The mystery of Faraday's homopolar generator

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Valve crossover for audio
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Easy attenuators
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Receiver design III: noise & sensitivity

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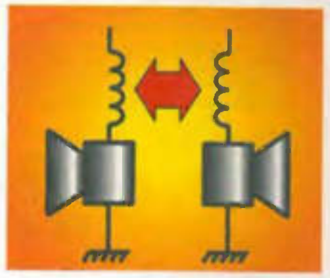
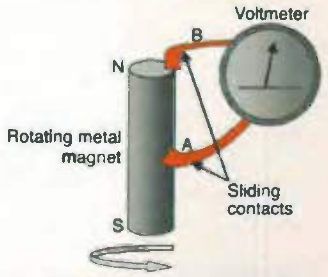
Robocop? Designed for bomb disposal and the like, this police robot can make decisions on its own. The month's top electronics news stories start on page 732.



When you mount two power amplifiers close together, coupling between their output inductors is inevitable. Probably the only article you'll find on the topic – from audio guru Doug Self – starts on page 749.



Dom DiMario has been replicating experiments carried out by Faraday 170 years ago. Has he discovered a phenomenon that cannot be explained by current theory? Turn to page 762...

Ethernet is very cheap to implement. You can pick up a PC interface card for under a tenner. Eddy Insam explains how you can put it to use for remote monitoring and control with the aid of this tiny low-cost Internet-ready controller – page 738.

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On the right track?

Thing are looking up for Railtrack. Its train announcements at a local station are computerised, but the computer seems to have feelings. It announces, "I am sorry to tell you that the ..." or "I am extremely sorry for the delay to your journey," although not "I am pleased to inform you..."

Use of the first person implies that the computer is accepting responsibility for events. Maybe it is frantically networking with other computers in distant signal boxes to ensure that future services will run smoothly – what a good idea.

On Tuesdays, the announcements are made by a male voice, whereas on Thursdays it is female. Perhaps it has a split personality – one like Marvin the paranoid android would probably be more suited to the post-Hatfield climate.

Could Railtrack be the first to have actually built a caring machine? Could feelings be given to other machines? Perhaps I could argue with the speed camera that's just flashed me.

But seriously, Railtrack and British Rail before it, have benefited from technology that saves money, and possibly even helps to make money too. For example, the move from steam to diesel and electric traction in the 1960s; the introduction of fully braked goods trains which eliminated the need for a guard; unmanned level crossing barriers; the mass signalling schemes of the 1970s and 80s which created signal boxes to cover large areas of track and which replaced many lever-frame boxes; driverless trains on the Docklands Light Railway. Even the infamous Beeching Report, which closed so many railway lines, called for massive new investment.

Technology does have its problems because of the change that it brings. The need to have a 'change management' plans with each new scheme is now apparent.

It is morally indefensible to throw a lot of people on the job scrap-heap without some concern about what they are going to do. Losing a job is like having a rug pulled out from underneath you – it hurts a lot. Big companies now thankfully recognise this and accept the responsibility to organise retraining schemes and outplacement services.

So it's relatively easy to cut costs, but how do you account for railway safety? Is it a bottomless pit into which vast sums of money are thrown to appease nervous passengers?

We seem to hold the railways close to our heart and expect them to be near perfect. It's still a very safe mode of transport, although the recent crashes have been horrendous.

Even more horrendous is the blind spot that we have for road transport. We tolerate over 3000 road fatalities every year as the price for the freedom of the road. Why don't we make more fuss about it? Why shouldn't cars be compulsorily fitted with radar or sonar linked brakes?

Investment in safety systems is all down to risk management. No system can be 100% perfect, and if you try to make it so the last 0.5% can cost as much as the 99.5% in effort. When the railways were split up there was a feeling that the safety regime was too tight and too costly.

Continental railways, it seemed, worked perfectly well with less regulation. For example there was more train cab radio control rather than costly signalling, and country stations had more basic facilities.

However, what seemed to have been overlooked was the chronic sate of the infrastructure in the UK. Railtrack was sold a dummy. Shareholders were blinded by the massive amount of property that Railtrack was about to inherit.

Future advice must be that when Governments 'sell off the crown jewels' check that they are not fakes. Industry is not exempt from this either – look at the sale of Rover to BMW.

Despite the present problems of Railtrack, the railways are going through an unprecedented renaissance. Urban and light rail schemes are being adopted by many cities and towns. Stations that were closed in the 1960s are reopening. The Channel Tunnel Rail Link is being constructed with talk of more high speed lines.

The break up of the railway industry has caused problems, particularly over responsibility for safety. However, it has opened up the network to more passenger and freight trains than ever before. Technology has an important part to play in speeding up this process and one sincerely hopes that the lessons of Southall, signal SN109 at Paddington, and Hatfield will be learnt and applied wisely. ■

Peter Marlow

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UPDATE

Chip provides glaucoma warning

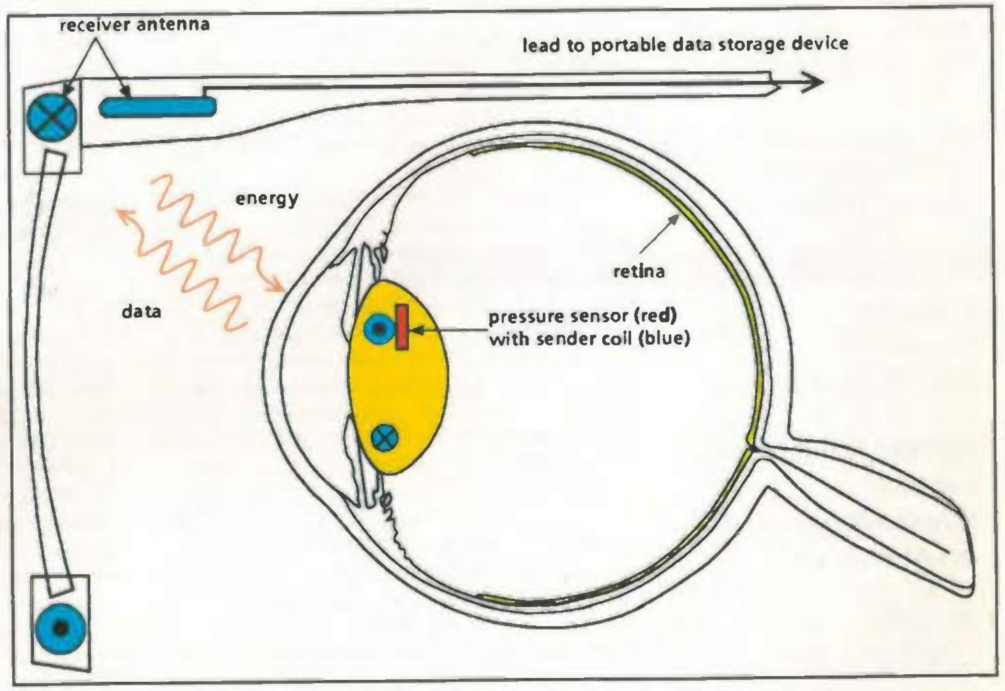
Scientists at the Fraunhofer Institute in Germany have developed a system that measures pressure inside the eye, warning glaucoma sufferers of danger.

A chip, pressure sensor and RF coil are inserted into an artificial lens in the eye. The external reading device, integrated in a pair of glasses, powers the circuitry through the RF link. This allows 24 hour measurement.

"Until now, measuring methods have virtually neglected fluctuations in pressure during the day and at night," said researcher Gerd von Bögel. A human study of the device is due soon, he said, while commercial devices could be ready in as little as a year.

In its normal form, glaucoma develops slowly and insidiously without pain. Fluid pressure inside the eye increases, reducing blood pressure to the retina. This can permanently damage the optic nerve and lead to blindness.

Lens incorporating a chip, pressure sensor and RF coil, provides early warning of pressure build-up in the eye, and 24-hour monitoring.



Solar-powered plane flies for 18 hours

NASA has flown Helios, its unmanned solar powered wing, for the first time at a significant altitude. An 18 hour test flight over Hawaii saw Helios reach over 23 000 metres, proving its 35kW of solar cells were up to the task.

Helios has a wingspan of 75 metres, yet weighs under 600kg. For the next test flight, Nasa's Dryden Flight Research Center hopes to get the wing above 30 000 metres.

By 2003, the team hopes to install hydrogen fuel cells that would enable the wing to continue flying through the night. Such a craft could then be used to provide communications at a much lower cost than satellites.

Flight at 30 000 metres is said to be similar to that on Mars.

BP buys Madrid facility for solar-panel production

BP is buying a semiconductor fab in Madrid from Agere Systems which it will turn over to production of photovoltaic solar panels.

By the end of next year, the plant will be one of the largest dedicated solar-cell facilities, with a yearly capacity of 60MW of crystalline silicon panels.

"Spain and the rest of Europe represent growing markets for solar power," said Harry Shrimp, chief executive of BP Solar. "This project in Madrid will enable us to continue to play a leading role in meeting demand."

The company plans to invest $100m in the factory which will create around 600 jobs.

BP claims to have nearly 20 per cent of the world market in solar panels. Last year its turnover was $200m and it produced over 40MW of photovoltaic equipment.

BP also recently opened a plant in Australia with a yearly capacity of 25MW. The firm calculates this level of production reduces CO_2 emissions by 35 000 tons.

Niobium capacitors replace tantalums

Tantalum shortages are bringing niobium into electronics as NEC announces the first niobium conductive polymer-type solid electrolytic capacitors.

Niobium is one of the alternatives being considered to replace tantalum in capacitors – as tantalum becomes less available and human rights issues are raised about some of its African sources.

According to NEC estimates, niobium is over 100 times more plentiful than tantalum.

"The construction is exactly the same as our tantalum conductive polymer capacitors," said an NEC spokesman. "It uses a new type of polymer, which is very difficult to make. Only NEC, Sanyo and one other company can make it."

There are other companies, claimed the spokesman, that make lower-performance non-polymer niobium types.

The company is claiming that its niobium capacitor, "has the quality and performance equivalent to the conventional tantalum capacitor," and that production yield is better, although it does admit that niobium capacitors will be slightly larger if all else is equal.

NEC anticipates that in the future, niobium capacitors will be used for high-capacity devices – about 1000μF – which were too costly to produce using tantalum, and will come to replace aluminium electrolytic capacitors.

NEC's niobium capacitor specification

Rated voltage	2.5V
Capacitance	220μF
Size	7.3 by 4.3 by 2.8mm
ESR	55mΩ
Part number	NB/PD0E227M
Samples	August 2001
Production	early 2002

Police robot works autonomously

Researchers at Sandia National Laboratories in New Mexico have unveiled a police robot that makes some 'how to' decisions on its own.

The idea is to free its operator to make the more critical 'what to do next' decisions during bomb-disposal and other law enforcement activities.

Even simple operations can be complicated on these robots. "Sometimes it's like playing a video game with a seven-lever joystick sitting upside down, with one eye closed, and with your boss looking over your shoulder," said researcher Phil Bennett. "Operators might think they're about to bump an object but they're really three feet away. Or they don't know if the robot will be able to fit between two cars or climb a flight of stairs. Often they don't accomplish these difficult tasks on the first try. The pressure can be intense."

The Sandia machine is based on a standard Wolverine robot from Remotec in Tennessee, with Sandia's own software – called SMART, an acronym for Sandia modular architecture for robotics and teleoperation.

During a demonstration in March, claims Sandia, the modified Wolverine, "shaved minutes off typical bomb responses even with the most skilled FBI robot operators".

Currently SMART automates commonly-needed robotic tools and behaviours in police work, such as automatic tool changes, tool placement, and bomb-disruptor aiming, as well as telerobotic straight-line movement in all directions.

New behaviours that could be included in future include path planning, machine vision, proximity sensing, obstacle avoidance, visual targeting and 'reachability' analysis.

An un-named "major Asian company" is looking at making flat screen TVs using technology from Oxfordshire-based Printable Field Emitters. The picture shows PFE's soon-to-be-revealed 15cm quarter-VGA (320×240) monochrome field emission display demonstrator. It does not need an expensive active matrix.

UK programmable analogue chips break into Japan

UK field programmable analogue array (FPAA) company Anadigm is breaking into the Japanese market through a deal with Omron.

"We are delighted with this alliance, and believe it will substantially accelerate the take-up of this exciting new technology," said Anadigm's CEO Mike Kay. Omron will sell Anadigm chips under the Omron brand.

The Japanese company has experience with the full custom analogue IC business, and sees a demand for FPAAs in applications needing under 10 000 units per month, said the company.

"Electronics engineers are seeking to design at higher and higher levels of abstraction, but a universal analogue HDL is not available, due to performance issues associated with analogue IC fabrication," said Kuninori Hamaguchi, CEO of Omron's semiconductor division.

"FPAAs offer a solution to this problem, which massively reduces design time and costs compared with analogue Asics."

The company will initially market FPAA technology in Japan, where it expects to establish, "a multi-million dollar market within two years".

It will also contribute to the development of FPAAs aimed at specific markets including industrial automation and medical equipment.

Anadigm's FPAAs are based on switched capacitor building blocks and design software from the company allows the chips to be configured on power-up. A wide variety of analogue functions are available.

New pressure sensor exhibits massive linear resistance change

The new pressure-sensitive material could provide useful user feedback from manual power tools.

Research at the University of Durham has yielded an electrically-conducting polymer with linear resistance change under pressure.

The composite material contains particles of metal dispersed throughout an elastomer. Resistance decreases from $10^{13}\Omega$ to just 1Ω under fingertip pressure.

"You can go from essentially an insulator to a conductor just under moderate pressure," said Professor David Bloor from the University. "That's an amazing range of any physical property, in any material... and under fairly non-extreme conditions."

Darlington firm Peratech has taken the material and demonstrated it in power tools, in a remote control unit and as a wearable PC interface.

Ported software that runs faster

Running code written for one processor on another normally means interposing adaptation software and a horrendous loss in speed.

Now a Manchester company is claiming it can port software in real time – with a speed gain.

"Overall we have seen speed-ups of 10, 20 or 30 per cent in complete applications," said Alasdair Rawsthorne, company chief technical officer of, Transitive Technologies.

Called Dynamite, it is described as 'code-morphing' software, running as a kernel around the target processor at run-time and converting original processor binary code on-the-fly.

Speed gains, said Rawsthorne, come, "simply because of re-optimisation of the code. Doing it at run-time means that the behaviour of the software can be observed as it runs and its performance improved."

The company is aiming the product at companies with a lot of existing code needing to be run on a new processor. "You can't buy a decent 68000 now," said Rawsthorne, "and you may have 100 engineer-years of 68000 code."

Dynamite has a re-usable core which is adapted for particular processors by attaching a front and back end. "We have adaptors for 68000, Sparc, Alpha, ARM, PowerPC and ×86 at the university," said Rawsthorne.

Transitive demonstrated ×86 code running on a MIPS processor at the recent Embedded Microprocessor Forum in the US. "We will be delivering a production version of the software at the end of September," he said.

According to Transitive, Dynamite is built upon a well-accepted method called optimising dynamic binary translation. Similar techniques have apparently been lab tested and used non-commercially by IBM and Hewlett Packard for specific CPU pairs and are also used in Transmeta's Crusoe processor which emulates an Intel ×86 processor in a VLIW machine.

