering commands from a keyboard employs a scanning encoder occasions when a simple latching circuit is required, independent of complex processors. For two buttons, a pair of cross-coupled NAND gates offers a straightforward solution, but when selection is to be made from any one of say eight buttons, clearly a more versatile method is called for. It is possible to employ a counter, whose clock is stopped when the desired number is reached, but this turns out to be rather messy as it requires a gate for each button. A better method is to use the MM74C922, which is a hexadecimal keyboard encoder with built-in latches, as well as debouncing. To achieve the desired aim of eight illuminated buttons in one row, some rearrangement is necessary and a 3-to-8-line decoder such as a DM74LS138 needs to be added, as illustrated in Fig. 1. The encoder scans the eight momentary push-button switches S0 to S7, using an internal clock of about 6kHz set by C1, though an external clock of up to 10kHz can be applied to pin 5 instead. The debounce time is some 22ms, determined by the value of C2 = 2p2 shown. The decoder sends one and only one of its eight outputs Y0 to Y7 into the low state, and the sinking current is sufficient to illuminate the respective LED. Push-buttons with built-in LEDs are very effective here, and only one current-limiting resistor (R1) is required. Another point to note is that the binary output ABC from the hex-encoder is also available for commands, depending on whether the circuit to be driven wants 8-line or 3-line inputs. In the latter format this circuit has been made up as a sub-board that conveniently mounts behind front panels, with a five-wire ribbon cable (6V, 5V, A, B, C) to the main PCB. The 8-line (Y0 to Y7) version needs a total of ten wires; as it turns out, the same PCB layout can be used, with either a 5-way or a 10-way connector being fitted during the assembly process.

C. J. D. Catto
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Milford Instruments Limited, Tel 01977 683665, Fax 01977 681465, sales@milinst.com
Timer for battery chargers

Many devices with NiCd or other types of rechargeable cell specify a time for charging. This is usually several hours and it is very easy to put a battery on charge and then forget about it.

This circuit was developed at the request of my son who was given a rechargeable strimmer at Christmas that required a charge of 8 hours.

On operating the ‘On/Start’ switch the output is live for a preset period of from 2 to 12 hours, after which it is off. Timing is reset by switching ‘On/Start’ off and then on.

The delay is provided by the ICM7242 Timer/Counter chip, which is connected as a monostable and triggered by switch-on. The Timer drives a TLP3063 optical isolator triac with zero crossing turn-on. A PNP transistor, BC212L, buffers the output of the timer as its maximum sink current is 3mA and the optical isolator needs about 5mA. The optical isolator, in turn, controls the gate of a TIC 226M triac. The maximum current for the TLP3063 is 100mA and this current is possibly sufficient for battery chargers up to about 20 watts, but having a larger triac makes the unit more versatile. For example it could be used to switch a light off in the house when unattended.

A jumper allows timing and switching functions to be tested over a short interval (20 secs to 2 mins).

By simply changing the value of the 470μF capacitor the delay range may be altered.

The low-voltage components were mounted on one PCB and the two triacs on another, with only the led drive connecting the two. The output connector is a panel mounting 13A socket (RS part number 847-455), with the ‘running’ neon indicating when that socket is live. The unit is housed in a 150x90x55mm box.

The timer chip, optical isolator and triac are available from RS (part numbers 264-793, 261-0211 and 649-403) and their Application Notes may be downloaded from the RS site. Other components came from Maplin.

Tony Meacock
Norwich

Ten year index: new update

Hard copies and floppy-disk databases both available

Whether as a PC database or as hard copy, SoftCopy can supply a complete index of Electronics World articles going back over the past nine years.

The computerised index of Electronics World magazine covers the nine years from 1988 to 1996, volumes 94 to 102 inclusive and is available now. It contains almost 2000 references to articles, circuit ideas and applications - including a synopsis for each.

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Signal Wizard 1.6
Real-Time Digital Filter

Signal Wizard 1.6 (formerly RTDF) is a unique real-time audio-bandwidth digital filter system with infinitely adjustable characteristics – all available at the click of a button. It uses a DSP unit that runs the filter and a friendly Windows-based interface that allows you to design and download any kind of filter you like, all within seconds. You don’t need to know about filter maths, DSP or analogue filter design – all you need to know is what kind of filter you want. With Signal Wizard you can do more than specify the gain of the frequency response – you can also specify the phase of any frequency, with a resolution of one hundred thousandth of a degree! If you don’t want to bother with phase, Signal Wizard will design with total phase-free distortion, no matter how complex or sharp the filter is. You are not limited to the design tool interface either - you can also import frequency responses as text files, specifying just magnitude or magnitude and phase. Once you are happy with the design, just download the filter and run it in real-time. Low-pass, High-pass, multiple-band or arbitrary are all possible.

Signal Wizard 1.6 - Key features

- Runs under Windows 95, 98, Me, 2000, NT and XP.
- Multiple pass, stop or arbitrary filters.
- Import mode for arbitrary frequency response.
- Zero-phase distortion or arbitrary phase.
- Rectangular, Bartlett, Hamming, Hanning, Blackman or Kaiser window functions.
- Deconvolution (inverse) or flipped filter options.
- Single (18-bit) or dual channel (16-bit) modes.
- Plots impulse and frequency responses as magnitude, dB, square, root, real, imaginary or phase.
- Extensive filter analysis statistics.
- Animate facility for tap adjustment.
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- Filter module holds up to 16 filters.
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19in. rack-mount monitor drawers
APW Standard Products has introduced the Cyberview family of rack-mounted interface devices for use in its iMserv & Paramount server cabinets and any other 19in practice equipment. The 1U drawers take up the minimum possible vertical rack space and, when flipped up, the screen sits in front of the verticals. An 1U flip-up monitor drawer houses a 15in or 17in TFT/LCD active matrix colour display panel. Suitable for all cabinets from 650 to 1000mm deep, the displays feature a wide viewing angle with resolutions up to 1280 x 1024. Also available are 1U keyboard/display drawers, giving an 84-key keyboard with trackball and a 15in. display. These units can also integrate an 8-port KVM switch, which, when tiered, is stackable to 64-way.
APW
Tel: 01895 237123
www.apw.com

IC maps Gigabit Ethernet into SDH network
Transwitch has introduced a device for full-duplex mapping of Ethernet traffic into the SONET/SDH transport network. The EtherMap-3 supports eight 10/100 Ethernet ports or one Gigabit Ethernet port to deliver a broad range of both SONET/SDH and Ethernet processing functionality. By incorporating both the new standardised link layer framing protocols GFP (Generic Framing Procedure), LAPS (Link Access Procedure for SDH) and LAPP (Link Address Procedure Framed-mode) and the new virtual concatenation (VC) standards, the IC will allow designs to implement private line Ethernet transport and Transparent LAN services in the wide area network. The device incorporates an Ethernet Media Access Control (MAC) function, a buffering strategy, 84-channel VT/TU (Virtual Tributary / Tributary Unit) mapping. VC-3/VC-4 mapping and virtual concatenation.
Transwitch
Tel: 01256 882158
www.transwitch.com

Inductors offer low current resistance
Pulse has introduced a series of inductors for DC-to-DC power supplies that offer both a low direct current resistance (DCR) rating and wide inductance range. Available with inductances from 0.4 to 6.2μH and current ratings from 9 to 73A, the components provide DCR ranges from 0.38 to 1.44mΩ. This allows power loss ratings to be kept to a minimum, 0.16 to 2.25W, placing the inductors among the most efficient on the market.
Pulse
Tel: 0033 84350448
www.pulseeng.com

Memory is bank-switchable
IDT has added to its family of bank-switchable dual-port memory with device speeds up to 200MHz and densities up to 9Mbits. Unlike traditional 9Mbit dual-port devices that rely on multiple internal die, these devices are available in a single configuration. The 36-bit and 18-bit devices feature selectable 3.3/2.5V I/O operations, with the

Motor driver ICs with brake function
Allegro Microsystems has a range of dual full-bridge PWM motor driver integrated circuits featuring a brake function. Each device is designed to control two DC motors bi-directionally and includes two H-bridges capable of continuous output currents of ±650mA and operating voltages to 30V. Motor winding current can be controlled by internal fixed-frequency, pulse-width-modulated (PWM) current-control circuitry. The peak load current limit is set by the user’s selection of a reference voltage and current-sensing resistors. The fixed frequency pulse duration is set by a user-selected external RC timing network. The capacitor in the RC timing network also determines a user-selectable blanking window that prevents false triggering of the PWM current-control circuitry during switching transitions. Two package styles are available: the A3968SA is supplied in a 16-pin dual-in-line plastic package, while the A3968SLE is supplied in a 16-lead plastic SOIC package with copper heatsink tabs.
Allegro Microsystems
Tel: 0033 4505 12359
www.allegromicro.com
3.3V options supporting speeds up to 200MHz and the 2.5V options supporting speeds up to 166MHz. The bank-switchable devices are organised into 64 banks within a common memory array, surrounded by multiplexing circuitry to allow each bank to be accessed by either port. The devices are capable of supporting frequencies up to 200MHz on buses of various widths, frequencies and voltage levels. The dual ports feature separate, independent clocks on each port to support communication between buses running at different frequencies, even with the two ports set at different voltage levels.