Dynamite can be deployed, said Transitive, above the target operating system (OS) in 'multiple binary' mode, executing different binary code streams simultaneously in a single target environment. It can also run under the subject OS interfacing to raw target CPU hardware and processing applications and OS loads of the subject machine transparently to the user.

www.transitives.com

RF identification chip small enough for bank notes

This tiny chip is a radio-frequency identification (RF ID) tag developed by Hitachi. The firm reckons the 'mu-chip', as it calls it, is small enough at $0.44mm^2$ to be embedded in paper and bank notes. Operating at 2.45GHz, the wireless chip stores its ID in a 128-bit ROM, reducing the chances of tampering.

Memory chips that test themselves

Embedded memory specialist Virage Logic has developed a memory design that includes self-test and repair circuits.

The technique removes the need for testing machines and laser modification of devices following production, said the firm. It should increase the yield of devices containing embedded memory.

"Today you cannot integrate large amounts of SRAM without solving the yield problem, which is a burden for one of the fastest growing segments – communications and networking," said Dr Yervant Zorian, Virage's chief scientist.

Called STAR, the system uses redundancy to provide spare memory bits and a processor dedicated to testing bits and assigning replacements.

Normally performed at the wafer-sort stage of manufacture, the testing results are stored in an on-chip 'fuse box' which is used to allocate spare bits each time the chip is powered-up.

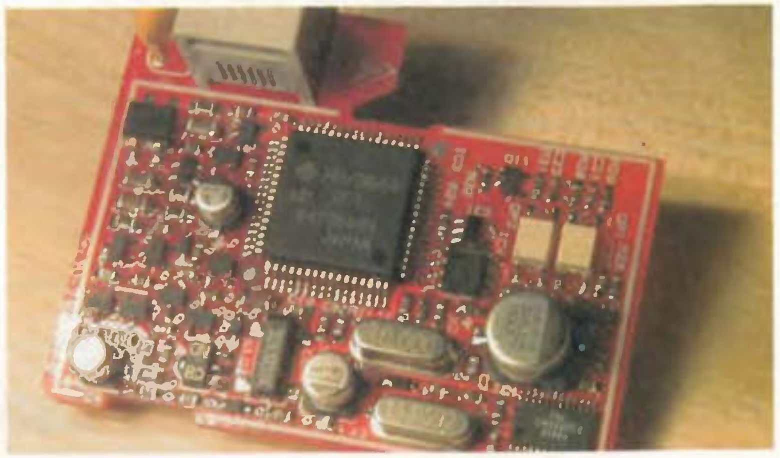
The firm claims the testing algorithms can pick up 99 per cent of faults.

This, said Zorian, "saves the customer literally millions of dollars by repairing otherwise defective ICs."

Virage has designed three embedded memories containing the self-test circuitry; 4Mbit and 512kbit single port designs and a 256kbit dual port version.

A licence for the design starts at $170 000 plus royalty fees.

Modem is powered by phone line



A phone-line powered modem has been developed by Edinburgh Communications using a V.22bis modem chip from Essex-based Consumer Microcircuits. The unit does not need any external power source, other than the phone line, and is aimed at metering applications. A full-duplex data rate of 2.4kbit/s is not fast, but is suited to the intended applications. At 3V, the modem consumes just 2.7mA, and under 2μA in power-save mode. The firms say the modem conforms to all European standards for connection to the telephone network, and EMC standards.

Scientists seek a successor to CMOS

IMEC, the Belgian research centre, has begun a study to discover the practical limits of CMOS and has initiated a program to explore alternatives to the workhorse chip-making process.

The study into the future life of CMOS will determine whether the process can be scaled beyond the 45nm level – perhaps as low as 22nm. IMEC reckons that mass production, not just the manufacture of lab devices, can be achieved at this level.

Lasting three years, the study will look mainly at front-end-of-line issues such as the gate stack, channel/substrate engineering, shallow junction formation, spacer technology and silicidation.

A major part of the experiment will focus on high-k dielectrics and metal-gate technology.

A second program called Emerald will look at device structures that will overcome the limits of CMOS below 35nm.

Examples of these structures have already been suggested from earlier IMEC studies. They include fully depleted silicon-on-insulators; high-mobility CMOS based on strained silicon/silicon germanium (Si/SiGe) layers; vertical devices; and double-/triple-gate and gate-all-around devices.

Designs will be studied to determine their suitability for mass production and compatibility with conventional CMOS transistors.

Meanwhile, back in the real world, IMEC has set up a two year program to develop a 0.18μm BiCMOS process for RF chips.

Heterojunction bipolar transistors will be based on silicon germanium carbon with CMOS devices having 3.5nm gate oxides and shallow trench isolation.

A fully integrated process including passives should be ready by the end of next year, IMEC said.

Intel researches two new memory technologies

Intel is researching two new technologies for memory, including polymer ferroelectric RAM and ovonix unified memory.

The polymer memory is non-volatile and is aimed at data storage. Polymer chains with a dipole moment store data by changing polarisation of the polymer between metal lines.

This technique has the beauty of avoiding the use of transistors completely for the memory array. Furthermore, the polymer layers can be stacked, so the memory array could be integrated above transistors in a chip.

Thin Film Electronics in Sweden has already fabricated some test structures up to 8Mbit in size which are said to have faster write speed than flash. However, read speed is slower.

Ovonix unified memory (OUM) uses a material called a chalcogenide, which changes from crystalline to amorphous when a current is passed through it. Thus the resistance changes, which can be sensed with a transistor, much like magnetic or ferro RAM.

A 4Mbit test chip is comparable in size to flash and DRAM, while read and write speed are 50 and 100ns, according to Intel. The material is compatible with CMOS.

STMicroelectronics and BAE Systems have also taken a licence for OUM from developer Ovonyx. ■

Virtual pets are coming to mobile phones. Intelliplus has signed a deal with Finnish firm Lumo Media to distribute the 'pets' in the UK. Virtual pets are reckoned to be the next big craze for teenagers and those wishing they were still teens. Pets are purchased like ring tones and logos for around £3, and maintained using SMS or WAP.

Can i/o via the web be simple?

Set up your own remote monitoring system using Internet protocols, Ethernet and a web server. Can you think of anything more daunting? Read on, you might be surprised how easy and cheap it is – especially if you've already dabbled with HTML. Later in the article, Eddy Insam provides a step-by step guide to help you get the simple hardware described here on line.

Building on my previous articles on Ethernet interfacing, the following shows you how to implement a real web server connected to your PC via a local-area network. This is a very basic design that will allow you to do things like display temperature, control a heater and switch lights remotely from any web browser.

Once you've grasped how it all comes together though, the opportunities are endless – especially now that fast wireless Ethernet links are becoming readily available.

I live in a very old house; it has a large basement with a large coal bunker where I keep many bits of fine wood that may come in useful some time. I need to keep the place at the right ambient temperature. This means a regular walk down the steps to check the walls for humidity. If there's a problem, I leave the lights or heating on for a few hours.

This is not a job I can delegate to a simple room thermostat, as it would leave the heater on most of the time and waste a lot of energy in the process.

Nowadays, I don't have to go all the way down the steps that often. I go to my PC, fire up my Internet browser and call up a web page. Up comes a screen page displaying the temperature in the room. If I want the lights or the heater on, I just click on a button on the screen.

How is it done? Well, I have an embedded web server in the basement. This web server fits in a small 125 by 125mm box fixed to the wall. It connects to the Ethernet LAN running around the rest of the house, **Fig. 1**. Pushing the description to the limit, I could call this a miniature dedicated 'private internet service provider'.

The display containing information from the remote controller unit is not static. Contents of the web pages are changed dynamically by data read by the unit via its I/O ports.

Similarly, I can control the heaters by making page requests from my web browser. I don't need any special software on my PC. All I need is my standard Internet web browser to give me a two-way simple I/O control.

The trend of using an Internet browser like a 'program' is gathering pace. There are certainly advantages to doing things this way – especially as any web-enabled piece of equipment can be used for controlling interfaces. This equipment includes WAP phones and palm-top computers.

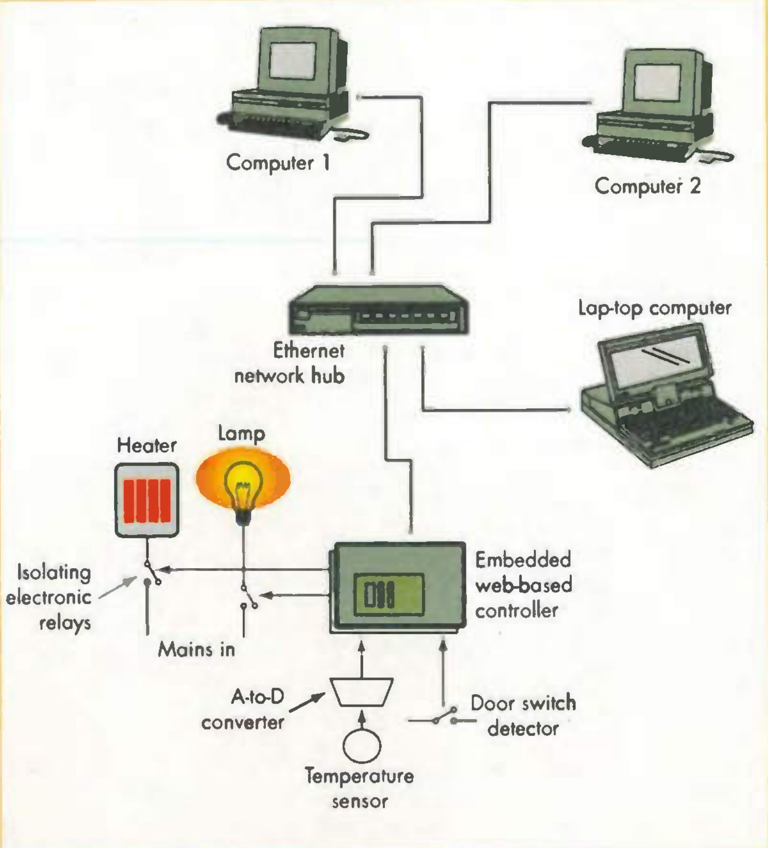


Fig. 1. A typical Ethernet network will have the various computers in the house or office connected to a central hub via standard network cable. The embedded controller connects to a spare slot in the hub and acts like a web server, accessible by all computers with a web browser by simply entering the right IP address.

So what is inside this box?

I have chosen the SitePlayer module (www.siteplayer.com) as the basis for my remote-control system, **Fig. 2**. This is a good example of current technology, and the kind of ready made product available for this sort of application.

SitePlayer is a 25mm-by-25mm module built as a piggyback board. I have chosen it mainly because it comes with a scripting web language that allows me to relate external events to internally-stored web page displays. There's more on this later.

Figure 3 shows the block diagram. The SitePlayer module consists of a Realtek RL8019AS Ethernet transceiver chip plus an 89C51 microcontroller. The Ethernet Transceiver chip conditions the Ethernet signals on the line and transfers the packets to the micro.

All the protocols are handled by the microcontroller. It also supports eight general purpose I/O ports plus one serial port. I am only using the parallel port in this design.

The micro has 64K of code space, with roughly 16K of program code dedicated for TCP stack and script interpreter code, and 48K of user programmable area to hold web pages.

The only external components required for this project are an external 10 BaseT line transformer, an RJ45 socket connector for the LAN, and some external standard circuitry to drive a mains controlled heater, a ceiling lamp and to sense a switch and the temperature sensor. **Figure 4** shows the full circuit diagram.

The 48Kbyte web area can include as much text, graphics, etc., as can be fitted in the space available. Web pages can contain special fields that are modified with data read from the I/O ports. Page or click requests will also cause the I/O ports to output different data. The web page becomes the 'user interface' Fig. 5.

Have you wired up your house yet?

Externally, a standard CAT network cable is used to connect to the rest of the local network around the house. This can include one or more PCs.

Network cables, connectors and crimping tools are now available from most electronics and computer shops. Many new houses and offices are now being fitted with network connections in every room as standard.

One of the best ways to get started is to buy a network 'kit.' These will include a couple of network cards for your computers – assuming you have more than one – a four or five port 'hub', and a few cables. The instructions supplied with this sort of kit are usually very good and will explain all you need to know about networking in detail.

Note that if you are only using one computer for this project – i.e. you don't have a hub – you will need to use a special changeover cable or adaptor to connect the computer to the module.

How does it all work?

In order to understand the basics, it is easiest to look at how things function from the PC's point of view.

When you fire up your Internet browser and request a web page, a number of events take place. At the lowest level, a full TCP connection is established between the PC and the remote server.

For the purposes of this article, think of TCP like a telephone call; something that you start by dialling a number, and that results in a fixed connection between the two points until one of the hand sets is put down. When the connection is live, bytes are exchanged in a way not too different from a plain RS232 wired link.

A few steps may need to precede this, such as obtaining the numeric address of the remote site. You can't phone anybody unless you know their telephone number. These numbers, or 'IP addresses', are usually stored in local tables, but they can also be found by making a separate call to a 'directory enquiries' server on the network, or on the Internet.

Of course, everybody connected to the world-wide web has to have a different address. Newcomers must be allocated a unique 'name and number'.

In theory, this should include my basement device. However, as I am running my own local network with no access to the outside world, I can cheat a little. I can allocate almost any numbers or web names I like to my interfaces. As a result, my project is called 'mybasement.com' and is associated with the address 192.168.1.250. This happens to be the default IP address of the SitePlayer module.

Once the TCP link is established, the browser and server can exchange bytes. The protocol used for web

Fig. 2. This one-square-inch module can hold up to 48K of web-page material, which can include text and images. Dynamic scripts are used to replace display fields with values read from the I/O ports. Similarly, request commands from the browser can set data outputs. The module also provides for eight parallel data bits plus one serial port for external control.

access is called HTTP. To request a page, the browser sends a 'GET' command string in plain text, usually followed by a file name and maybe a few other parameters, also in plain text. Commands are terminated with double line feeds.

The server simply replies with the contents of the requested file. Web pages are made of plain text, and use a formatting language known as HTML. This language allows for the inclusion of commands such as font and colour changes by the use of predefined bracketed 'tags' within the text.

If no file name is supplied, a default name such as 'Welcome.html' or 'Index.html' is substituted instead. When the file has been fully transmitted, the server closes the TCP connection, and that is the end.

To get another page, the browser must open another TCP connection, and send another GET command with the name of the new file.

A typical web site contains many HTML files. There is usually one main file called 'Welcome.html' or 'Index.html' and various other support files including bitmap images, java applets, etc.

In a standard page request, the browser will first request – and display – the main page. Then it will scan through the contents of this page looking for links to other files. For each link it finds, it will make further GET requests for each, one at a time. Each of these requests involves the opening and closing of a new TCP connection.

As the files arrive, they are presented to the user. If they are bitmaps for example, they are displayed at their correct location on the screen. Text files are very short and can display very quickly. Images, even compressed ones, are bigger and can take more time to download.

The HTTP/HTML scheme allows the various parts of a file to be displayed in any order, and give an impression of speed. Following this logic you can now understand why some web pages can load and display very quickly and why some pages display a blank rectangle while the image that goes within it is still being downloaded.

It should also be clear how straightforward it could be to implement a web server using a low-end microprocessor. The server only needs to wait for GET requests, and then return a string of pre-stored characters. The only limitation is the availability of enough ROM or RAM space to hold all the HTML files and data.

Of course I have simplified the argument a little. I am talking about the simplest of HTTP protocols, but this is all you need to know at this stage.

And now for the clever bit

The SitePlayer module has an internal flash memory big enough to hold up to 48K of web page data. This can include text, images, sounds, animations and anything else that can fit in that space.

Tools supplied with the module allow you to edit the HTML pages in the normal way on a PC and download them into the module's memory in compiled form.

This is not all; the module has a feature known as scripting. This links fields on its stored web page with internal variables in the microcontroller and also with external events such as I/O port data. The module will replace the tags with the actual values that are dynamically updated at display time. In order to do this, some non standard codes need to be added to the standard HTML text.

For example, assume there is a register variable inside the microcontroller that has been given the name port1 in the script, which happens to contain the decimal value 57. If the original HTML page contained the text: "*The value is ^port1*", it will be displayed – i.e. sent to the browser – as "*The value is 57*".

This scripting can be used in many ways. For example, the source text:

```
<a: img srce "led^n.gif">
```

can be used to display one of a number of pre-stored bitmaps depending on the value of *n*, i.e. led1.gif, led2.gif... This could be used to emulate a seven-segment numeric display for example.

Similarly, codes sent back from the browser can activate outputs in SitePlayer. Clicking on a web page button can cause an I/O output pin to change.

The scripting language is relatively simple, and can be learnt in an hour or so. Figure 5 shows my crude attempts at web design. More can be achieved with the imaginative use of graphics, buttons and java scripts.

The hardware

Figure 4 shows the full circuit. The parallel port signals are standard 8051 'PORT1' type ports. These can be set as output only or as bi-directional input-output.

I have used two pins as outputs to drive two mains controlled devices – a fan heater and a ceiling lamp. One input bit is used to sense a door latch, and the remaining five bits are used to read the output of an a-to-d converter using an LM335 diode temperature sensor. You can use any combination of input/output in your own designs. All the definitions are done in the software anyhow.

The line connection requires a 10BaseT transformer and RJ45 socket. It may be worth salvaging these from an old Ethernet PC card, as they can be difficult and expensive to obtain in one-offs.

I have used the LF1S022 Lan-Mate combined unit, which consists of a transformer and connector integrated into a single container. Addresses for web sites advertising these parts are given later.

Getting hold of the parts

The SitePlayer module can be imported directly from the manufacturers. The one-off retail cost is $49.95. Support software is separately available as a CD for $9.95, or it can be downloaded for free. Please note that shipping, VAT and handling can easily add another 50-60% to these costs. For information on obtaining Ethernet transformers see the web sites mentioned in a further panel within this article. An integrated RJ45 transformer can be also obtained from the SitePlayer web sales site. The author has a small stock of modules and transformers. Please contact him for more details.

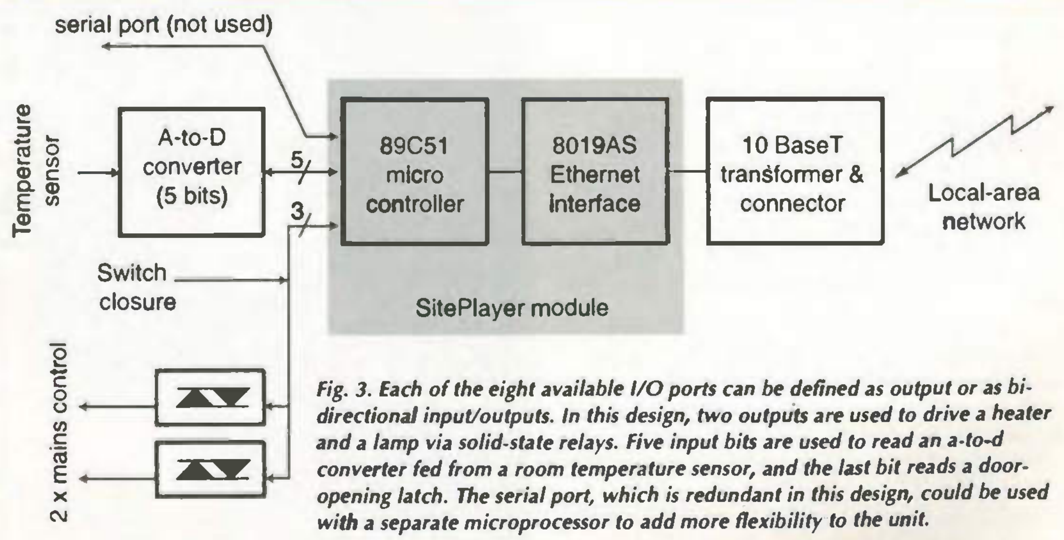


Fig. 3. Each of the eight available I/O ports can be defined as output or as bi-directional input/outputs. In this design, two outputs are used to drive a heater and a lamp via solid-state relays. Five input bits are used to read an a-to-d converter fed from a room temperature sensor, and the last bit reads a door-opening latch. The serial port, which is redundant in this design, could be used with a separate microprocessor to add more flexibility to the unit.

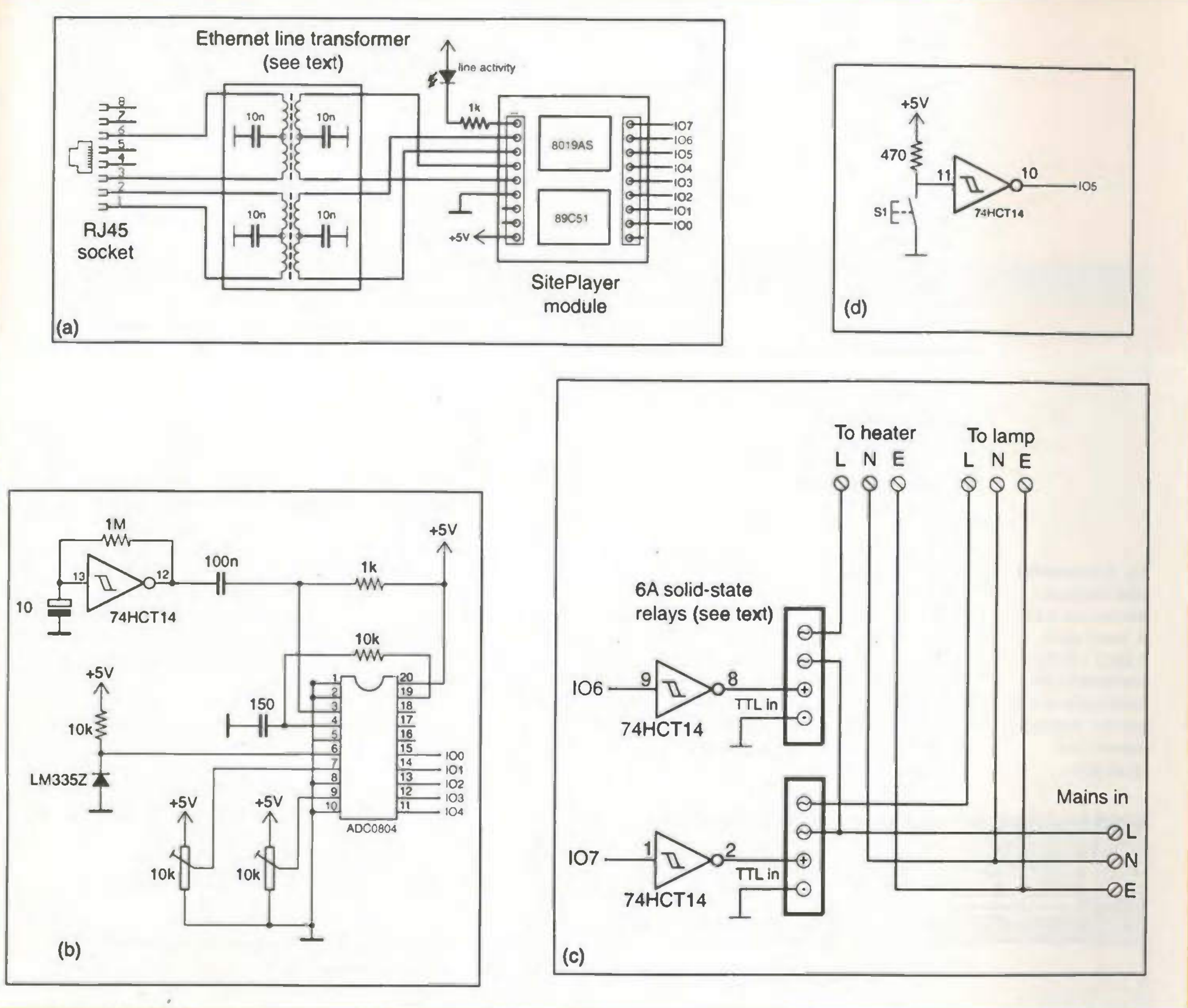


Fig. 4. The SitePlayer module is mounted on a socketed 18-pin piggyback module and needs only a transformer as an external network component. Its LED flashes when there is relevant activity on the Ethernet bus. The a-to-d converter is autonomous; only five bits are used to give a temperature span of 5 to 37°C with a one degree resolution. The range can be changed to other settings by adjusting the two 10kΩ pots. The zero crossing solid state relays that I used can switch up to 6A each from a TTL level input.

Figure 4b) shows the a-to-d converter circuit I used to sense local temperature. It incorporates a National ADC0804 chip in self-run mode.

A Schmitt trigger inverter – part of a 74HCT14 – forms a free running oscillator to start a new a-to-d conversion every few seconds or so. Only the five most significant output data bits are used. These connect to the lower five I/O pins of the SitePlayer module.

One nice thing about the ADC0804 is that you can set the analogue conversion end points to cover almost any two voltage limits. The two 10kΩ variable resistors trim the minimum and maximum end points so that the desired temperature range spans the full digital range of 00_{16} to FF_{16}.

The LM335 diode gives a steady 10mV change per °C around its defined output level of 2.98 volts at 25°. I have set my potentiometers to give a temperature span of 5 to 37°C. This is convenient as it corresponds to a resolution of one degree per bit, making life easy when computing the display assignments on the HTML web page.

Figure 4c) shows the circuit used to drive the ceiling lamp and the heater. I opted for ready-made solid-state relays; these can cost a bit more than using separate triac circuits, but they have other advantages such as encapsulation and environment proofing.

The models I used – namely Crydom MP or CX series, available from Farnell and others – take TTL level signals and can switch mains voltages at between 2 and 10 amps depending on model. If you use a different type, ensure they use zero voltage switching.

Note that I have driven the relay modules via the spare hex inverters. This is not strictly necessary, but I added them here just in case something goes wrong. I'd rather have the 74HCT14 blow up than my expensive Ethernet biscuit.

Fig. 5. Minimalist user interface. Should you want it, much more realism can be obtained by the imaginative use of colours, bitmaps, buttons and javascript.

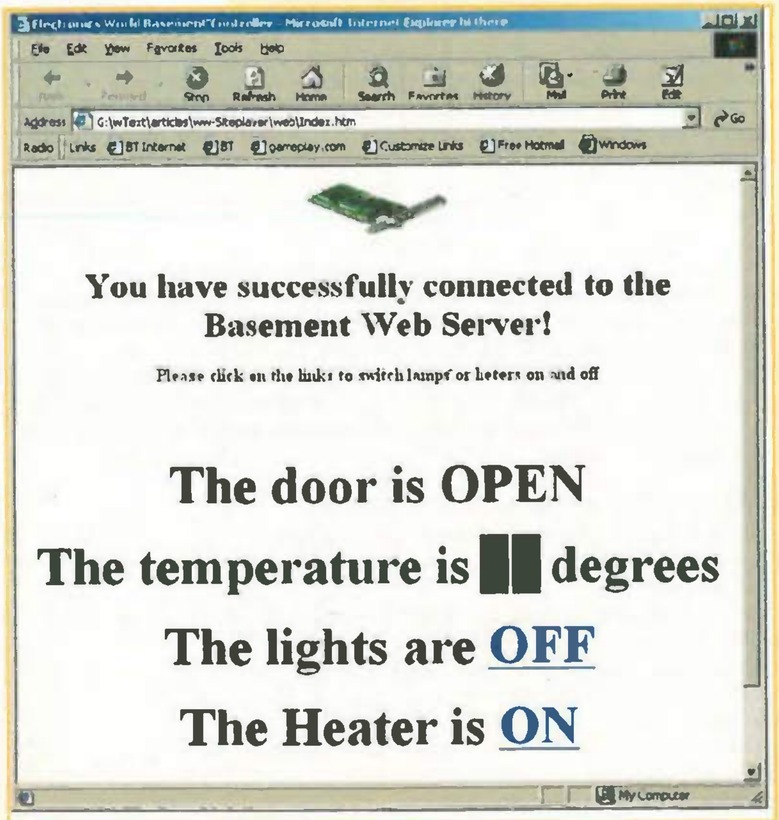


Figure 4d) shows the remaining input used to read the door latch. I use a simple burglar alarm magnetic reed switch, also connected via one of the spare hex inverters, again in case something goes wrong.

The whole thing is powered by a single regulated 5V supply – not shown. Current consumption is around 60mA.

What's next?

Once the electronics have been wired up, the rest of the job needs to be done in the software – not at the module end, but at the PC end.

You will need to connect the unit to your network. If you have a hub, connect the unit to the hub via a standard CAT5 network cable. If you connect the module directly to your PC, i.e. if there are no other devices on the network, do not forget to use a crossover LAN cable or adaptor.

The first stage is to get the module to act as a web server only, i.e. without any of the scripting facilities; these can be added later on. This makes it a bit easier to get things running.

You will also need the following files from the SitePlayer CD: 'sitelinker.exe', 'siteplayerpc.exe' and 'siteplayerserialdemo.exe.' These can be downloaded from the SitePlayer web site. You will also need the various jpeg image files.

I have included detailed step-by-step instructions, as the SitePlayer module manual can be a bit Spartan in places. It may also be useful to go through, or print, the various SitePlayer documentation files supplied as .pdf files.

Creating the control software

The stages involved are as follows:

1. Create a new directory 'c:\SitePlayer' (or similar) in your PC. This is where all the program files are going to reside, copy all the above SitePlayer files here. Create a new directory 'c:\SitePlayer\web\.' This is where all the web files in your project are going to reside.