IDT
Tel: 01372 366112
www.idt.com

Watch out for miniature crystal
Fox Electronics is offering a miniature watch crystal that measures 7.0 x 1.5mm with a profile of 1.4mm for real time clock (RTC) applications. With a frequency of 32.768kHz, the new FSX327 is optimised for a 12.5pF load capacitance. Frequency tolerance is ±20 PPM at 25°C and frequency stability is ±0.035 ±0.01 PPM over -40 to +85°C. Turnover temperature range is ±20 to ±30°C operating temperature is -40 to +85°C, and storage temperature is -55 to +125°C. Minimum insulation resistance is 500MΩ at 100 VDC and maximum equivalent series resistance is 65kΩ.

Fox Electronics
www.foxonline.com

Voltage reference with 50ppm/°C drift
Texas Instruments has introduced a family of low-dropout, series-mode CMOS voltage references offering accuracy of 0.2%, a SOT23-3 package size, and 500µA output current. Power consumption is 50µA (max). The REF30xx family features 1.25V, 2.048V, 2.5V, 3.3V and 4.096V output voltages. The devices are able to source up to 25mA of output current and provide a supply range up to 5.5V. The references do not require a load capacitor, and are stable with any capacitive load. Unloaded, the devices can be operated on a supply within 1mV of output voltage.

Texas Instruments
Tel: 0449 8161 80 33 11
www.ti.com

WCDMA test for adjacent channel leakage
Rohde & Schwarz is offering firmware for its SMIQ03HD signal generator and FSU spectrum analyser to support ACLR (adjacent channel leakage ratio) measurements of WCDMA signals. According to the supplier, with power amplifiers, adjacent channel leakage must be low, especially on the downlink. For a single-carrier WCDMA signal, the signal generator features ACLR of +77dB in the adjacent channel and +82dB in the alternate channel. Compared to previously available performance, that means 7dB more dynamic range and beside SCPA's (single-carrier power amplifiers), producers of basestations are making increasing use of MCPA's (multicarrier power amplifiers) for up to four channels.

Rhode & Schwarz
Tel: 01252 818888
www.rsuk.rohde-schwarz.com

STS-3/STM-1 transceiver with 311MHz clock
TDK Semiconductor is offering a SONET/SDH line interface unit which operates at 155.52Mbit/s (STS-3 or STM-1) rates and provides a synchronized clock for backplanes operating at 311MHz speeds. The 78P22S4 interfaces to a 75Ω coaxial cable using CMI coding and provides all necessary transmit and receive circuitry to interface to a digital framer.

TDK Semiconductor
Tel: 002 8443 7061
www.tdksemiconductor.com

FPGA is a live system power-up
Actel has announced availability of a "live-at-power-up" 72,000-gate anti-fuse FPGA for radiation-intensive applications, such as low-Earth orbiting satellites and deep space probes.

P47 Power backplane
Schroff has expanded its range of power backplanes with P47 connections for Compact PCI systems. The backplane supports the connection of up to four power supply units in parallel, with separable fault signals FAL# (Fail) and DEG# (derating of outputs). This allows for a higher-level monitoring unit to carry out logical operations on the signals before they are forwarded to the CPU. This can be used for studying the monitoring of redundant power supplies in high-availability systems at prototype stage. The System Management Bus as specified in PICMG 2.09 is integrated on the board, so with the definable geographical address of each slot, information on the status of each power supply in the system can be monitored at a higher level.

Schroff
Tel: 01442 240474
www.schroff.co.uk

The RTXS-S family uses hardened latches, which the firm says eliminates the need for software-based triple module redundancy (TMR) and maximises the total number of logic gates available to the designer. The RTXS-S devices offer total ionising dose
performance in excess of 100Krad; inherent single-event latchup immunity; greater than 63MeV-cms/mg single-event upset performance; and hot-swap compliant I/Os and cold-sparing capabilities. The family ranges in density from 32,000 to 72,000 typical gates (16,000 to 36,000 ASIC gates).

Acal Power Solutions has introduced a family of 10W DC-DC converters manufactured by IBEK of Switzerland and packaged in a 2 x 1in metal case measuring only 10.5mm high. The converters are available with input voltage ranges of 9 to 18, 18 to 36 or 36 to 75V DC with output voltages of 3.3, 5, ±5, 12, ±12, 15 or ±15V DC. Typical output voltage noise at 20MHz bandwidth is only 60mV p-p. Continuous no-load and short-circuit protection are provided as standard and a shutdown function is available as an option.

Acal Power Solutions
Tel: 01252 858727
www.acalelec.co.uk

Audio playback DAC with SACD interface
Analog Devices has introduced a single-chip stereo digital audio playback design which comprises a multi-bit sigma-delta modulator, digital interpolation filters and a continuous-time differential current output DAC. The audio DAC includes a separate Super Audio CD (SACD) bit-stream and external digital filter interface. The AD1955 supports a 24-bit, 192kHz sample rate and provides 123dB of dynamic range using its mono mode and is fully compatible with all known DVD audio formats, said the supplier. The 5V chip also is backwards compatible, supporting 50/15µs digital de-emphasis intended for "redbook" compact discs, as well as de-emphasis at 32 and 48kHz sample rates. It has a 120dB specified signal-to-noise ratio and 120dB of dynamic range (both not muted at a 48kHz sample rate, A-weighted stereo).

Analog Devices
Tel: 01932 266000
www.analog.com

Smartcard goes remote with IDC termination
Targeting applications such as set top boxes and digital encryption, board-mount smartcard connectors with IDC termination from Tyco Electronics are designed to allow for the smartcard slot to be located remotely from the main board. The smartcard connector features a ribbon cable with strain relief, various cable lengths and versions offering 8 and 16 contacts.

Tyco
Tel: 020 8954 2356
www.tycoelectronics.com

Dual cathode varactor of 15pF capacitance
For applications needing close tuner diode matching, Zetex has introduced dual common cathode hyperabrupt varactors. Two devices, the ZMDC831BTA and ZMDC832BTA offer high tolerance CV characteristics, low leakage and an accordingly low phase noise performance. Nominal capacitances for the 831B and 832B are respectively just 15pF and 22pF for a reverse bias voltage of 2V and a frequency of 1MHz. Reverse voltage leakage current is typically as low as 0.2nA. A maximum footprint of 2.2mm by 2.2mm is required by the component's SOT323 outline. Typical applications for the ZMDC dual varactors include voltage controlled oscillators and tuned phase lock loop circuits.

Zetex
Tel: 0161 622 4444
www.zetex.com

Instrumentation amp on a 2.7V supply
Linear Technology has introduced the LTC2053, a zero-drift instrumentation amplifier that features rail-to-rail input and output, works on a single 3V supply and is available in the tiny MSOP-8 package. It has a maximum of 10µV offset voltage, a 50nV/°C offset drift and a high common mode rejection ratio of 116dB, which is gain independent. According to the supplier, this level of DC accuracy exceeds the precision specifications of instrumentation amps that until now have been only available in the bigger DIP and SO packages and require dual supplies to operate.

Linear Technology
Tel: 01276 677676
www.linear-tech.com

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Zetex
Tel: 0161 622 4444
www.zetex.com

Fan specialist Papst has added product specifying tools to its web site. Working alongside the existing pressure and airflow unit converter tools, the Airflow and Pressure Drop Calculators are designed to help engineers specify the company's fans. The objective of this tool is to obtain an initial estimate of what airflow a fan needs to produce, and to deduce what back pressure the fan must overcome to eradicate excessive heat from a system. To establish what back pressure the fan needs to overcome to deliver the required flow rate, the user then enters details of the relevant aperture size. By clicking on 'calculate' the result is displayed as a value in m² and is the total available open area for air to travel through.

Papst
Tel: 01264 333388
www.papstpic.com

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Power resistors in a chip
Welwyn Components is finding applications for its ranges of standard and custom surface mount resistors in the design of DC-DC converters, where the drive efficiency is placing heavy demands on the specifications of components such as chip resistors. The thick film PWC series (Pulse Withstanding Chip), is available in four standard sizes from 0805 to 2512, it offers a resistance range from 1R0 to 10MΩ, tolerance to 0.5 per cent and typical TCR of 100ppm/°C. Its special design permits an enhanced power rating (1.5W at 70°C for 2512) and higher Limiting Element Voltage (500 for 2512). The PCR series of precision chip resistors offers any resistance value within a specified range of 10R to 1MΩ, at a tolerance of 0.1% and TCR of 50ppm/°C.

Welwyn Components
Tel: 01489 583858
www.welwyn-llt.co.uk

Dual-port comms RAM is 9Mbit
Cypress Semiconductor is offering a 9Mbit dual-port RAM. The CY7C0853V provides 9Mbit of synchronous, pipelined dual-ported memory capable of buffering large packets of data between two independent clock domains. Configured as a 256k x 36-bit wide device, it provides up to 9.6Gbit/s of bandwidth and allows for interface to wide busses. Unlike alternative bank-switchable devices this is a true dual-port, providing simultaneous read and write access to any cell in its memory array from either of its two ports. In addition, the two ports may operate at independent clock speeds, allowing complete decoupling of the devices being interfaced. The devices are available in a 172-pin BGA package at up to 133MHz.