2. Design a simple web page using a normal web or plain text editor. Start with something simple, you can always change it later. Feel free to add one or two simple images, but keep them small. If you are using a commercial web editor, remove any unnecessary HTML fields. Many web editors add many extra non-display fields, which are of no use here. Remember that you have a 48K total limit. Name your web file 'Index.htm', and place it in your 'web' directory. Ensure all image files you used are also in this same directory.
3. Using a text editor, create a new file named 'myproj.spd,' place it in your base directory. This will be your Definition file. Copy the text from List 1 to it. If your files are in different directories, ensure the entry named $SitePath points at the directory where you keep your web files, and that the other entries point at your base directory. Entry $InitialIP defines the target IP address to be used for the download from the PC. This must be the same as your SitePlayer module; the default is 192.168.1.250 unless you have changed it. More on this later.
4. Run 'SiteLinker.exe.' Use the Menu option File/Open to select and open your definition file 'myproj.spd.' Nothing will be shown on the screen. If this is the first time that you've run this program, take the opportunity to use the Menu option Configure/Editor to select Notepad as your choice editor. Next, use the Menu option Download/Make Config file to compile the source. The generated file will be called 'myproj.spb'. This name was taken from the $Sitefile entry in the description file. The file just created is a compiled version of all your source data, and it is the file that will be downloaded to the SitePlayer module. Now, use the Menu option Download/Download SitePlayer to send the compiled file to the flash EPROM in the SitePlayer module via the Ethernet. Ignore the 'no password' message.

5. Run the Internet Browser in your PC – for example Internet Explorer. Type 'http::/192.168.1.250' on the address line and click the GO button. You should see your web page, which is now coming from your SitePlayer module.

Call yourself lucky if all this happened smoothly and

List 1

Sample Definition file. Please copy this file to your base directory and name it myproj.spd. Ensure that all paths point to the directories where you keep your own files.

```
; Sample Myproj.spd Definitione file
;
; Lines with a semicolon in first column are comments and ignored by the compiler
;
; Compile this file with 'SiteLink.exe' which takes this file, and the contents
;  of your web directory, and creates a new compiled file 'myproj.spb' which is
;  then downloaded to the SitePlayer module.

;
; Sets the name or description of the device
$Devicename "Electronics World Demonstration heater controller"

; Sets SitePlayer to find its IP address from a DHCP server
; if your PC is enabled to do so)
$DHCP on

; Sets password for downloading web pages and firmware, don't need any to start with
$DownloadPassword ""

; Sets password for browsing web pages, donít need any to start with
$SitePassword ""

; Sets SitePlayer's IP address to use if no DHCP server is available
$InitialIP "192.168.1.255"

; Sets the binary image compiled file name that will be created, enter the full path
$Sitefile "C:\SitePlayer\myproj.spb"

; Sets the root path were your web pages are
$Sitepath "C:\SitePlayer\WEB"



; Definitions for SitePlayer standalone direct port I/O
; We need these definitions so that they can be viewed with a ^objectname replacement
; in your web HTML files. They can also be input by making an object of the same name in a form


        org 0ff00h
p1              ds 1                    ; Port 1 all 8 bits

        org 0ff11h
io0             ds 1                    ; Port 1 Bit number 0
io1             ds 1                    ; Port 1 Bit number 1
io2             ds 1                    ; Port 1 Bit number 2
io3             ds 1                    ; Port 1 Bit number 3
io4             ds 1                    ; Port 1 Bit number 4
io5             ds 1                    ; Port 1 Bit number 5
io6             ds 1                    ; Port 1 Bit number 6
io7             ds 1                    ; Port 1 Bit number 7
```

you got a result first time. The first problem you may encounter is compilation errors. These are mainly due to commands in the definition file not pointing at the right directories, so these should be easy to fix. There may also be entries in your HTML web page file the compiler may not understand. Keep your HTML file simple to start with.

Your main problem will be addressing errors. It may be that your network is not able to access the SitePlayer default address for various reasons – usually subnet masking. One way around this is to change SitePlayer's IP address to be accepted by the masks in your machine.

In my case, the Ethernet address of my machine is 192.168.0.1, and the subnet mask is 255.255.255.0. When I first started using SitePlayer, I could not access it as it had a default address of 192.168.1.250. I had the choice of either changing my subnet masking to something like 255.255.0.0 – easy to do, but not such a good idea if your PC is also used in other networks – or change SitePlayer's own IP address to something like 192.168.0.2.

You can change SitePlayer's IP address using the supplied utility 'SitePlayerSendDemo.exe.' This program communicates with SitePlayer via the serial port, not via the Ethernet. You will need to add some extra hardware and voltage level conversion circuitry to connect the serial port in the module to the serial port in your PC. See the SitePlayer documentation for more information.

Alternatively, if your PC supports DHCP, it can assign an IP address to SitePlayer automatically. Ensure there is a $DHCP on entry in your 'myproj.spd' descriptor file.

More information

See 'PC Interfacing Via Ethernet', Eddy Insam, *Electronics World*, May and June 2001 issues. These articles comprised a description and general introduction to TCP/IP in embedded systems.

Visit www.eix.co.uk/Ethernet for more detailed information on the above.

SitePlayer's web site is www.SitePlayer.com. Demonstration programs can be downloaded from here.

www.haloelectronics.com and www.pulseeng.com supply 10BaseT transformers.

Adding scripting

Once you have your server operational, you can now add scripting facilities. Try the following steps to enable simple scripting and to allow the browser to read the I/O data pins on the microcontroller:

1. Open a text editor on your web file 'Index.htm'. Somewhere near the bottom of the page, add the line <H3>BIT 5 OF PORT1 IS ^io5</H3>. The H3 is only there to make the text look bigger, ^io5 is the variable denoting bit 5 of port1. You can see that some of these variables have been pre-defined in the descriptor file 'myproj.spd'.

2. Compile and download the newly generated '.spb' file to SitePlayer, by repeating items 4 and 5 in the previous section. The browser will now display the extra line of text 'BIT 1 OF PORT1 is 0'. Now press the switch connected to pin 5 in Fig. 4d) – assuming you have wired one already. Next click on the 'refresh' button on your PC browser. The following line of text will now be displayed: 'BIT 5 OF PORT1 IS 1'. The web page reflects the change in the micro's data port.
3. As an extended demonstration of the above, copy bitmap files '0_blk.jpg' and '1_blk.jpg' to your web directory. These files contain image bitmaps of seven-segment LED indicators. Now add the following line anywhere in your 'Index.htm' web file: <img src="^io5:1_blk.jpg" width='40'height='60'>. Repeat step 4 above. The resulting display will now show a seven-segment bitmap showing either a '1' or a '0' reflecting the data value on the port. The entry ^io5:1 returns the first digit of the numeric variable as a single character which defines which of the two jpg files is to be displayed. A full list of commands available and more examples can be found in the SitePlayer manual.
4. The full port can also be read as one byte by using the variable 'p1' – also defined in the definition file. Ensure that all the ten LED jpg files are in the web directory. Add the following text to our 'Index.htm' file: <img src="^p1:3_blk.jpg" width="40" height="60"> <img src="^p1:2_blk.jpg" width="40" height="60"> <img src="^p1:1_blk.jpg" width="40" height="60">. Recompile and reload as described above. The display will now show a three digit number reflecting the binary data pattern in port1, changing from '000' to '255'.

So much for reading data *from* SitePlayer. Sending data *to* it is a little bit more complex:

1. Create a new text file 'myproj.spi' in the web directory. You want three text lines in this file. You must ensure the third line contains no text, i.e. the file ends with two new lines.

```
HTTP/1.0 302 Found
Location: /Index.htm
<third line should have no text>
```

2. Edit the 'Index.htm' web file. Add a new line:

   <H3>port 6 = ^io6</H3>.

   The command 'io6=^io6~1' is used to read bit 6, invert it, and place it back (toggle). The reference to myproj.spi causes a link to what is basically another HTML file. This file does nothing but cause a link back to your index page, while at the same time updating the variable, including the I/O port bit on the micro.

3. Compile and download the file as already described. Run the browser and click on the port 6 link. See how the output of this pin toggles on every press using an oscilloscope, or connect a LED to it. Also note how the seven-segment displays change. This is because the full port is read/write.

The above is just a summary of what can be done. Many possibilities exist – especially when the basic software is used in conjunction with more powerful languages like javascript, and also when SitePlayer is used with the serial port option.

You don't need to purchase the SitePlayer device to see all this in action. There is a free simulator for the PC available on the SitePlayer web site. You can devise and test your own pages using this simulator.

To end with...

I wonder if the PC of the future is going to be nothing more than a sophisticated web browser, with all computing power done externally across the Internet, or within the peripheral units around it? It will be interesting to see what will happen to all the various operating systems. After all, every browser should work the same way. Maybe it will also mark the end of dedicated, platform-specific software.

Technology is fast catching up. New products such as Bluetooth will remove the need for running wires around the house. A few manufacturers are already working on combined Bluetooth/network appliance interfaces. ■

The Author

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List 2

A minimal sample web page. Copy this file to your web directory. Ensure that all the named image files are also present in the same directory.

```
<html>
<body bgcolor="#CCCCCC">
<b><H1>You have successfully connected to the Basement Web Server!</H1>
<H3> BIT 6 OF PORT1 IS ^io6</H3>.
<H3> BIT 7 OF PORT1 IS ^io7</H3>.
<H3> BIT 6 OF PORT1 IS <img src="^io6:1_blk.jpg" width='40' height='60'>
<H3><a href="ww-web.spi?io5=^io5~1">CLICK TO TOGGLE PORT 5 = ^io5</a></H3>
<H3> THE FULL DATA BYTE IS
<img src="^p1:1_blk.jpg" width='40' height='60' align="absmiddle">
<img src="^p1:2_blk.jpg" width='40' height='60' align="absmiddle">
<img src="^p1:3_blk.jpg" width='40' height='60' align="absmiddle">
</H3>
</body>
</html>
```

Improving valved test gear

You can pick up valve-based surplus test equipment very cheaply, but the valves within will no doubt have deteriorated due to their age. Wim de Ruyter explains how to get round this problem.

Inspired by Dave Mien, who produced a MOSFET-based 6V6 equivalent recently, I tried to synthesise a 6AU6 pentode. This will not work with one MOSFET alone. The mega-ohm impedances needed can only be achieved via cascode circuits, as they are in transistorised oscilloscopes.

In the solution described here, the bottom element of the cascode is a 'superfet' circuit. It comprises a junction fet and a p-n-p bipolar transistor. The resulting transconductance is the reciprocal of the unbypassed source resistor.

Any desired transconductance can be realised. The strong negative feedback action of this source resistor makes the input circuit ultra linear. When the unbypassed resistor is a metal film type, the long term stability is much better than could ever have been achieved using valves.

Adjusting the circuit to suit the valve

By selecting the right junction fet pinch off voltage, you can adapt the circuit to the valve it has to replace. The gate voltage of the mosfet is stabilised by a 12V zener as shown. With DC-coupled amplifiers this must he higher. A service manual for the instrument in question is indispensable.

The maximum drain voltage of the BS5125 is 600V. Gate-source capacitance is 160pF, and the dissipation one watt. This will suffice in most oscilloscope amplifiers. Only the junction fets need to be matched in balanced oscilloscope amplifiers.

Often, stabilised power supplies use cathode-follower circuits. Here, the valve can be replaced by a mosfet without needing any peripheral components. Such circuits are always negative feedback loops. Of course an appropriately-rated heat sink must be added in this case.

Often, the price difference between valves and semiconductors is small, but in many cases, valves cost considerably more. I have been unable to synthesise a pentode circuit in which the suppressor grid is essential. Also the remote cut-off pentode is still a problem. Fortunately such pentode circuits are seldom used in measuring equipment.

Implementing the idea

So far so good, but how do you connect the semiconductors and associated components to valve sockets? Very simple, when the circuit is mounted on an inverted valve socket. You can then use lengths of stainless-steel wire to plug the inverted socket into the original valve socket.

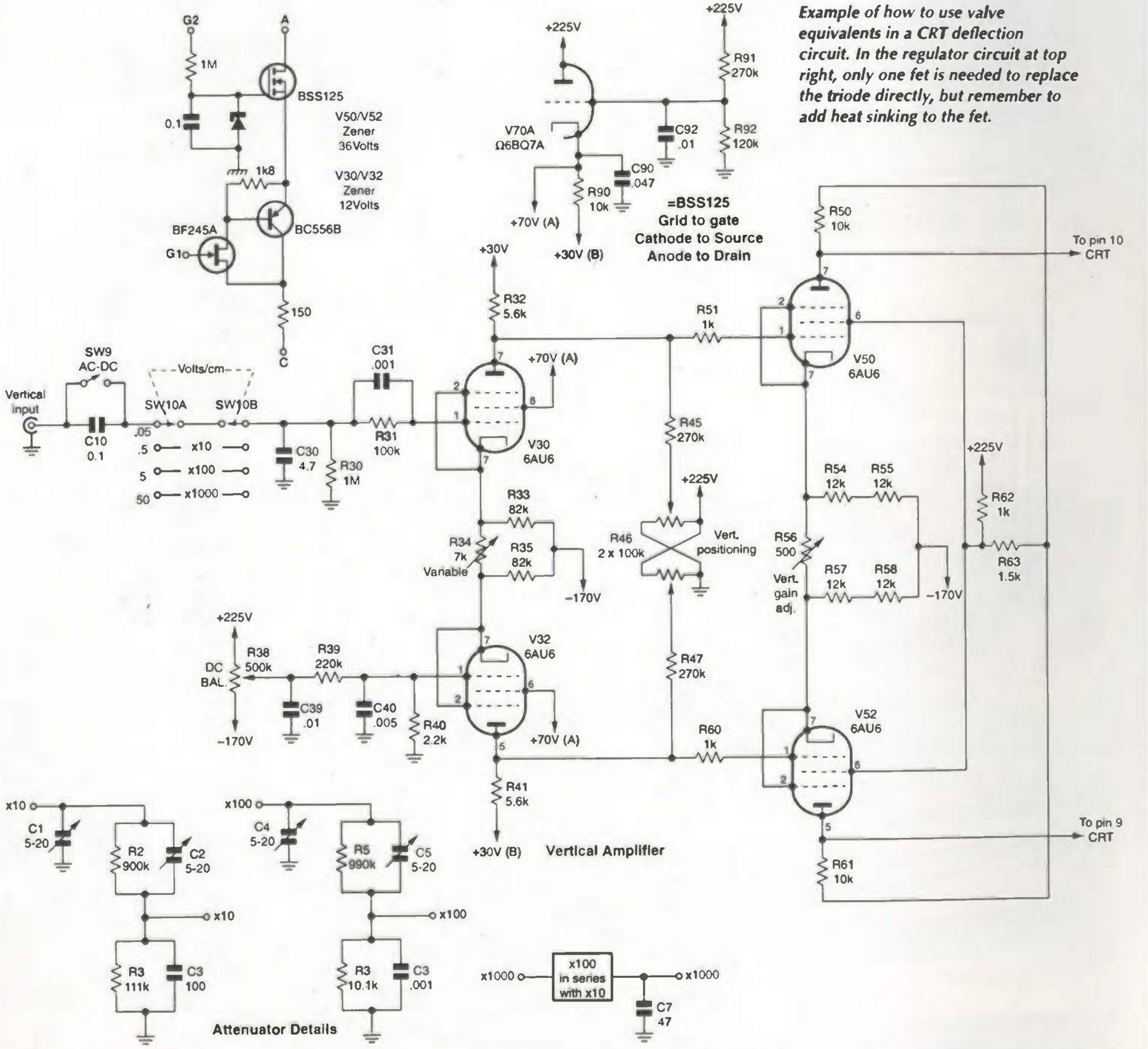
Suitable stainless wire is available as welding rod. It is a millimetre in diameter. The length of the 'double male' connector pins is half an inch, in accordance with the accepted standard.