Cypress Semiconductor
Tel: 01707 378799
www.cypress.com

Testing ADSL loop in the field
The LX100 from Yokogawa Martron is a portable test tool for the field troubleshooting of ADSL services over copper cable. The unit displays test data required for effective troubleshooting, including attenuation, noise, TDR measurements, burst noise waveform and complex impedance. Applications include verifying the signal/noise margin necessary for ADSL services, determining the locations of loading coils and bridge taps, estimating the source of crosstalk noise and burst noise, and impedance measurement. It can measure noise down to low levels (~140dBm/Hz), and will carry out measurements on attenuation levels of up to 100dB. There is an auto test mode and the instrument is fitted with a PC compatible PCMCIA memory card slot.

Yokogawa Martron
Tel: 01494 459200
www.martron.co.uk

Controller for Pentium 4
Semtech has announced the SC1474 dual-phase power supply controller to supply both V(core) and VID voltages for the mobile Intel Pentium 4 processors. It delivers the 0.600V to 1.750V core voltage at up to 40A, and the 1.2V, 300mA VID power. The core voltage is set by a 5-bit DAC accurate to 0.85 per cent. The dynamic current-sharing feature automatically balances the average current in each phase, eliminating hot spots caused by mismatched trace impedance and component tolerance variations, said the firm. A linear regulator controller delivers the 1.2V, 300mA power.

Semtech
Tel: 02380 769008
www.semttech.com

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August 2002 ELECTRONICS WORLD
An Electronic Universe

Inability to explain fundamental concepts to a wide audience leads to a severe problem in communication: to being regarded as a 'technician' who hides ignorance behind jargon. The ability to design circuits, but inability to explain everything, causes a frustrating lack of self-confidence for engineers in the boardroom. Jargon without clear explanation leads to shunning by a society which doesn't appreciate mere description of crucially important phenomena, e.g., 'capacitance' and 'inductance', and who want proper underlying explanations. So electronics, as jargon-dominated trivia, is being left out of newspapers and TV, despite the increasingly important reliance of society and science upon electronics.

History
A century ago, 'electronics' was the name of the latest and most prestigious science. But the researchers ended up in chaos, with Ampere's original theory of 'current' finally culminating in the calculated typical 1 mm/s flow of drifting 'electrons', versus Heaviside's 300,000 km/s transverse electromagnetic (TEM) energy wave (whose exact speed, like the local speed of light in a medium, is determined solely by the dielectric insulating material, such as air or plastic, between the conductors, not by the nature of the conductors themselves). This particle-versus-wave problem was not a new problem; it had its roots originally in 1680 when Christian Huygens proposed that light is waves, in direct opposition to Isaac Newton's particle theory. Eventually, in 1927, Niels Bohr invented a 'correspondence principle' to suppress critics by accepting 'particle-wave duality', permitting whichever calculation was appropriate for the problem in hand. Consequently, explanations became submerged by semi-empirical equations, while experimental electronics applications flourished.

It is obvious that even if the entire mass of the cable was electrons, they would carry negligible kinetic energy travelling at 1 mm/s (since the kinetic energy equation is $E = \frac{1}{2}mv^2$). Hence, a 1 mm/s electron current cannot be the predominant mechanism of energy transfer. Ivor Catt (b. 1935) started developing TEM wave-based explanations with David Walton and Malcolm Davidson in May 1976, and published them between 1978-88 as Wireless World articles, which unfortunately were produced in an abstruse manner (absurdly rejecting electric current and displacement current out of hand using Ockham's razor, without including a proper replacement theory or using the new facts which they established to produce an understanding of the unanswered problems in science). Continuity of electric current in a circuit: a Science Fiction Story
Once upon a time, everyone grasped the basic law of

electric currents that currently only flow in complete circuits. It was a simple theory, which was consistent with the known facts.

Sadly, it was a misleading and false theory, because the electric current cannot know if there is a break in the wire at one point until it arrives there, travelling at the speed of light for the dielectric.

Whenever any cable is connected to a power source, the power source will deliver power to the cable, because it has no way of telling whether there is an open circuit or a load at the other end. Only when the electric energy arrives at a break, is the circuit proven open. In the intervening period, electric energy flows at 300,000 km/s as if there is no break. So electric current will flow in an open circuit.

It is important to stop at this stage, and carefully examine what happens in the cable that has been carrying electric energy towards the unconnected (open circuit) wire ends of the cable. First, the cable itself acquires an electric charge (like a pair of capacitor plates connected to a power source). Due to the electric charge, an electric field occurs between the wires of the cable. Second, when the electric energy arrives at the break in the circuit, it has no place to go except to bounce back, which it does, always at the speed of light.

When we close the switch and energy goes off into the open-ended cable at the speed of light for the dielectric between the wires, not knowing that an open circuit exists at the end of the cable:

1. Ohm’s Law is violated because, in his equation $V = IR$, R is the circuit resistance, which is infinity if there is an open circuit.

2. Kirchhoff’s First Law is violated since the law says electric current requires a complete circuit.

Both these problems arise because these old Laws assume instantaneous action at a distance, i.e., that the electricity knows whether or not it faces an open circuit before it even sets off at the speed of light when the switch is closed! Ivor Cutt’s research in computer circuits disproved such nonsense.

Oliver Heaviside around 1875 corrected Ohm’s Law by adding to resistance the term Z, which is the impedance of the dielectric used in the cable. If there is nothing between the wires in the cable, Z is the impedance of the fabric of free space (vacuum), 377 ohms.

The corrected version of Ohm’s law reads: $V = I(R+Z)$. If there is no resistance, Ohm’s law becomes $V = 1Z$.

Hence, any 377 volt source will initially send a 1 amp electromagnetic pulse (EMP or transverse electromagnetic wave, TEM wave, depending on preference) travelling at 300,000 km/s (if the dielectric is vacuum), into a pair of wires it is connected to, regardless of whether there is a load or an open-circuit at the other end.

A consideration of what happens when the 1 amp of energy reaches the open circuit and reflects back, is the basis for the ingenious calculation (below) by Cutt, Davidson, and Walton. This proves that “a capacitor is a transmission line”, i.e., that electric current as presently taught in electronics and physics, is an old deception which needs replacement by the new theory presented below.

In the Dec 1978 issue of *Wireless World*, p 51, Ivor Cutt, Malcolm Davidson and Dr. David S. Walton produced the most original and brilliant theoretical calculation in electronics since Maxwell’s day: they calculated the real mechanism of charging of a pair of wires (open ended power transmission line) through a resistor by 300,000 km/s energy being delivered to it, with the energy bouncing back and forth as it charged up, giving a mathematical formula exactly the same as that empirically found for a charging capacitor. We hereby set out clearly their basic mathematical proof of “a capacitor is a transmission line” and that ‘static’ electricity is indeed in constant c speed motion:

1. Because the pair of open ended wires being charged up through resistor R are in open circuit, their impedance is that of free space. $Z = 377\Omega$.

2. When the switch is closed sending energy at potential V volts through the resistor into the wires, the voltage of the energy in the wires is $VZ/(R+Z)$, which will move at the speed of light for the dielectric (air, vacuum, plastic, or whatever) between the wires.

3. When the energy arrives at the open or loose ends of the wires, it will bounce back at the same speed, colliding with more incoming energy which is continuously arriving at potential $VZ/(R+Z)$. This adds to the incoming energy potential (since electric fields are scalar, direction does not matter in voltage contributions). This gives the pair of wires $2VZ/(R+Z)$ volts.

4. If the length of the wires is x, and the speed of light c, then the number of 2-way passes of the light speed energy in the wires in time t will be simply: $n = c/t(2x)$. 5. Each additional reflection at each end of the wires, continues to increase the voltage potential of the existing potential, although due to the difference between R and Z the increase will be by decreasing amounts, since the differential increase on the n number 2-way pass will be: $2((R-Z)/(R+Z))n.VR/(R+Z)$. 6. Summing (with a geometric series) all the contributions from n reflective passes of the energy up and down the wire while energy is being put in continuously with potential $VZ/(R+Z)$, gives a total voltage in the wires of $V[1-((R-Z)/(R+Z))]$. 7. In the simple case, R is much larger than Z, so that R >> Z. 8. Since R >> Z, it follows that as n becomes very large (as it will do very, very quickly, since the speed of the energy is nearly 300,000,000 m/s), the voltage formula reduces to simply: $V[1-e^{-2ZdR}]$. 9. Since we have shown (in step 4 above) $n = c/t(2x)$, the voltage at time t is $V[1-e^{-2ZdR}]$. 10. The term in the exponent above, $cZ/d = 1/C$, where C is capacitance of the pair of wires, so we arrive at the standard result for a charging capacitor: $V[1-e^{-2\pi fRC}]$. Hence Cutt, Davidson and Walton discovered the correct mechanism of electricity, proving that both ‘static’ and current are continuous 300,000,000 km/s electromagnetic energy flows and showing that the traditional exponential charging formula for a capacitor is merely an approximation to the numerous small steps of bouncing 300,000 km/s TEM wave energy which actually occur in the real physical process.