A small circuit board can easily be fastened to the central tube of the valve socket by soldering. I used double-sided un-etched circuit board, with one side forming a screen between input and output. The bottom element of the cascode is mounted on one side, the top element on the other. Strip board may be easier to use, but you lose the screening properties of the double-sided board.

Remember that electrically contacting materials must maintain their connection integrity in humid conditions, so materials other than stainless steel are not advisable.*

Valve pins are of Invar steel. This is an alloy of steel with 36 percent nickel and hence inherently stainless. ■

*I would advise deburring, and perhaps rounding, the ends of the steel wire so as not to scrape any coating off the socket of the original equipment. You might consider trying solder-coated copper wire. It's very easy to tin thick wire with the aid of plumbers' flux, but remember to clean it thoroughly afterwards with meths or soapy water. Also, if you replace some valves in a piece of equipment with a lot of valves, it might be worth checking the effects of the reduced heater-winding load – Ed.



Example of how to use valve equivalents in a CRT deflection circuit. In the regulator circuit at top right, only one fet is needed to replace the triode directly, but remember to add heat sinking to the fet.

Crosstalk

in amplifier output inductors

In applications like home cinema, numerous power amplifiers need to be packed together as compactly as possible. Do their output coils interact? And if so, does it matter? Doug Self investigates.

When designing a stereo power amplifier, the issue of interchannel crosstalk is always a concern. Now that amplifiers with up to seven channels for home theatre are becoming more common, the crosstalk issue is that much more important, if only because the amplifier channels are likely to be more closely packed.

But here I deal with only one important aspect of it. Almost all power amplifiers have output coils to stabilise them against capacitive reactances. A question often raised is whether inductive coupling between the two is likely to degrade crosstalk. It is sometimes suggested that the coils – which are usually in solenoid form, with length and diameter of the same order – should be mounted with their axes at right angles rather than parallel, to minimise coupling. But does this really work?

I think I'm pretty safe in saying there is no published work on this, so it was time to produce some. The magnetic coil coupling could no doubt be calculated – though not, without a good deal of work, by me – but as is so often true in the glorious pursuit of electronics, it was quicker to measure it.

The coils I used were both 14 turns of 1mm diameter copper wire, overall length 22mm and diameter 20mm. This has an inductance of about 2μH, and is pretty much an 'average' output coil. It is suitable for stabilising amplifiers of various kinds up to about 150W/8Ω. Above that you might need to look at the wire diameter.

Different coils will naturally give somewhat different results, but extrapolation to whatever component you are using should be straightforward; for example, twice the turns on both coils means four times the coupling. Coil diameter should be considered, and so on.

Figure 1 shows the situation in a stereo power amplifier. The field radiated due to the current in Coil A is picked up by Coil B and a crosstalk voltage is added to the output signal at B.

Figure 2 shows the experimental setup. Coil A is driven from a signal generator with a source impedance of 50Ω, set to 5V rms. Virtually all of this is dropped across the source resistance, so Coil A is effectively

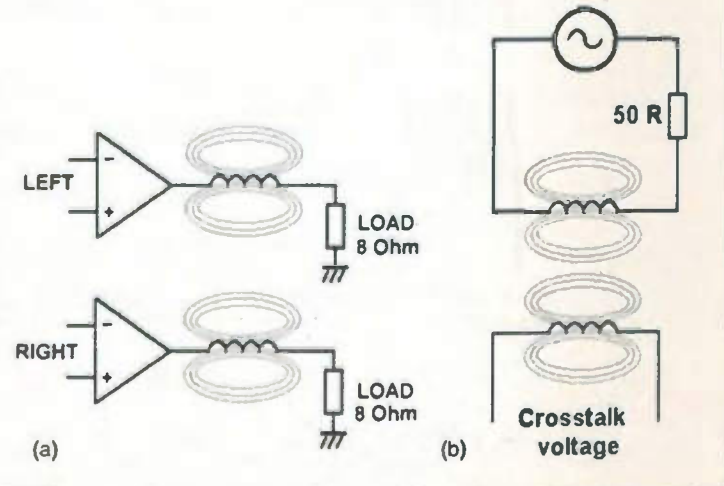


Fig. 1a). Magnetic coupling of output coils in a stereo power amplifier, a), and the experimental circuit for coaxial coils, b). The 'transmitting' coil A is driven with an effectively constant current, and the voltage across the 'receiving' coil B measured.

driven with a constant current of 100mA rms.

Figure 3 shows the first result,taken with the coils coaxial and their ends touching. This proved, as expected, to be the worst case for coupling.

The crosstalk rises at 6dB/octave, because the voltage induced in Coil B is proportional to the rate of change of flux, and the magnitude of peak flux is fixed. Capacitive crosstalk can be neglected because of the very low impedance of the receiving circuit.

The 6dB/octave rise shows that this is clearly not the same as conventional transformer action, where the frequency response is essentially flat. In a transformer, the primary inductance is much greater than the circuit series impedance. As a result, the magnetic flux that couples with the secondary halves when the input frequency doubles. The voltage induced in the secondary remains the same.

The worst-case crosstalk – coaxial coils touching – at 20kHz was taken as the 0dB reference, referred to as CAL on the plots. This represented 2.4mV rms induced across Coil B. Now 100mA rms of current flowing in Coil A corresponds to 800mV rms across its associated 8Ω load, so this gives a total crosstalk figure from channel to channel of –54dB at 20kHz in the worst case.

Crosstalk carries on deteriorating above 20kHz but no-one can hear it, so I for one don't care. All crosstalk figures given below are at 20kHz; I do appreciate that the ear is a good deal more responsive at 1kHz, but given the 20kHz figure, the crosstalk can be worked out for any desired frequency using the 6dB/octave characteristic.

The load resistance of the receiving – i.e., crosstalked to – channel has very little effect as the voltage induced in the coil is at a very low impedance indeed. It is essentially the impedance of the coil.

I then separated the coils by 10mm at a time. With each increment, the crosstalk dropped by 10dB, as you can seen from Fig. 3. At 110mm spacing, which is quite practical for most amplifier designs, the crosstalk had fallen by 47dB from the reference case, giving an overall crosstalk of 54+47=101dB total. This is a very low level, and only occurs at the very top of the audio band.

At 1kHz, where the ear is much more sensitive, the crosstalk will be some 25dB less. This brings it down to –126dB total, which I can say with some confidence is not going to be a problem. This is obtained with what looks like the least favourable orientation of coils. Coil-to-coil coupling is –32dB at 50mm, and the figure at this spacing will be used to compare the configurations.

The next configuration tested was that of **Fig. 4**. Here, the coils have parallel axes but are displaced to the side. The results are in **Fig. 5**; the crosstalk is now –38dB at 50mm. With each 10mm spacing increment the crosstalk dropped by 7dB. This setup is worse than the crossed-axis version but better than the coaxial one.

The final configurations had the axes of the coils at 90°; the crossed-axis condition. The base position is with the corners of the coils touching, **Fig. 6**. When the coil is the mid position, still touching, crosstalk almost vanishes as there is a cancellation null. With the coils so close, this is a very sharp null and exploiting it in quantity production is quite impractical. The slightest deformation of either coil ruins the effect.

Moving Coil A away from B again gives the results in **Fig. 7**. The crosstalk is now –43dB at 50mm, only an improvement of 11dB over the coaxial case; turning coils around is clearly not as effective as you might suppose. This time, with each 10mm spacing increment the crosstalk dropped by 8dB rather than 10dB.

The obvious next step is to try combining distance with cancellation as in **Fig. 8**. This can give a good performance even if a large spacing is not possible. **Figure 9** shows that careful coil positioning can give crosstalk better than –60dB (–114dB total) across the audio band, although the spacing is only 20mm.

The other curves show the degradation of performance when the coil is misaligned by moving it bodily sideways by 1, 2, 3 and 4mm; just a 2mm error has worsened crosstalk by 20dB at 20kHz. Obviously in practice the coil PCB mounting hole won't move – but it is very possible that coils will be bent slightly sideways in production.

Figure 10 gives the same results for a 50mm spacing, which can usually managed in a stereo design. The null position once more just gives the noise floor across the band, and a 2mm misalignment now only worsens things by about 5dB. This is definitely the best arrangement if the spacing is limited.

When there are coupling problems, there is always screening as a possibility. Pieces of metal cost money, so this should be a last resort, not least because they appear to be relatively ineffective. Metal screens are 100% effective at eliminating capacitive crosstalk, but the situation with magnetic coupling is very different. I spent some time doing informal experiments with bits of steel and aluminium, and they proved relatively ineffective, improving things by only a few decibels.

In summary

Coil orientation can indeed help, but only if applied intelligently. Simply turning one coil through ninety degrees gives an improvement of only 11dB. If it is carefully aligned to cancel out the coupling though, there

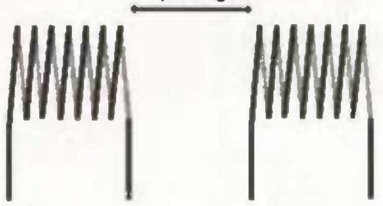


Fig. 2. Physical coil configuration for the measurement of coaxial coils.

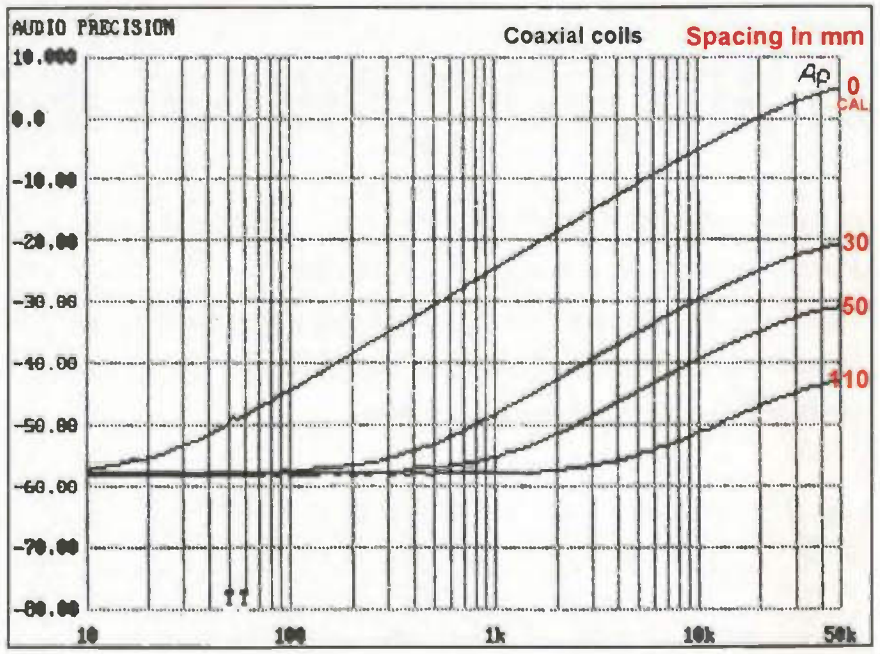


Fig. 3. Crosstalk versus spacing for coaxial coils.

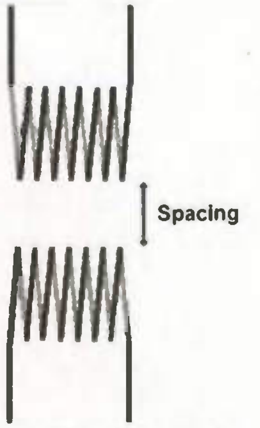


Fig. 4. Coil configuration for non-coaxial parallel-axis coils

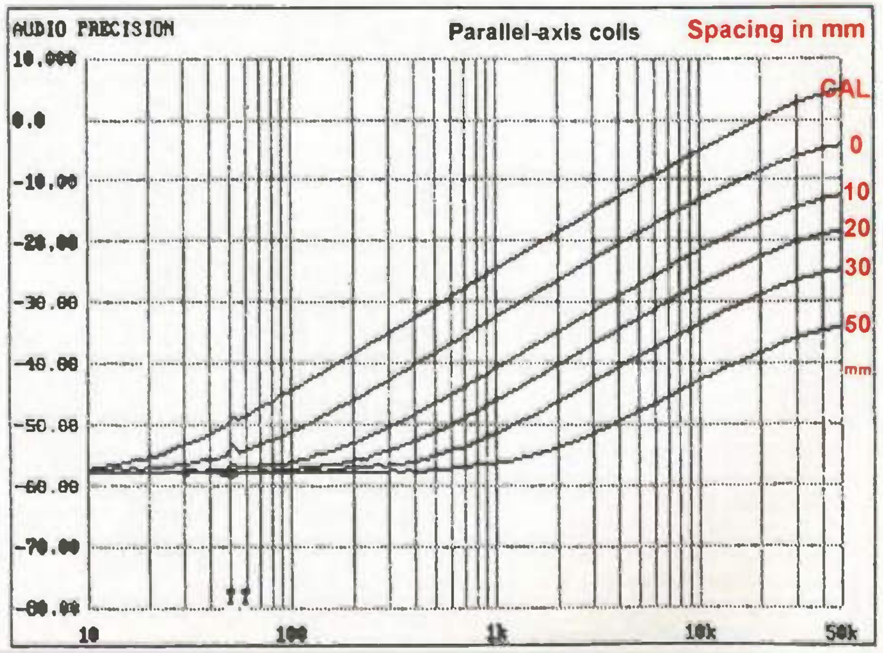


Fig. 5. Crosstalk versus spacing for parallel-axis coils.

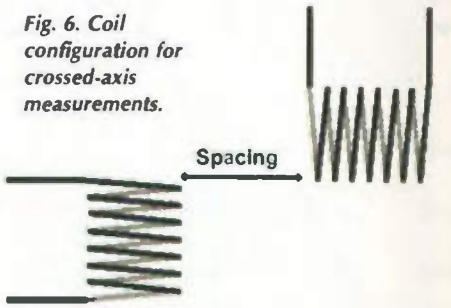


Fig. 6. Coil configuration for crossed-axis measurements.