All ‘Static’ Charge is Oscillating 300,000,000 km/s Standing Waves of Electromagnetic Energy

1. “Energy can only enter a capacitor at the speed of light.”

2. “Once inside, there is no mechanism for the energy current to slow down below the speed of light.”

3. “The steady electrostatically charged capacitor is indistinguishable from the reciprocating, dynamic model.”

4. “The dynamic model is necessary to explain the new feature to be explained, the charging and discharging of a capacitor and serves all the purposes previously served by the steady, static model.”

(I. Cutt, Electromagnetism 1, Westfields Press, St. Albans, 1994, p 5).

In addition to this proof that the capacitor is a transmission line, the same thing was done for the inductor, treating it as square-shape for simplicity of calculation, with a lot of maths solved by a computer program by Ivor Cutt and Michael S. Gibson. The basic
concept is a bit like the charging capacitor, but there is cross talk between the adjacent windings of the inductor coil so that: "The inductor is a time-delay and energy trap. A voltage step enters and travels back and forth through the device, with gradual trapping of energy inside." The computer iteration solution gave a lot of small steps which shows that the correct (experimentally known) exponential induction curve is just an approximation to the c speed energy flow physical mechanism and was the proof that was published by Catt and Gibson in Proc. IEEE, vol. 75 (1987), p 849.

Ivor Catt also did the analysis for a simple oscillator circuit, containing a capacitor and inductor. Traditionally, the circuit is analysed by equating the potential (voltage) across the capacitor with that across the inductor: 
\[ v = (1/C)\frac{d}{dt} \int L \frac{d}{dt} \] where C is capacitance and L is inductance. (These terms come from Maxwell's "displacement current" formula for a capacitor, i = C dv/dt, and the Faraday equation for self-inductance by a coil of wire or inductor, i = -L dV/dt.) Differentiating gives an accelerating current equation, d2i/dt2 = -i/LC. This is then solved as a case of simple harmonic motion, giving the sine wave voltage variation curve. sin (ωt), where ωt2 = 1/LC.

The problem with this traditional analysis is that, as Catt states, it: "assumes that when current is switched into the inductor, it appears instantaneously at all points in the inductor; the use of the single, lumped quantity L implies this. Similarly, it is assumed that the electric charge density at all points in the capacitor is the same... Work on high-speed logic systems led to a reappraisal of the conventional analysis." Ivor Catt's reappraisal of the oscillator circuit on the basis of real c speed energy flow, shows that the conventional sine wave solution is only an approximation to the reality, which is a large series of small steps due to c speed energy reflections in the circuit, in which the capacitor behaves as an open-circuit transmission line, while the inductor behaves as a short-circuited transmission line. Catt showed that the underlying mechanism is that the bigger the values of the capacitor or inductor, the smaller is each bouncing pulse of current between the capacitor and inductor, so more time passes while the capacitor charges and discharges, thereby reducing the 'resonant frequency' of the circuit. Catt published the full mathematical proof in Proc. IEEE, vol. 71 (1983), p 772.

Experimental Proof from the Discharge of a Charged Cable into an Oscilloscope: "A one metre section of 50Ω coaxial cable was charged up to a steady 10 volts via a 1 MΩ resistor, then suddenly discharged into a long piece of coax. A 5-volt pulse 2 metres wide was found to travel off at the speed of light for the dielectric. The voltage was half of what one would expect. It appears that after the switch was closed, some energy must have started off to the left, away from the now closed switch; bounced off the open circuit, and then returned all the way back to the switch and beyond."

"This paradox is understandable if one postulates that a steady charged capacitor is not steady at all; it contains energy, half of it travelling to the right at the speed of light, and the other half travelling to the left at the speed of light. Now it becomes obvious that when the switches are closed, the rightwards-travelling energy will exit first, immediately followed by the leftwards-travelling energy after it has bounced off the open circuit. Any apparently steady field is a combination of two energy currents travelling in opposite directions at the speed of light."

(1. Catt, Electromagnetism I, Westfields Press, St. Albans, 1994, pp 13-14, condensed here.)

The Nature of the Electron as Derived from Catt's Results

The above experimental proof, conducted by Ivor Catt when working out the theory of mutual inductance (cross-talk) in computer circuits while at Motorola, Phoenix, in the 1960s, leads to the question of what happens to the magnetic fields from each opposing component of the c speed energy oscillating in the capacitor plates. The answer is that the magnetic fields are vectors which curl in one direction around the direction of the energy flows and since there is equal energy flow in each possible direction in 'static' electricity, the magnetic fields from each equal and opposite energy flow cancel each other out exactly, while the scalar electric fields simply add up.

It is interesting to consider what we mean by 'cancel out'. Do the two components of the magnetic field magically dematerialise energy by disappearing (thereby breaking the law of conservation of energy)? Or is the cancellation just a superposition of fields that cannot be measured by a compass needle for the reason that the compass needle is equally pilled in two opposite directions?

The answer can be found by calculating the total electric energy of a capacitor, and seeing whether this is the complete energy, or whether the total input energy shows that there is also an unobserved magnetic field present in all 'static' electric charge. The capacitance of a pair of wires is C = QV, where Q is electric charge on either conductor (each conductor having equal and opposite charge), and V is the potential difference (voltage) between the charged wires. The electric energy stored in a capacitor is E = (1/2)CV2, whereas the magnetic energy is E = (1/2)LI2, where L is the self-inductance of a 2-wire power cable.

[Since electromagnetic energy in a capacitor has half its energy in magnetic energy and half in electric field, E = (1/2)CV2 = (1/2)LI2, so CV2 = LI2, which upon employing Ohm's law as Z = V/I proves that Z = (LC)-1/2, and L = C2. These very useful results also apply to a transmission line since "a capacitor is a transmission line".]

The problem is that when we measure the energy going into the capacitor, we only usually measure the electric energy, not both electric and magnetic energy, even though every wire carrying a new energy flow has a measurable corresponding magnetic field around it. If we measure the electric plus magnetic energy supplied to the capacitor, it is E = CV2, exactly double the electric field energy in the charged capacitor! Hence, half the energy in the capacitor must be present unobservable magnetic fields with opposing curls from each equal and opposite 300,000km/s energy flow.

This factor of (1/2) difference also occurs when comparing the equation for kinetic energy, E = (1/2)mv2, with Einstein's total energy equation for mass, E = mc2. This analogy between electromagnetic energy in capacitors and energy in general physics is not a coincidence. By reduction of the previous capacitor situation down to a unit charge, we see that every apparently 'static' charge in the universe has an effect, a charged capacitor plate with electromagnetic energy oscillating in all directions at speed c. From this, we see that the individual electron is, as Catt's experiment proves, "a standing wave of energy" (Catt, private correspondence). Furthermore, Catt in a January 1986 Wireless World article points out that a standing wave (sin wave) is a "camouflaged circle:"

From this, I argue the nature of an electron using Catt's findings: an electron, as a unit charge, is a pulse of pure electromagnetic energy going around in a tiny circle (due to the inverse-square nature of gravity becoming very great
on a tiny distance scale), so the electromagnetic energy is bent into a circular orbit due to its own mass, $m = E/c^2$ (from $E = mc^2$). Remember, light has no 'rest mass' because light is never at rest, but light does have transit mass. We thus find that the electron is a spinning electromagnetic 'black hole' of radius $R = 2Gm/c^2 = 2E/c^4$. Since the effective gravity of a loop can be calculated on the basis that the entire mass of the loop is located in its centre. See Newton’s Principia for an ingenious geometric proof of this (Newton proved that the gravity of the Earth can be calculated correctly by treating the mass as all being located at the centre).

This model of the electron has a spherically symmetric electric field at large distances compared to the electron radius $R$, since the electric field lines are scalars radiating outwards equally in all directions at right angles to the loop at each point on the loop, but it has an asymmetric magnetic field due to the fact that the magnetic field loops around each point on the electron loop, creating a toroid or ring doughnut-shaped magnetic field which at long distances is a dipole magnet, hence the known magnetic moment of the electron. The spinning of the electron ring at speed $c$ explains the spin of the electron as utilised in quantum mechanics to explain the anomalous Zeeman effect (spectral line splitting when the emitting atoms are in a magnetic field). The reason why most atoms are non-magnetic is the Pauli exclusion principle, which forces every electron in the atom’s electron shells to have an opposite spin compared to its neighbours. This results in the magnetic fields normally cancelling each other in the sense of producing an unobservable net magnetic field, although orbital variations in the electron shells or orbitals of some elements do produce a slight net magnetic field due to the asymmetry of a small proportion of electrons in the material. This effect produces our magnets.

**Problems in original Wireless World presentation**

The lack of application of Catt, Davidson and Walton’s work from electronics to general science (including derivation of Maxwell’s equations, quantum mechanics, fundamental particle physics, relativity, mechanisms of fundamental forces and their inter-relationships, etc.), led them into a wilderness of suppression - akin to the famed Aristarchus of Samos who discovered the solar system theory in Ancient Greece, but was ridiculed and suppressed for nearly two thousand years until the theory was developed in detail by people who appreciated its value. It is important to note that some of the inferences of Catt, Davidson and Walton were misleading in matter of detail. For example, they disastrously asserted (December 1980) that there is "no electrical current", while what they actually prove is that energy is normally propagated by transverse electromagnetic mechanism, not by electron drift, and that capacitors charge, store energy and discharge at the speed of light, with no mechanism for the stored energy to slow down below that speed therefore proving that apparently static electrons have in fact speed-of-light oscillating speed and are TEM waves. Although this is correct, and proves that static charge or normal electrical energy transfer does not comprise of 1 mm/s electron drift, it does not prove the existence of electric currents in other circumstances, and electrons can be lost from a circuit due to electron emission and chemical reactions, so a drift current can in fact actually exist, although as Catt, Davidson and Walton assert, electric current is not the mechanism of energy transfer in electricity. This 1 mm/s electric current is to the 300,000 km/s TEM wave of electron spin and orbit at 90 degrees to each other. what the 1 m/s mild air breeze is to the 500 m/s air molecule bombardment speed.

The real issue is whether the concept of electric current, as the number of Coulombs of electric charge passing a point in a circuit each second, is really applicable to mains AC power supply, where the net drift of electrons is zero! Clearly this calculation and the whole concept in such a situation is in serious error and we should be careful not to apply the concept of ‘electric current’ or its calculation in Coulombs/second to mains AC electricity, since applying such a scalar equation to a vector situation where the resultant is zero will evidently give a completely false answer. What we must do instead is to refer to mains AC as ‘electric power’ not ‘electric current’, and measure the electric power in Watts (Joules/second of energy). It is important that this is not obvious: it is an analogy to the situation in physics where ‘weight’ and ‘mass’ were not distinguished for centuries and even Cavendish, when first determining the mass of the earth, called his experiment “Weighing the Earth”. Today, students are banned from doing this because the important distinction between weight (which is force) and mass (which is matter) is finally appreciated. We should therefore not belittle Catt, Davidson, and Walton for dismissing the 1 mm/s ‘electric current’ from situations where it is not applicable!

In regard to the original dismissal of “displacement current” by Catt, Davidson, and Walton, they failed to distinguish that what they were dismissing was Maxwell’s physical interpretation of displacement current, not Maxwell’s mathematical equation of it. Subsequently, Professor D.A. Bell, writing in the August 1979 issue of Wireless World, headed his article “No Radio Without Displacement Current”, and showed that radio transmission involves the mathematical equation for current being equivalent to the rate of charge of electric field (multiplied by the appropriate electromagnetic constant), i.e., so-called displacement current. The general problem with Catt, Davidson, and Walton’s research presentation was the lack of careful restriction of their discoveries to the area in which they were proven to be valid. If they had carefully stated that they were only dismissing electric current where electrical energy flow in transmission lines and capacitors was concerned, they would have avoided producing confusion in their ignorant readers and would have avoided giving the scientific world an excuse to argue that their discoveries were incompatible with well-established facts such as electron motion in vacuum TV picture tubes which implies an electron drift current in the cathode supply wires due to electron loss.

Part 2, ‘The Electronic Big Bang’, will be published later.

**Bibliography**


Catt, Ivor, Electromagnetism 1, Westfields Press, St Albans, 1994.


Keyboard input for PIC projects

For many PIC microcontroller based projects one of the design problems that needs to be resolved is how to input commands or set up information to a PIC program.

Writing a few links to a spare port of the microcontroller that is read on reset, or a few switches that are scanned when the PIC software is running may be sufficient. However, a low cost alternative is to use a standard PC keyboard. These keyboards cost only a few pounds and it is an input device that we are all familiar with. As a bonus there are three LEDs that can be controlled by the PIC program to show program status.

Within the article all data generated by the keyboard is given in hexadecimal 'NN'h form to distinguish between data and key characters such as function key F1. The PIC keyboard software was written for the 16F877 microcontroller but should work with most PIC microcontrollers, however only the 16F877x and 16C74 family has the built in serial port used for testing.

AT keyboard

All current PCs are supplied with an AT style keyboard that have a PS/2 type connector. The keyboard was designed by IBM to be software configurable so that there is no need to manufacture different keyboards for different countries. Only the key tops need changing between countries not the keyboard circuit. This software flexibility allows keys to be added. For example, recent addition of the Euro currency key (£), and some keyboards now include dedicated internet browser keys.

Keyboard internals

Internally these low cost AT keyboards consist of the keys sitting on a moulded clear rubber mat, this mat is placed on top of two plastic sheets with conductive circuit tracks printed on them. This conductive pattern is a 22 by 6 matrix where pressing down a key will make the connection between the two layers at a unique intersection. The keyboard controller continually scans this matrix and determines which key position has been pressed and sends this data to the PC.

The keyboard controller board is a small single-sided printed circuit board consisting of a surface mount controller (hidden under black protective coating), a few discrete components, 18 wire links and the three keyboard LEDs. Figure 2 shows the keyboard viewed from underneath, for clarity the two conductive sheets have been removed but they connect to the edge connector at the top of the printed circuit board.

Power supply

The keyboard will work off a 5-volt supply, so the same supply can power both the PIC circuit and keyboard. However the electrical characteristics sticker on the base of my 'Ever Green Touch' keyboard (manufactured in China) states that it requires 5V at 170mA. It is hard to imagine that a single customised controller chip requires all this power so I measured the current and found that it was only 8mA, and with all three LEDs on the keyboard consumed a total of 20mA. This is many times...
what the PIC microcontroller consumes, but if you are considering a battery powered application, then the current the keyboard requires will need to be taken into account.

Keyboard controller
The original keyboard design had a single chip microprocessor, but now a customised controller chip is used. This keyboard controller chip takes care of all keyboard matrix scanning, key de-bouncing and communications with the computer, and has an internal buffer if the keystroke data cannot be sent immediately. The PC motherboard decodes the data received from the keyboard via the PS2 port using interrupt IRQ1.

The one thing that these keyboards do not generate is ASCII values. With a typical AT keyboard having more than 101 keys, a single byte could not store codes for all the individual keys, plus these keys along with shift, control, or alt, etc. Also for some functions there is no ASCII equivalent, for example 'page up', 'page down', 'insert', 'home', etc.

When the keyboard controller finds that a key is being pressed or released it will send this keystroke information, known as scan codes, to the PIC microcontroller. There are two different types of scan codes - make codes and break codes.

make code
A make code is sent whenever a key is pressed or held down. Each key, including 'shift', 'control' and 'alt', sends a specific code when pressed. Cursor control keys, 'delete', 'page up', 'page down', 'ins', 'home' and 'end', send extended make codes. The make code is preceded by 'E0'h to indicate an extended code. The only exception is the 'pause' key that starts with a unique 'E1'h byte.

break code
A break code is sent when a key is released. The break code is the make code preceded by 'F0'h byte. For extended keys the break code has an 'E0'h preceding the 'F0'h and make code value. The only exception is the 'pause' key as it does not have a break code and does not auto-repeat when held down.

key code
Every key is assigned its own unique code so that the host computer processing the information from the keyboard can determine exactly what happened to which key simply by looking at the scan codes received. There is no direct relationship between the scan code generated by a particular key and the character printed on the key top.

The set of make and break codes for each key comprises a scan code set. There are three standard scan code sets numbered 1, 2, and 3 - stored within the keyboard controller. Scan code set 1 is retained for compatibility for older IBM XT computers. Scan set 3 is very similar to the set 2 but the extended codes are different. Scan code set 2 is the default for all AT keyboards and all scan codes discussed here are from this set.

scan code
If, for example, you press 'shift' and 'A' then both keys will generate their own scan codes, the 'A' scan code value is not changed if a shift or control key is also pressed. Pressing the letter 'A' generates '1C'h make code and when released the break code is 'F0'h, '1C'h.

Pressing 'shift' and 'A' keys will generate the following scan codes:

The make code for the 'shift' key is sent '12'h.
The make code for the 'A' key is sent '1C'h.
The break code for the 'A' key is sent 'F0'h, '1C'h.

The break code for the 'shift' key is sent 'F0'h, '12'h.

If the right shift was pressed then the make code is '59'h and break code is 'F0'h, '59'h.

By analysing these scan codes the PC software can determine which key was pressed. By looking at the shift keystroke the software can distinguish between upper and lower case.

Keyboard commands
The main purpose of the keyboard is to accept typed data and send this information to the host computer, however there are several commands that can be sent to the keyboard controller. Figure 3 shows some of the more common keyboard commands. There are other commands that can be used to change make or break codes for individual keys, but the commands given here are the most useful. The possible keyboard response to these keyboard commands is given in Fig. 4.