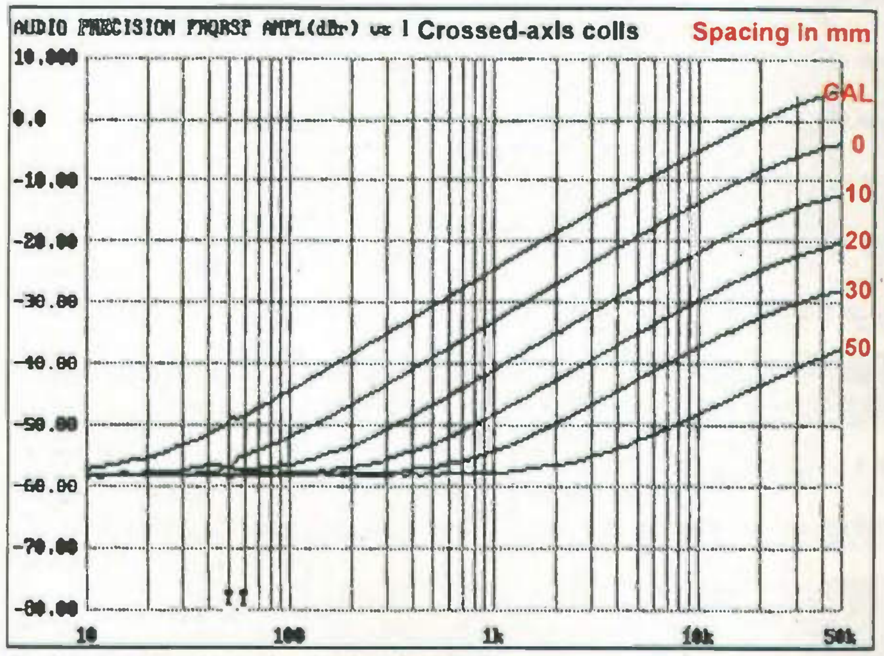


Fig. 7. Crosstalk versus spacing for crossed-axis coils.

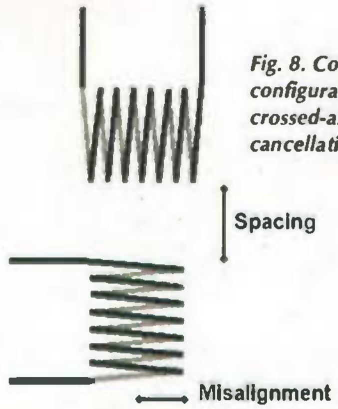


Fig. 8. Coil configuration for crossed-axis with cancellation.

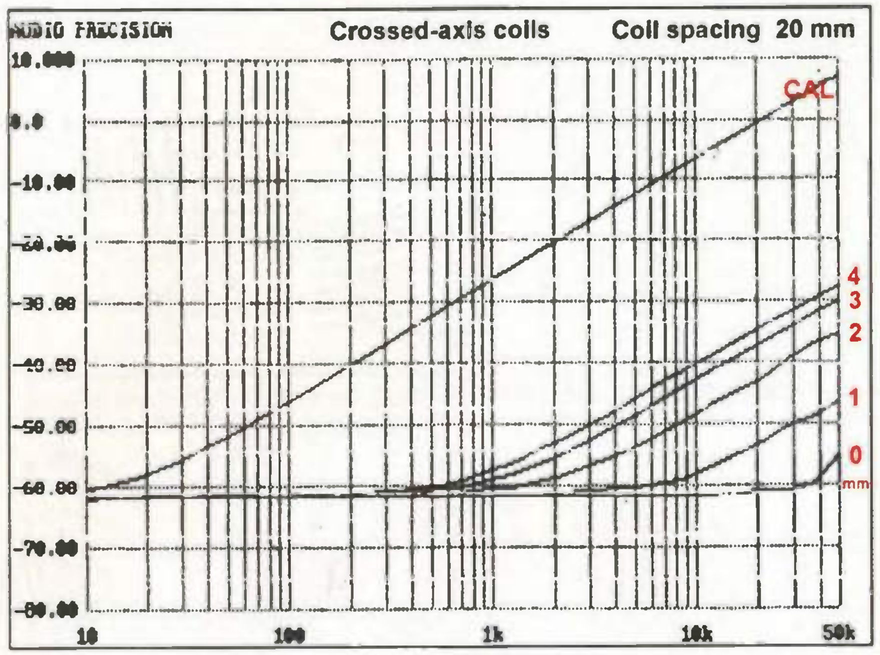


Fig. 9. Crosstalk versus alignment for crossed-axis coils spaced at 20mm, using cancellation.

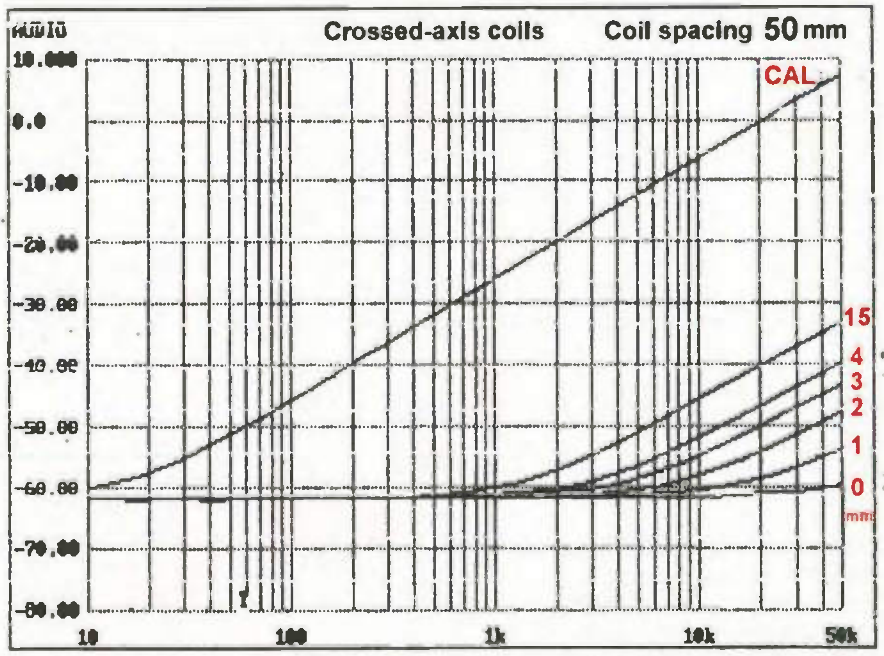


Fig. 10. Crosstalk versus alignment for crossed-axis coils spaced at 50mm, using cancellation.

is a big improvement; see how –38dB in Fig. 5 becomes –61dB in Fig. 10 at 20kHz. The alignment needs to be accurate to plus or minus half a millimetre, but this should be no problem with normal PCB layout techniques.

On a typical stereo amplifier PCB, the output coils are likely to be parallel – probably just for the sake of appearance. However, their spacing is unlikely to be less than 50mm unless the output components have been deliberately grouped together for some reason – and I can't think of a good one.

As with capacitive crosstalk, physical distance is cheaper than anything else, and if the results are not good enough, use more of it. In this case the overall crosstalk at 20kHz will be 54+38=–92dB total, which is probably already well below other forms of interchannel crosstalk. A quick quarter-turn of the coil improves this to at least –114dB at 20kHz. At 1kHz it will be even lower. It should do. ■

Morse code

the first communications revolution

Ian Poole looks at communications technology from the days before wireless was capable of carrying voice – Morse code.

Today, the internet and mobile phones are looked upon as being a telecommunications revolution. New generations of mobile phone are about to hit the market, and the internet has become an accepted part of everyday life.

But over a hundred years ago an equally important communications revolution took place with the introduction of the Morse telegraph system. This linked towns and cities that had previously relied on messages carried by riders on horseback. Even countries and continents were linked by this new technology changing society for ever.

An unlikely person

The inventor of the new system was a most unlikely person. Although the name of Morse is well known, few people realise that he was an artist of some renown – in fact today he is judged to be one of the finest artists to come from North America.

Born in 1791 to Jedidiah Morse, a distinguished geographer and preacher, Samuel Findley Breese Morse was an indifferent scholar, although he took an interest in the lectures on electricity and he enjoyed painting. Finally he managed to graduate from Yale in 1810 after which he became a clerk for a book publisher in Boston Massachusetts. However he was unhappy in this career and he became an itinerant artist. His portraits are judged to be some of the finest ever painted by a North American artist.

To improve his style Morse made some trips to Europe and in particular to England. On returning from one trip in 1832, shortly after the electromagnet had been discovered, he entered several conversations about electricity and communications with fellow passengers. Out of this exchange of ideas, a telegraph system began to develop.

The first telegraph system

It took some time for Morse to develop his system because most of his time was taken with his painting. However, to speed progress, he enlisted the help of others, including a man named Alfred Vail who was a gifted mechanic. Progress became much faster and a system using dots and dashes to represent letters was developed by 1838.

The next stage was to secure funding for a trial system, but this proved to be very difficult. Undeterred by the lack of interest, Morse continued without most of his other partners, finally securing funding for a 65km line between Washington and Baltimore.

Accordingly, on 24 May 1844 Morse sent the first message, "What hath God wrought!" The content of this message was chosen from the Bible (Numbers Chapter 23, verse 23) by Annie Ellsworth, the young daughter of a good friend.

With the successful opening of the first line, people and organisations quickly saw the benefits it could bring. Accordingly its use spread very rapidly. Within a year many new companies came into being to run lines to all parts of America.

New lines were installed, often along side railroad tracks. Towns that were previously isolated now had communications links with the outside world. Previously, communications relied on stagecoaches or the Pony Express, requiring days or weeks to cross the whole of the USA. In fact so successful was the Morse telegraph that only 19 months after the Pony Express was started it had to be discontinued.

Not only was the Morse system used in the USA: Europe and the rest of the world used it because of its effectiveness and simplicity. The first European line was set up between Hamburg and Cuxhaven in 1847, and many others quickly followed suit. Soon the need to link countries across seas and oceans was realised and in 1866 a submarine cable link was established between Britain and the USA, and by 1871 a link to Australia was established.

As a result of the vast use of the telegraph, stations were set up in towns, and in many remote areas repeater stations were required. Often, new towns arose as a result of these. One famous example is Alice Springs in the Middle of Australia. A line was required from Adelaide in the South to Darwin in the North from where messages could be sent to London and a repeater was needed in the middle of the desert. Alice Springs was chosen because of it was located near a suitable pass in the McDonnell range of mountains.

Morse keys

Since the very first keys were produced an enormous variety have been manufactured. The first key used by Vail when he was preparing for the telegraph demonstration in 1844 was very simple device, as you can see from the diagram. However it sufficed for the first transmission.

Further developments were made, but around 1848 a key known as the 'Camelback' was introduced. As a spring was not considered useful, the exaggerated curve shifted the balance towards the back of the key, keeping it open when not in use. Although springs were added later, the same style was retained for many years.

Key styles changed and straight levers became more common. In the US, James H Bunnell patented what he called a 'Steel Lever Key,' which first appeared in 1881. Keys like this were used for very many years and can still be bought today. In the UK though, keys tended to be much heavier like the ones used by the Post Office.

Just after the turn of the century more revolutionary changes to key design took place. With many operators spending many hours operating, many suffered from a complaint known as 'glass arm' or 'telegraphers paralysis'. To overcome this, people started to investigate the possibility of mechanical semi-automatic keys. The first successful variety was known as the Vibroplex. By pressing the 'paddle' to the right, a series of dots were made. Dashes were made manually by pressing the paddle to the left. This greatly reduced the level of movement required to send Morse Code and successfully reduced the occurrence of glass arm.

As time went on, electronics technology progressed and eventually fully automatic keyers capable of sending dots and dashes were designed. The first ones used valves

typical KOB - key on base. This includes a camelback key and sounder. This example was made around 1860.

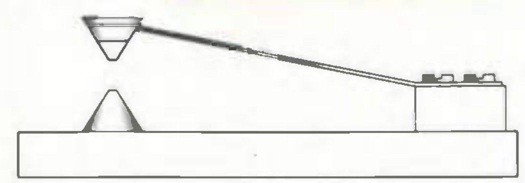
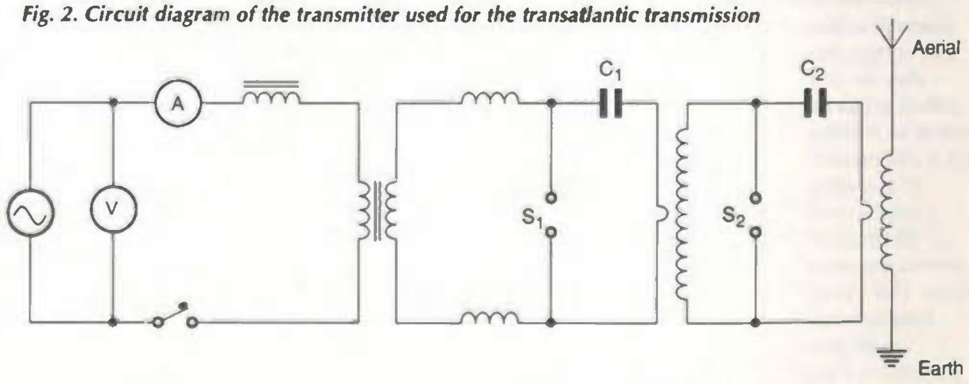


Fig. 1. Vail's correspondent was used to transmit the message between Baltimore and Washington in 1844.

A steel lever key. This particular example was manufactured in the USA around 1920.

A key used by the British Post Office. Note the heavy brass construction. This key was manufactured around 1900.

Fig. 2. Circuit diagram of the transmitter used for the transatlantic transmission



and were introduced around the early 1940s, although the introduction of transistors and then integrated circuits enabled them to be built more easily as well as allowed greater amounts of flexibility.

Now it is possible to use computers to generate and read Morse. However most people still enjoy using a key to send Morse. Besides, computers are not as adept as the human mind at receiving Morse – especially when there is any interference.

Famous times

In its life, the Morse code has been used to carry many historic messages. It was used to carry information about the end of the American Civil War. After General Grant accepted the surrender of General Lee on 9 April 1865, he sent the message: "General Lee surrendered the army of Northern Virginia this afternoon on terms proposed by myself the accompanying additional correspondence will show the conditions fully".

The Morse code was widely used for wireless transmissions, especially in the very early days when it was not easy to modulate a signal to carry sound. As a result many historic transmissions were made using Morse code. One was the first transatlantic radio transmission made by Marconi in 1901.

Transmitting across the Atlantic

Marconi had been developing his wireless system for several years and had been installing his equipment on ships. For them to be of particular use they needed to be able to communicate over large distances.

Marconi felt that it would be possible to send signals across the Atlantic and accordingly he set about proving whether this was possible. He set up two stations, one at Poldhu in Cornwall, and the other in Newfoundland.

The Poldhu station sent out the letter "s". This letter was chosen for two reasons. The first was that it would be easy to recognise. The second was probably more important. The transmitter was a very new design and it could not be trusted to transmit dashes without the risk of a breakdown!

Finally on 12 December 1901, Marconi was convinced that he could hear the weak set of dots through the static. Later he was able to send messages across the Atlantic using the system, and created headline news on both sides of the Atlantic.

Distress signals

Another incident of major importance occurred in 1912 during the maiden voyage of the Titanic. In order to cut the time of the crossing the liner steered a northerly course to shorten the distance travelled. Unfortunately she struck an iceberg on the night of 14 April and although she had been declared unsinkable, the great liner took on water and slipped to the bottom of the seabed. The two wireless officers, Jack Phillips and Harold Bride stayed at their posts until the last moments of the ship sending out the distress signals CQD and SOS. CQD was the recognised maritime distress signal, and SOS was due to replace it.

The distress signals were heard on both sides of the Atlantic as well as by the SS Carpathia that was about 93km away. As a result Carpathia steamed towards the stricken liner only to reach it an hour and twenty minutes after it sank to rescue some 700 survivors. Nevertheless over 1500 people died in the tragedy.