Keyboard self test
When the keyboard is first powered up it runs a self-diagnostic test, this test primarily looks for keys that are 'stuck' down. All the LEDs on the keyboard will also briefly switch on and off as part of this self test. When the keyboard is plugged into a PC you may be forgiven for thinking that this was part of the PC start-up sequence as it happens around the same time as the PC is powering up and
also running diagnostic tests.

After running the self-test the keyboard processor sends 'AA'h byte if everything is working correctly. If the keyboard processor finds a fault it will send 'FE'h byte. If the keyboard reports a fault then the PC BIOS will display 'Keyboard error or no keyboard present' followed by the less than useful message 'Press F1 to continue' (!).

**'ED' keyboard LED command**

The keyboard processor does not switch the 'Num Lock', 'Caps Lock', and 'Scroll Lock' LEDs whenever the appropriate key is pressed. Control of these LEDs is done by the host computer sending LED on/off commands to the keyboard processor. The keyboard LEDs and the corresponding keys are independent of each other.

To tell the keyboard which LED to turn on or off, send command 'ED'h and wait for the keyboard to respond with acknowledge byte ('FA'h). Then send the binary number '00000ABC' where the 'A' bit is the state of the 'Caps Lock' LED, 'B' is the state of the 'Num Lock' LED, and 'C' is the state of the 'Scroll Lock' LED. Logic '1' is LED on, '0' for LED off. The keyboard will then respond (again) with 'FA'h indicating that it has successfully received the information.

The most significant five bits in the byte containing the LED information must be zero. If any of those bits is set then the keyboard processor will respond with 'FE'h (error) and wait for a properly formatted byte. There are no mechanisms for asking the keyboard controller the status of these LEDs, if you are using the LEDs and need to know which are on or off then the PIC program will need to store this information.

**'EE'h echo test**

As the name suggests this command echoes back the command value. It can be used as a quick test to make sure that the keyboard is connected and working.

**'FO' set scan code command**

If you want to change to a different scan code set, send 'FO'h command byte to the keyboard. The keyboard processor will respond with 'FA'h (acknowledge). Then send '01'h, '02'h, or '03'h for scan codes sets 1, 2, or 3.

When the new scan code is received the keyboard will again reply with 'FA'h.

To find out which scan code set is currently being used by the keyboard send '00'h instead of a new scan code set number. The keyboard will then respond with scan code number '01'h, '02'h (default) or '03'h.

All the scan codes presented here are those actually generated by the keyboard. When the keyboard is plugged into the PC the BIOS may translate some of these scan codes for compatibility reasons. Consequently a PC program may report slightly different scan codes for some keys.

**'F2'h device identity command**

The keyboard will respond to this command with 'FA'h (acknowledge) followed by the keyboard device type numbers 'AB'h, '83'h. When the keyboard is plugged into a PC the computer needs to know what type of device is connected to which PS/2 port. Other PS/2 devices can also be connected, such as a PS/2 mouse, which will respond with ID number '00'h, '00'h.

**'FF'h keyboard test command**

If the keyboard is wired to the same 5-volt supply as the PIC, then it is possible that the self test result will appear before the PIC microcontroller has initialised, particularly if the PIC power up timer is enabled. If the keyboard is already powered then sending command byte 'FF'h will force the keyboard to reset and run the self-test. This command is acknowledged by the keyboard ('FA'h) before the self test is executed. Alternatively use the 'F2'h command to get the keyboard device id number.

**Typematic**

When you press and hold down a key on the keyboard that key becomes typematic. This means the keyboard will keep sending that key's make code until the key is released. The typematic delay is a short delay between the sending of the first and second make scan code. Typematic rate is how many characters per second will appear after this initial typematic delay. The typematic delay can range from 0.25 second to 1 second and the typematic rate can range from 2 characters per second (cps) to 30 cps.

**'F3'h set keyboard repeat rate**

These typematic values can be changed using the 'F3'h command (set auto repeat rate), send 'F3'h and the keyboard will respond with 'FA'h byte. Then the keyboard waits for the data byte that specifies the auto-repeat delay and rate.

With the exception of the 'pause' key, all keys will auto repeat. The default delay is 500ms and the auto repeat

---

**Fig. 7. Serial data sent from keyboard to PIC, data is read on the falling clock edge.**

<table>
<thead>
<tr>
<th>RXbits</th>
<th>Conv = false</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>2</td>
</tr>
<tr>
<td>3</td>
<td>4</td>
</tr>
<tr>
<td>5</td>
<td>6</td>
</tr>
<tr>
<td>7</td>
<td>8</td>
</tr>
<tr>
<td>9</td>
<td>10</td>
</tr>
<tr>
<td>11</td>
<td>Conv = true</td>
</tr>
</tbody>
</table>

interrupt generated by falling clock edge.

line controlled by keyboard processor.
default is 10 characters per second. It is unlikely that these default values will need to be changed, but there may be circumstances where longer delays are needed to allow the PIC to process information between key presses.

**Keyboard serial data**

The AT keyboard transmission protocol is a serial format, with one line providing the data and the other line providing the clock. The data length is 11 bits with one start bit (logic 0), 8 data bits (lsb first), odd parity bit and a stop bit (logic 1). The clock rate is approximately 10 to 30kHz and varies from keyboard to keyboard.

The communications protocol is bi-directional, but as there are only two lines the handshaking between keyboard and PIC is more complicated. Unusually the keyboard generates the clock irrespective of the direction of data flow. The keyboard communications protocol is a strange mix with elements of both synchronous (separate data and clock) and asynchronous (start/stop bits) data transmission.

Both the keyboard clock and data lines are open collector outputs and require pull-up resistors to +5V. The PIC microcontroller has internal pull-up resistors on Port B which are enabled in the "iniPIC" routine, if the keyboard is connected to another port then external pull-up resistors will be needed.

**How the code works**

The keyboard clock signal is connected to RB0 and used to generate an interrupt on the falling edge. The keyboard data line is connected to PIC port RB1. Running the iniPIC routine initialises the various register options, sets the timer prescaler and initialises the variables. In program keybd.asm, the serial communication port is initialised. The TimerOverflow flag is the TOIF flag of the 8-bit timer 0, this flag is set whenever the timer has counted up to 255 and starts counting again at 0. This flag is used to indicate a timeout and various counts are then automatically cleared. Without this, if the received data becomes corrupt and the TXbits count is wrong, then all following data will be decoded incorrectly. An alternative method if the timer is being used within the application program is to use the watchdog timer.

Variables TXbits and RXbits are counters indicating which bit in the serial keyboard data is being sent or received. The Conv flag is set whenever the data had been received from the keyboard. ReceiveDataFlag is the serial communication RCFIF flag that is set whenever data is received from the PIC via the serial port (keybd.asm only). This value is stored in variable TX and the ToKey routine is called.

**Receiving data from keyboard**

The keyboard will transmit data to the PIC microcontroller as soon as a key is pressed if both the clock and data lines are high, as this indicates idle status. If the clock line is held low by the PIC microcontroller then the keyboard cannot send and the keyboard controller will buffer the keystroke data.

Variable RXbits keeps track of which bit is being received, as RXbits is incremented on each interrupt. Variable keywork stores the bit pattern of the data received from the keyboard. This is achieved by setting the carry flag according to the logic status of the data at port RB1, then using the rotate right PIC instruction to shift the carry bit.
into the keywork variable. If RXbits = 10 this indicates the PIC is processing the parity bit, however this bit is ignored by the PIC program. On receiving RXbit = 11 (stop bit) the Conv flag is set indicating the end of data. Setting this flag causes the routine FromKey to be called from the main program loop. FromKey routine clears the Conv (convert) flag and sends the received keyboard data (contained in variable char) to the PrntHex (print hex) routine in the keybd.asm code.

This PrntHex routine converts the binary data into the ASCII suitable for display. Adding 48 to a binary decimal number converts that number to its ASCII text equivalent, if the number is greater than 9 then adding 55 will convert the hexadecimal number into an ASCII character. The PrntHex routine then calls the SendPC routine. This routine waits for the TXIF flag to be set, this indicates that the serial communications TXREG (transmitter register) is empty. TXREG register is loaded with the char data and this data is automatically transmitted via the serial port to the PC. These routines are not required in keybd.asm.