Finale

With new and more sophisticated forms of communication now available, the use of Morse has declined considerably in recent years. Many years ago the use of radio began the decline of landline Morse systems. This was compounded by the introduction of the teleprinter that used a keyboard to enter messages, and provided a printed copy at the far end. In Britain the Post Office discontinued the use of landline Morse in 1932, although its use continued until the 1960s in both the USA and Australia.

The use of Morse continued for many years after this particularly for ships, although teleprinters and then computers combined with the growing use of satellite communications spelled the end. Finally from midnight on 31 January 1999, international regulations no longer required ships at sea to be able to make distress calls in Morse.

Despite this some ships – especially those from the third world – still use it as a low cost alternative to the more expensive satellite systems. Radio amateurs still use it widely because it offers advantages in terms of the simplicity of the equipment and being able to make contact under conditions when other forms of communication would not be able to get through. As a result its use will continue for many years to come, continuing a tradition that is over 150 years old. ■

A	.-
B	-...
C	-.-.
D	-..
E	.
F	..-.
G	--.
H	....
I	..
J	.---
K	-.-
L	.-..
M	--
N	-.
O	---
P	.--.
Q	--.-
R	.-.
S	...
T	-
U	..-
V	...-
W	.--
X	-..-
Y	-.--
Z	--..

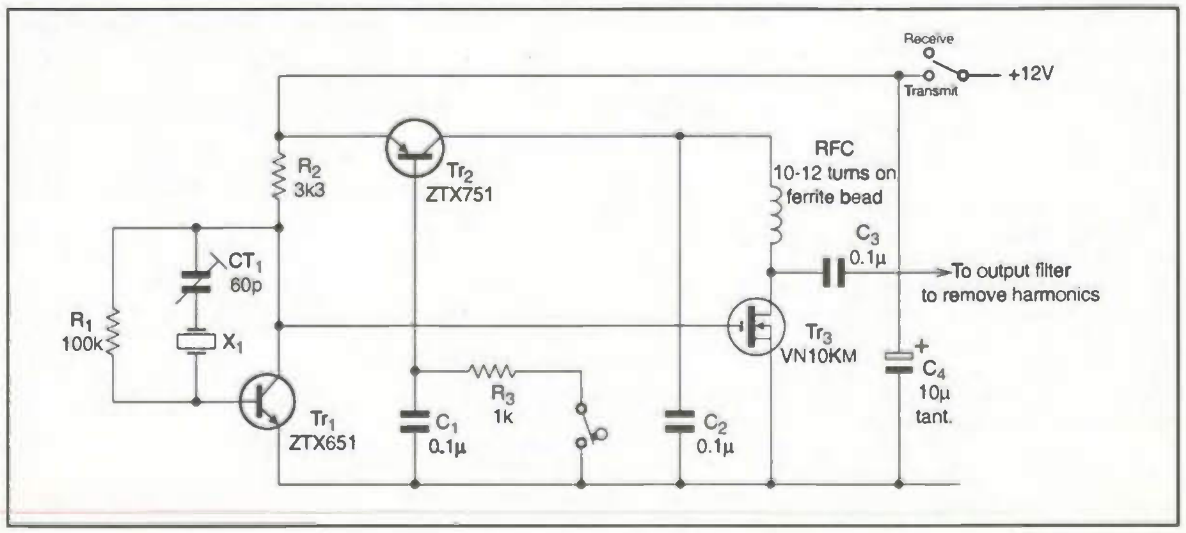
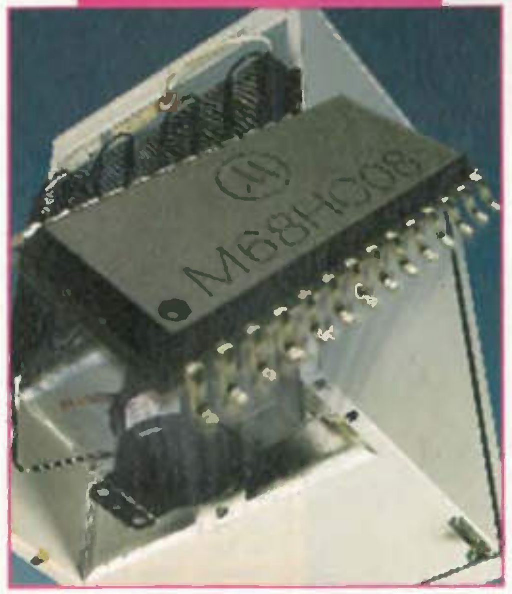


Fig. 3. The Oner – circuit diagram of a simple low-power Morse transmitter that can occupy less than an inch cubed. It can be built in an evening yet is still capable of providing contacts over distances of several thousand miles. This circuit is reproduced with kind permission from George Burt GM3OXX.



Many refrigerators still use rudimentary electromechanical control systems. But with microcontroller alternatives becoming so cheap, and energy consumption becoming such an important issue, the arguments for resisting change are weakening. Here, William Mackay of Motorola explains the benefits of using a low-cost microcontroller.

The digital fridge

In most of today's domestic appliances there are many opportunities to replace dated mechanical parts, improve and standardise inflexible manufacturing methods and develop economic appliance control using a microcontroller solution.

Complemented with new flash technology, the MC68HC08 Microcontroller family provides a convenient development environment and manufacturing flexibility for low-cost appliance embedded solutions. This article discusses a low cost solution for a domestic refrigeration appliance.

Compressor control

Basic conventional fridge control typically uses a bi-metallic thermostat to control the fridge temperature. The thermostat simply applies power to the compressor based on a coarse mechanical temperature setting.

This method has been used in the past for some time and is now rapidly becoming dated, due mainly to the increasing demand for more efficient control of the appliance and for an increase in the number of functions that the appliance can perform.

Temperature is more accurately controlled using negative temperature co-efficient thermistors – NTCs for short – with the microcontroller analogue-to-digital converter module. Figure 1 shows conventional compressor control and a microcontroller solution.

In contrast to conventional compressor control, a more efficient and long-term cost-effective solution can be implemented using the microcontroller to control the compressor motor using a triac and a relay. The PTC thermistor connected in series with the start winding is no longer required and the sequence of events required to start the induction motor is now under software control.

The relay is energised and closes the contact to apply line voltage to the start and run windings simultaneously. The triac is fired on the first zero-crossing point of the line voltage, and every successive zero-cross detected for a period of approximately 40ms.

In normal operation, after this time the compressor will have started and the triac will be in the non-conducting state until another start sequence is invoked. The compressor remains powered and running while the relay remains energised. With the PTC thermistor out of the circuit, considerable cost and a few watts of power can be saved.

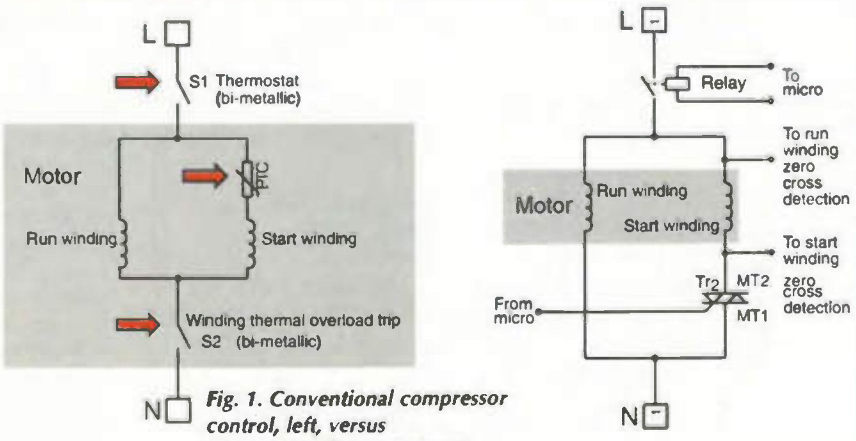


Fig. 1. Conventional compressor control, left, versus microcontroller compressor control, right.

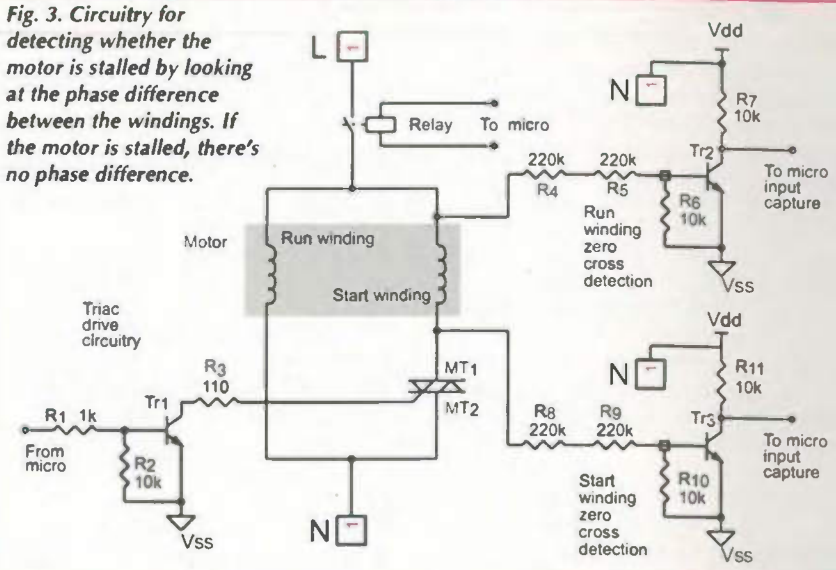


Fig. 3. Circuitry for detecting whether the motor is stalled by looking at the phase difference between the windings. If the motor is stalled, there's no phase difference.

Refrigeration system

An MC68HC908KX8 Microcontroller forms the heart of the refrigeration system, providing an adaptable platform for a range of functions suitable for low to high-end appliances.

Using the microcontroller in a refrigeration appliance can provide various possibilities for developing improved operational efficiency and versatility. As the system is under software control there is better scope for improving system efficiency with more accurate electronic temperature measurement and compressor control. This is complemented by additional functionality provided from the device feature set. A schematic of a typical implementation is shown in Fig. 2.

Functional overview

The temperature of the system is primarily dependent on inputs from two NTC thermistors, and a potentiometer. One thermistor detects the fridge ambient air temperature and the other detects the evaporator temperature. The potentiometer is used to select the desired ambient air temperature of the fridge.

Other application features included are an audible alarm, used to alert the user of a 'door-open' condition or over and under

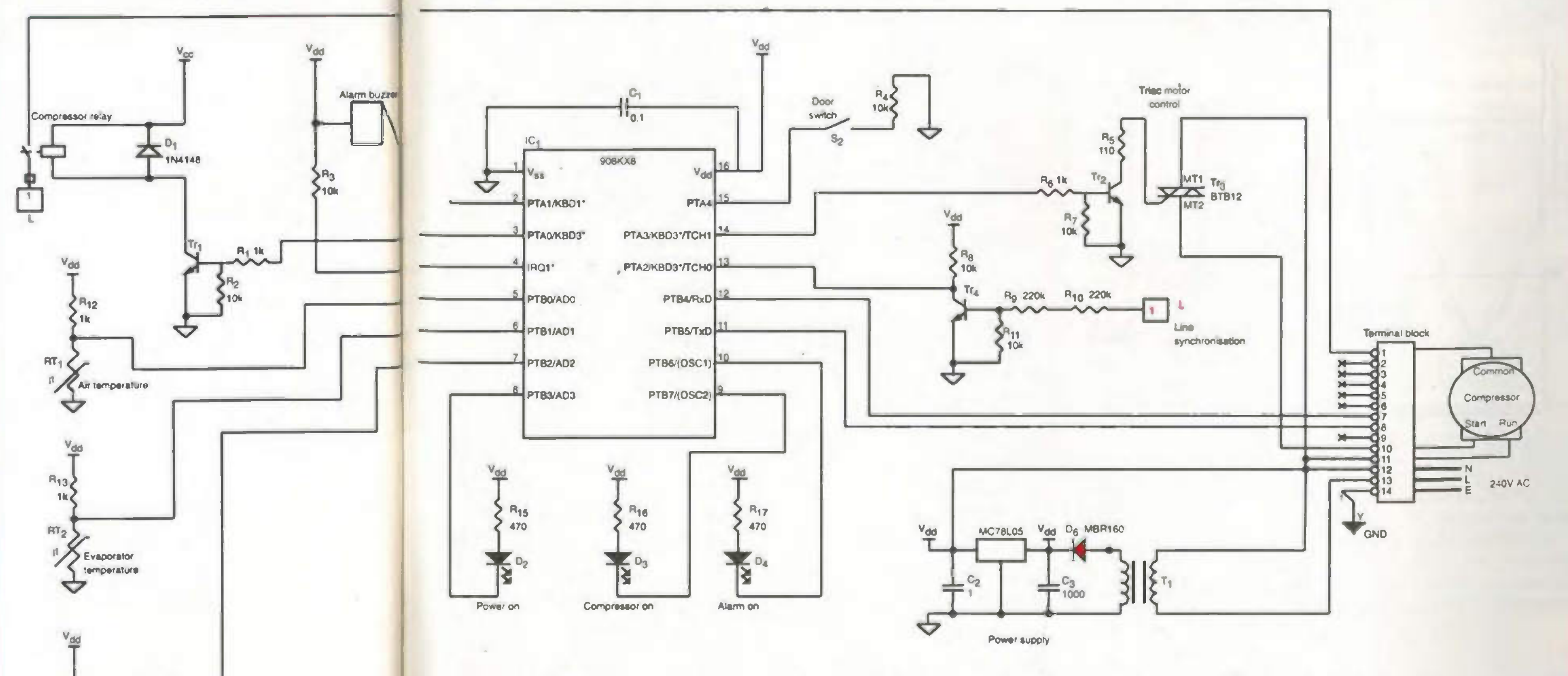


Fig. 2. A typical implementation – domestic refrigerator control system using Motorola's HC908KX8 flash microcontroller.

temperature conditions. A door switch determines the status of the door as either open or closed. There are three system status LEDs. One indicates power-on, another indicates when the compressor is powered, and the third LED is a visual indication of a 'door open,' 'over temperature' or 'under temperature' alarm condition.

Stall and rotation detection

Compressor motors in today's refrigeration appliances are thermally protected by a bi-metallic contact embedded in the motor windings. This contact is in series with the AC line supply to the motor and will open and remove power to the motor when overheating occurs. This can be caused by various sources, for example, by electrical start-up failure that can result in a motor stall condition.

If the motor is stalled, overheating occurs fairly rapidly and may cause permanent damage to the motor. Mechanical failure of the compressor pump is another potential problem source. This may impair the rotation of the motor and again result in a stall condition.

Detection of a motor-stall condition can be achieved efficiently and economically using the input capture features of the device. This has the added advantage of monitoring the rotation of the motor during the normal run condition, **Fig. 3**.

After start-up and during normal running conditions, the continuous rotational motion of the motor and magnetic field induces a voltage in the now redundant start winding. In this condition there is a measurable phase difference between the start and run windings.

If the motor is not rotating, that is, stalled, then there is no measurable phase difference between the start and run windings. This is illustrated in **Fig. 4a)** and **4b)**.