Sending data to the keyboard
When the PIC microcontroller needs to send data to the keyboard, the routine ToKey is called. ToKey sets the clock

![Fig. 11. Summary of main program loop for keybd.asm.](image)

```plaintext
call iniPIC
loop
  if conv = true then
    begin
      if keydata = 'A' then
        begin
          send LED command ('ED'h)
          wait for ack ('FA'h)
          send LED on (b'00000111')
        end
      if keydata = 'B' then
        begin
          send LED command ('ED'h)
          wait for ack ('FA'h)
          send LED off (b'00000000')
        end
      end
    end
  
  if TimerOverflow = true then
    begin
      if RXbits = 0 then
        begin // not sending data
          RXbits = 0
          Keydata = 0
          TimerOverflow = false
        end
      end
    end
  goto loop
```

![Fig. 12. Summary of interrupt routine.](image)

```plaintext
if TXbits > 0 then
  begin
    if TXbits < 9 then
      begin
        if TX[TXbits] = true then
          begin
            RB1 = '1' // output
            invert parity bit
          end
        else
          begin
            RB1 = '0' // output
          end
      end
    
    if TXbits = 9 then
      output parity
    if TXbits = 10 then
      begin
        make RB1 an input
        RB1 = '1' // stop bit
      end
    
    if TXbits = TXbits + 1
      if TXbits = 12 then
        TXbits = 0
      end
    else
      begin
        RXbits = RXbits + 1
        if RXbits = 11 then
          begin
            keydata = keywork
            Conv = true
            RXbits = 0
            keywork = 0
          end
        else
          begin
            if RXbits = 10 then
              exit // do nothing
            end
          
      end
    end
  
  if RB1 = true then
    begin
      keywork = keywork + '1'
    else
      keywork = keywork + '0'
    end
end
```

![Fig. 13. HyperTerminal screen showing the scan codes when A, B, C, insert, and pause keys are pressed on the keyboard.](image)
line low for 60 milliseconds using timer 0. Bringing the clock line low prevents the keyboard from transmitting data. While the data line is held low the clock line is set to input and the keyboard will start generating a clock signal.

To make a port pin an output a '0' is sent to the TRISB (data direction register), a '1' sets that relevant port pin to an input. Data to be transmitted is output on the clock interrupt and read by the keyboard on the rising clock edge.

**PIC software**

Sending the scan codes to the PIC is a useful demonstration (and functional test) of the keyboard to PIC connection. It allows specific keyboard scan codes to be verified but it is of very limited application. The main function of this software is to use the keyboard as an input device to a PIC microcontroller. Rather than send the scan code to the PC, the scan value should be checked for various scan codes and appropriate data values modified within the PIC application program.

Assembler listing keybd.asm shows a simple method of reading the keyboard scan codes and if specific keys are pressed, then the keyboard LEDs are turned on or off. The program looks for the letter ‘A’ (scan code ‘46’h), when this is pressed all the LEDs are switched on (variable led determines which LEDs are switched on). When the letter ‘B’ is pressed (scan code ‘3E’h) all the LEDs are switched off. All other key presses are ignored. These keyboard keys and which LEDs are activated can be changed, or values changed when specific keys are pressed.

**Testing the interface**

When the PIC is programmed with the keybd.asm code any
make and break scan codes will be sent as ASCII characters to the PIC serial port. This requires the 74LS14 and two resistors to be fitted. A suitable three-wire serial cable to connect the PIC to the PC's serial port will need to be made. The Windows Hilgraeve HyperTerminal (supplied with Windows) program can be used to view these keyboard generated scan codes as they are transmitted by the PIC software as text. The program properties should be set up as follows - direct to com, speed as 57600 baud, 8 bits, no parity, no flow control and one stop bit.

Figure 13 is a HyperTerminal screen showing the self test passed byte followed by the scan codes for letters A (make code = '1C'h, break code = 'FO'h, '1C'h), B (make code = '32'h, break code = 'FO'h, '32'h), C (make code = '21'h, break code = 'FO'h, '21'h).

Followed by the extended scan codes generated when pressing the insert key (make code = 'E0'h, '70'h, break code = 'E0'h, 'FO'h, '70'h) and eight byte extended code when the pause key was pressed (make code = 'E1'h, '14'h, '77'h, 'E1'h, '14'h, 'FO'h, '77'h, no break code).

Figure 14 shows an interactive Windows program displaying the keyboard response to various commands sent to the keyboard from the PC via the serial communications port. The four buttons (reset, keyboard id, echo, and scan code) when pressed will send that particular command to the keyboard and the keyboard's responses can be seen. The three LEDs can be switched on or off and when the button marked 'LED' is pressed this command is sent to the keyboard and the appropriate LEDs should be lit on the keyboard.

**Fig. 17. Keyboard function key scan codes.**

<table>
<thead>
<tr>
<th>key</th>
<th>make</th>
<th>break</th>
</tr>
</thead>
<tbody>
<tr>
<td>F1</td>
<td>'05'h</td>
<td>'00'h, '05'h</td>
</tr>
<tr>
<td>F2</td>
<td>'06'h</td>
<td>'00'h, '06'h</td>
</tr>
<tr>
<td>F3</td>
<td>'04'h</td>
<td>'00'h, '04'h</td>
</tr>
<tr>
<td>F4</td>
<td>'0C'h</td>
<td>'00'h, '0C'h</td>
</tr>
<tr>
<td>F5</td>
<td>'03'h</td>
<td>'00'h, '03'h</td>
</tr>
<tr>
<td>F6</td>
<td>'08'h</td>
<td>'00'h, '08'h</td>
</tr>
<tr>
<td>F7</td>
<td>'83'h</td>
<td>'00'h, '83'h</td>
</tr>
<tr>
<td>F8</td>
<td>'0A'h</td>
<td>'00'h, '0A'h</td>
</tr>
<tr>
<td>F9</td>
<td>'01'h</td>
<td>'00'h, '01'h</td>
</tr>
<tr>
<td>F10</td>
<td>'09'h</td>
<td>'00', '09'h</td>
</tr>
<tr>
<td>F11</td>
<td>'78'h</td>
<td>'00', '78'h</td>
</tr>
<tr>
<td>F12</td>
<td>'07'h</td>
<td>'00', '07'h</td>
</tr>
</tbody>
</table>

**Fig. 18. Keyboard key pad scan codes.**

<table>
<thead>
<tr>
<th>key</th>
<th>make</th>
<th>break</th>
</tr>
</thead>
<tbody>
<tr>
<td>Num lock</td>
<td>'77'h</td>
<td>'FO'h, '77'h</td>
</tr>
<tr>
<td>/</td>
<td>'E0'h, '4A'h</td>
<td>'FO'h, 'F0'h, '4A'h</td>
</tr>
<tr>
<td>+</td>
<td>'7C'h</td>
<td>'FO'h, '7C'h</td>
</tr>
<tr>
<td>-</td>
<td>'7B'h</td>
<td>'FO'h, '7B'h</td>
</tr>
<tr>
<td>Enter</td>
<td>'E0', '5A'h</td>
<td>'FO', 'F0', '5A'h</td>
</tr>
<tr>
<td>0</td>
<td>'71'h</td>
<td>'FO', '71'h</td>
</tr>
<tr>
<td>1</td>
<td>'69'h</td>
<td>'FO', '69'h</td>
</tr>
<tr>
<td>2</td>
<td>'72'h</td>
<td>'FO', '72'h</td>
</tr>
<tr>
<td>3</td>
<td>'7A'h</td>
<td>'FO', '7A'h</td>
</tr>
<tr>
<td>4</td>
<td>'6B'h</td>
<td>'FO', '6B'h</td>
</tr>
<tr>
<td>5</td>
<td>'73'h</td>
<td>'FO', '73'h</td>
</tr>
<tr>
<td>6</td>
<td>'74'h</td>
<td>'FO', '74'h</td>
</tr>
<tr>
<td>7</td>
<td>'6C'h</td>
<td>'FO', '6C'h</td>
</tr>
<tr>
<td>8</td>
<td>'75'h</td>
<td>'FO', '75'h</td>
</tr>
<tr>
<td>9</td>
<td>'7D'h</td>
<td>'FO', '7D'h</td>
</tr>
</tbody>
</table>

**Fig. 19. Keyboard interface circuit diagram.**

**Fig. 20. Components required for keyboard interface.**

<table>
<thead>
<tr>
<th>IC1</th>
<th>PIC 16F877</th>
</tr>
</thead>
<tbody>
<tr>
<td>IC2*</td>
<td>74LS14</td>
</tr>
<tr>
<td>C1</td>
<td>10nF</td>
</tr>
<tr>
<td>C2</td>
<td>1nF</td>
</tr>
<tr>
<td>C3, 4</td>
<td>15pF</td>
</tr>
<tr>
<td>R1, 2*, 3*</td>
<td>470W</td>
</tr>
<tr>
<td>X1</td>
<td>20MHz crystal</td>
</tr>
<tr>
<td>CN1</td>
<td>6 pin mini-DIN (PS/2)</td>
</tr>
<tr>
<td>CN2*</td>
<td>9 pin 'D' serial data</td>
</tr>
</tbody>
</table>

* optional

**Fig. 21. Power interface wiring list.**

|   +5V  | PIC pin 1 (mcl) |
|   +5V  | PIC pin 11 |
|   +5V  | PIC pin 32 |
|   +5V  | PIC2 pin 14 |
|   +5V  | CN1 pin 2 |
|   0V   | PIC pin 12 |
|   0V   | PIC pin 31 |
|   0V   | IC2 pin 7 |
|   0V   | CN1 pin 5 |

**Fig. 22. '9' pin serial communications link.**

R2 - CN2 pin 2 (tx)
R3 - CN2 pin 3 (rx)
0v - CN2 pin 5 (gnd)

**Fig. 23. Wiring of the keyboard 6 pin mini-DIN PS/2 socket - viewed from the solder side.**

| pin 1 | - no connection |
| pin 2 | +5v |
| pin 3 | - no connection |
| pin 4 | - PIC RB1 (data) |
| pin 5 | - 0v |
| pin 6 | - PIC RB0 (clock) |
The ‘AA’h is the result of the keyboard self test, ‘FA’h is the command acknowledgement for the device identity request. The keyboard responds with device type ‘AB’h and ‘83’h. The two ‘FA’h bytes are acknowledgement of the scan code query command and keyboard processor responds with scan set 2. The final two ‘FA’h are for the LED command acknowledge. The program will also show any make or break codes if any keys are pressed on the keyboard. This Windows program (two versions are available, one for Windows 95/98/ME and the other for Windows XP) and the two PIC assembler source code programs (keybd.asm and keybd1.asm) will be available from EW – just email j.lowe@cumulusmedia.co.uk stating which one you’d like.

Construction
The PIC circuit can be built using strip board, the 20MHz crystal can be connected to 5V for correct operation. The two inverters and series current limiting resistor are for the optional PC serial communications. They are not necessary for the keyboard connection. The PIC expects to interface to a serial line driver which in operation would invert the data, as a serial driver IC is not used then the data has to be inverted.

Care is needed when wiring the PS/2 socket – particularly for the power connection. Remember to observe the keyboard self test when the keyboard is plugged into the socket. All the LEDs should briefly flash if the wiring is correct. If not then disconnect the power supply and check the wiring.

Acknowledgements
My thanks to Andrew Thomas for help with the PIC programming.
PIC is a registered trademark of Microchip Technology Incorporated, USA.
Windows is a registered trademark of Microsoft Corporation.

Assembler listings

```
; keybd.asm
; PIC AT-keyboard reader
; Written by Roger Thomas
; MPASM 23 January 2002

__config H'0F02'

TMRO EQU H'01'; timer0
STATUS EQU H'03'; register
C EQU H'00'; carry flag
Z EQU H'02'; zero flag
R0 EQU H'05'; page bit
PORTB EQU H'06'; port B
RB1 EQU H'01'; keybd data
INTCON EQU H'08'; register
IRQ_RB0 EQU H'01';

interrupt
T01F EQU H'02'; timer0
IRQ_RN EQU H'07'; irq
OPT_REG EQU H'01';

register
TRISB EQU H'06'; port B
PIR1 EQU H'0C'; peripheral
RCIF EQU H'05'; serial comm
XCSTA EQU H'18'; serial comm
XRCRG EQU H'1A'; serial comm
TXSTA EQU H'18'; serial comm
SPBRG EQU H'19'; serial comm
TEMP EQU H'20'; irq handler
IRG EQU H'2A'; irq handler
IRQS EQU H'2B'; irq handler
IRQSTKEQU H'2C'; irq handler

CHAR EQU H'2D'; output
RXBITS EQU H'2E'; bit count
TXBITS EQU H'2F'; bit count
KEYDATA EQU H'30'; keybd

org 4 ; interrupt
MOVF IRQ0
SWAPF STATUS,W
BCF STATUS,RP0
MOVF IRQS
MOVF TEMP,W
MOVF IRQSTK
CALL IRQ
MOVF IRQSTK,W
MOVF TEMP
SWAPF IRQ,W
SWAPF IRQW,F
GOTO IRQ4

irq
; if conv = true then
CALL FROMKEY

; if conv = true then
CALL INIPIC

main
MOVLW D'09'
SUBWF D'10'
BTFSS STATUS,Z
GOTO IRQ1
MOVF PARITY,W
MOVF TX
MOVF IRQ1
MOVF TXBITS,W
SUBWF D'10'
BTFSS STATUS,Z
GOTO IRQ2
MOVF RXBITS,W
MOVF TRISB
BCF STATUS,RP0
GOTO IRQ4

port
MOVF PORTB
SUBLW D'10'
BTFSS STATUS,Z
GOTO IRQ3
MOVF PARITY,F
GOTO IRQ4

main1
MOVF PORTB,RB1
BTFSS STATUS,C
GOTO IRQ3
MOVF RXBITS,F
IRQ3 GOTO IRQ4END
IRQ6 MOVF RXBITS,F
IRQ6 INCF RXBITS,F
SUBLW D'11'
BTFSS STATUS,Z
GOTO IRQ7
IRQ5 GOTO IRQ4END
IRQ6 GOTO RXBYTES,F
MOVF RXBYTES,W
SUBLW D'12'
BTFSS STATUS,Z
GOTO IRQ5
MOVF RXBYTES,F

calc
KEYWORD EQU H'31'; keybd
CALL TOKEY

main3 MOVF RCREG,W
MOVF TX
MOVF KEYDATA
GOTO IRQ7

flag
IRQ MOVF TXBITS,W
; end
GOTO IRQEND
IRQ MOVF RXBYTES,W
MOVF RXBYTES,W
SUBLW D'10'
BTFSS STATUS,Z
GOTO IRQ3
```

August 2002 ELECTRONICS WORLD 57
```assembly
MOVWF IRQSTK ;
CALL IRQ ;
MOVF IRQSTK,W ;
MOVWF TEMP ;
SNAPF IRQs,W ;
MOVWF STATUS ;
SNAPF IRQm,F ;
SNAPF IRQm,W ;
RETIE ;
MAIN
CALL INITIC ;
LOOP BTFSS FLAGS,CONV ;
GOTO MAIN1 ;
BCF FLAGS,CONV ;
MOVF KEYDATA,W ;
SUBLW H'1C' ;
BTFSS STATUS,Z ;
GOTO MAIN4 ;
MOVWF H'ED' ;
BTFSS TX ;
CALL TOKEY ;
BSF FLAGS,ACK ;
MOVWF H'07' ;
; end
MOVWF LEDS ;
GOTO MAIN1 ;
MAIN4
BTFSS FLAGS,ACK ;
GOTO MAIN5 ;
MOVF KEYDATA,W ;
SUBLW H'FA' ;
BTFSS STATUS,Z ;
GOTO MAIN5 ;
MOVF LEDS,W ;
MOVWF TX ;
CALL TOKEY ;
BSF FLAGS,ACK ;
CLRF LEDS ;
; if RXbits = 11 then
SUBLW H'32' ;
BTFSS STATUS,Z ;
GOTO MAIN1 ;
MOVWF H'ED' ;
BTFSS TX ;
CALL TOKEY ;
BSF FLAGS,ACK ;
CLRF LEDS ;
; keydata = keywork
MOVF KEYDATA,W ;
SUBLW H'32' ;
BTFSS STATUS,Z ;
GOTO MAIN1 ;
MOVWF H'ED' ;
BTFSS TX ;
CALL TOKEY ;
BSF FLAGS,ACK ;
CLRF LEDS ;
; goto irq7
MOVF KEYDATA,W ;
SUBLW D'10' ;
BTFSS STATUS,Z ;
GOTO LOOP ;
; if RXbits = 10 then
MOVF TXBITS,W ;
SUBLW D'00' ;
BTFSS STATUS,Z ;
GOTO LOOP ;
; if RXbits = 0 then
CLR RXBITS ;
CLR KEYDATA ;
BCF INTCON.TOIF ;
GOTO LOOP ;
; if TxCB = true then
IRQ MOVF TXBITS,W
BTFSC STATUS,Z
CLRF TMRO
GOTO IRQ
; begin
MOVF TXBITS,W
TOKEY BCF INTCON,IRQ_EN
```
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<td></td>
</tr>
<tr>
<td></td>
<td></td>
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<thead>
<tr>
<th>COMPANY</th>
<th>PAGE NO:</th>
</tr>
</thead>
<tbody>
<tr>
<td>CAMBRIDGE MICROPROCESSOR SYSTEMS LTD</td>
<td>45</td>
</tr>
<tr>
<td>CRICKLEWOOD</td>
<td>45</td>
</tr>
<tr>
<td>CROWNHILL</td>
<td>IBC</td>
</tr>
<tr>
<td>DISPLAY ELECTRONICS</td>
<td>38</td>
</tr>
<tr>
<td>JK MICROSYSTEMS</td>
<td>63</td>
</tr>
<tr>
<td>JPG ELECTRONICS</td>
<td>45</td>
</tr>
<tr>
<td>J &amp; N LTD</td>
<td>2</td>
</tr>
<tr>
<td>JOHN'S RADIO</td>
<td>31</td>
</tr>
<tr>
<td>LABCENTRE</td>
<td>OBC</td>
</tr>
<tr>
<td>MILFORD</td>
<td>35</td>
</tr>
<tr>
<td>PICO</td>
<td>35</td>
</tr>
<tr>
<td>QUASAR</td>
<td>11</td>
</tr>
<tr>
<td>TECSTAR</td>
<td>43</td>
</tr>
<tr>
<td>TELNET</td>
<td>IFC</td>
</tr>
<tr>
<td>TIE PIE</td>
<td>33</td>
</tr>
<tr>
<td>SEETRAX</td>
<td>4</td>
</tr>
<tr>
<td>STEWARTS OF READING</td>
<td>4</td>
</tr>
<tr>
<td>SURREY ELECTRONICS</td>
<td>63</td>
</tr>
<tr>
<td>WESTDEV</td>
<td>43</td>
</tr>
</tbody>
</table>

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