The zero-cross detect circuit conditions the above waveforms into usable digital signals forming a series of pulses switched between 0V and V_{dd} at half the period of the AC line frequency. In this case, that's every 10ms.

Digital signals derived from the zero-crossing points of the run and start windings are conveyed to the microcontroller input capture pins. A sample of the waveforms applied to the input capture pins is shown in **Fig. 4c)**.

In summary

The refrigerator is an appliance that in the past was controlled primarily by mechanical parts. It provided the user with little controllability and no operational information.

Today's consumers are demanding better efficiency and more intelligent appliances. Using the HC908KX8 flash microcontroller, the system design can benefit from a flexible development life cycle.

Using a microcontroller can also provide attractive appliance features, and from a manufacturing point of view, a range of appliances can be accommodated using the microcontroller as a standard hardware platform.

Printed circuit boards can be standardised for anything from low end to high-end appliances, and application software versions can be implemented with ease using the programmability and in-application re-programmability of the 8K flash. This minimises development time and reduces manufacturing costs while improving time-to-market. ■

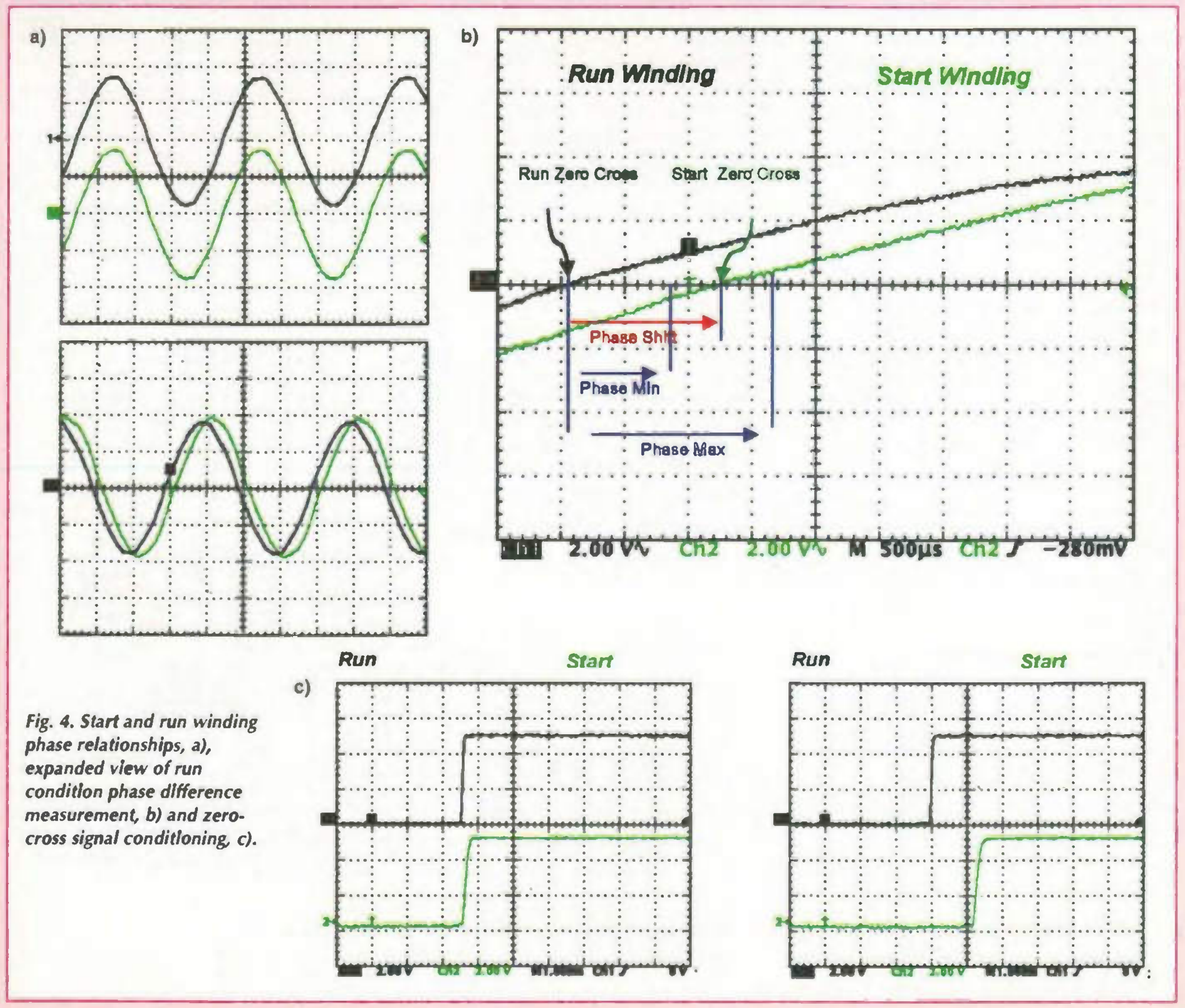


Fig. 4. Start and run winding phase relationships, a), expanded view of run condition phase difference measurement, b) and zero-cross signal conditioning, c).

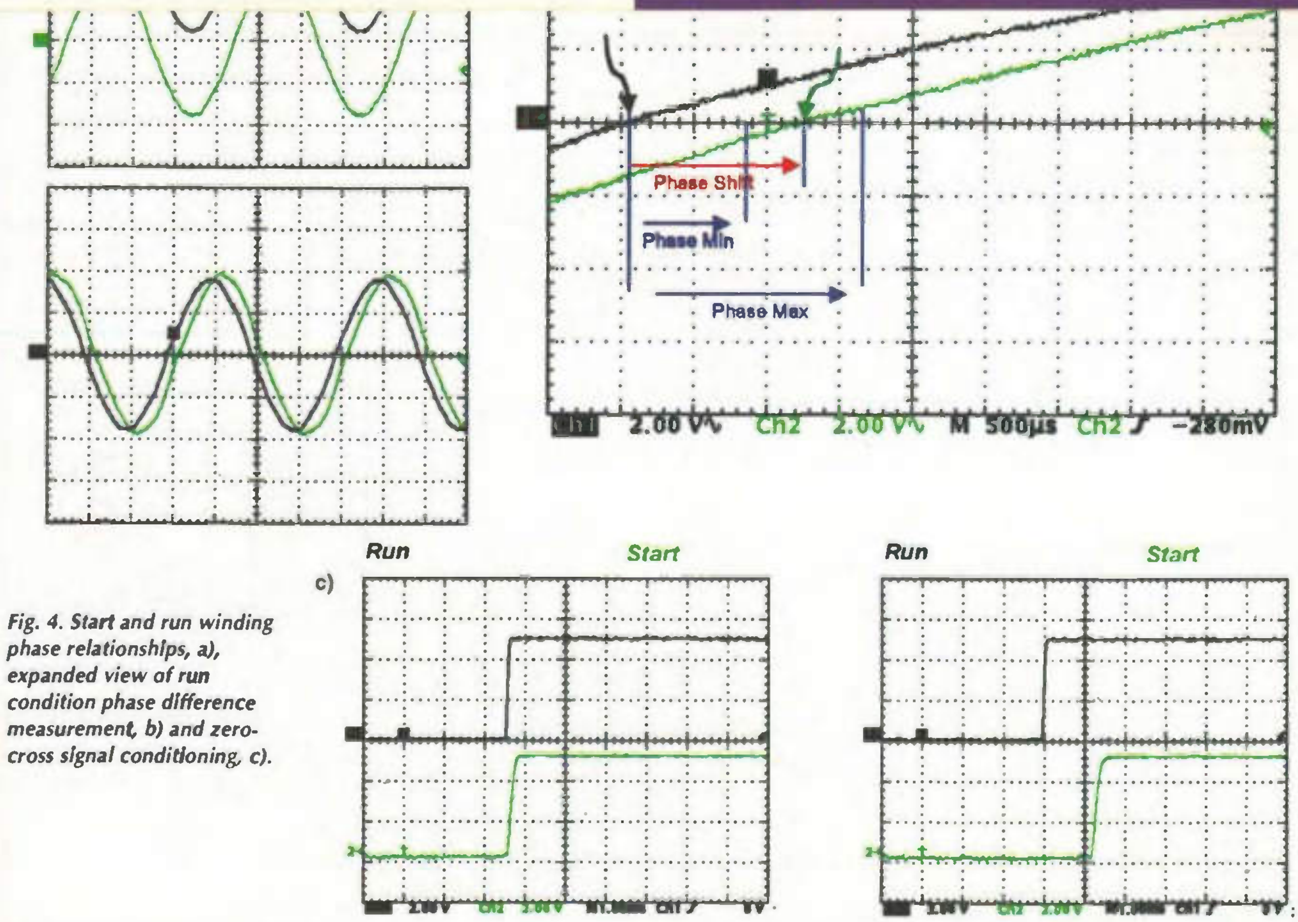


Fig. 4. Start and run winding phase relationships, a), expanded view of run condition phase difference measurement, b) and zero-cross signal conditioning, c).

Faraday's homopolar generator

The homopolar generator was discovered 170 years ago, yet no theory is available to give a convincing explanation of the results of the experiments described here, argues Dom Di Mario. You can replicate these experiments with magnets and a few pieces of copper.

It is unlikely that you will find the description of the Faraday homopolar generator in any university textbook on electromagnetism. The closest you will come across is the plain homopolar generator.

The fundamental difference between the two adaptations of the homopolar generator is that in Faraday's device, the magnet generating the magnetic field moves or rotates together with the conductor in which the voltage is supposed to be generated.

Faraday first reported its operation[1] in 1831. **Figure 1** shows a practical implementation of the discovery. A ring magnet together with a copper disc that's three or four times the diameter of the magnet are involved.

The disc is glued to the magnet and the combination is rotated while the voltage is measured between two sliding contacts. One contact is placed at the centre of the assembly, the other allowed to slide from the centre up to the edge of the copper disc.

The maximum reading takes place when contact A is at, or just beyond, the radius of the magnet. Actually, you will read almost the same voltage regardless of whether the contacts are between the centre and the radius of the magnet or between the radius of the magnet and the outer edge of the disc. Obviously the first solution is a more practicable implementation.

As a generator, this set-up seems to have only drawbacks: the output voltage is generally rather low, just more than 2mV with my set up using a 4V DC electric motor rotating between 20 and 40rev/s and a ring magnet of 36mm diameter. Other experimenters have reached hundreds of millivolts or more with powerful magnets and motors.

This is a DC machine. The presence of sliding contacts introduces a voltage drop that could make the generator totally useless, given the high current that it takes.

Yet, there are some rather puzzling aspects concerning this generator: as the magnet is moving together with the copper disc, you would expect that no voltage would be generated. In fact, the voltage is definitely there. It is the same as you would have with a stationary magnet and a rotating disc.

With a stationary magnet, the voltage can be explained using existing laws of induction. It is also possible to imagine that when the disc and the magnet rotate together, a voltage is generated in the wire connected to the brush; this happens regardless of its orientation, its size, whether it is reduced to a needle – even with the wire magnetically shielded, except for the contact point.

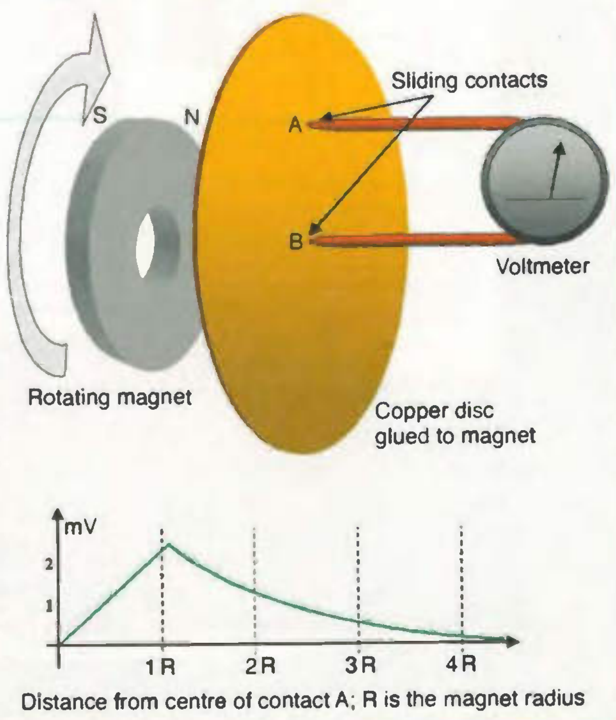


Fig. 1. Homopolar generator – there is no difference in the output voltage when the disc is rotated and the magnet is stationary or when both the disc and the magnet are rotated.

However the real issue surfaces when you consider the set-up's operation as an electric motor: you should expect to see rotation of the disc/magnet combination after a certain voltage is applied across the sliding contacts. From a practical point of view it will be quite difficult to send, say, hundreds of amps through the sliding contacts, but it is the theoretical implication that remains without a proper answer: against what is the magnet/disc moving?

The only solid material available is the sliding contacts, but the direction of the current is such that the disc appears to exert no force on them for a certain orientation, or if they are shielded. Are the disc and magnet moving without a counter force?

Conversely, if you use the device as a generator and apply a load, you are unable to see where the back-torque is exercised. The generator seems to generate electricity without a corresponding input loading, except for friction losses.

Because of the implications, you would expect a rush of experiments, but there is little activity in this area. Major contributions towards the implementation of a Faraday homopolar generator came from a series of experiments carried out by the late Bruce dePalma with his N-machine[2]. Despite the fact that the device showed no apparent increase of input power when a load was applied, there was no further development. This was probably because the results were not so clear cut and certain assumptions had to be made in order to compute the power losses.

Also Stefan Marinov[3] has been busy with the theory and construction of machines based on the same principle: his Modrilo and Siberian Coliu motors were two ingenious implementations, but he was duly ignored by his fellow academicians.

Contributions also came from P. Tewari with his 'Drive space generator'[4] and L. Mencherini[5] who tried to demonstrate that the magnetic field should be considered stationary outside a magnet even when it is in motion, an hypothesis put forward by Faraday himself.

A basic experiment

From an experimental point of view, there are other configurations that should be explored, with the hope of finding a clue that will provide a better understanding of the underlying mechanism.

If you use a common cylindrical metal magnet, you can dispense with the copper disc. There is a voltage between either of the magnetic poles and the centre point of the magnet. The exact position and shape of sliding contact B at the magnetic pole is not critical, although you get a higher voltage when the contact is placed off axis or right on the edge.

The voltage will decrease if contact A is moved away from its centre position between the poles until you will have no voltage when the contact reaches either end of the magnet. The maximum reading was 2.4mV using the same electric motor as in the previous experiment, with a magnet 25mm long and 6mm diameter.

With this configuration, you could make the same assumptions about counter torque as you would for an electric motor, but doing so would yield the same set of unanswered questions. The major difficulty in generating a usable amount of energy is to have a very low resistance for the sliding contacts.

Resistance of the contacts must be close to zero. If it isn't, the voltage drop across the contacts could easily be equal to the generated voltage – leaving nothing to drive an external load. This can be achieved only with the use of sliding contacts lubricated with mercury, as in the N-machine.

Even with the contacts lubricated in this way, the resistance of mercury could be too high if there is too much of it. This means that the brushes, or whatever mechanical solution is devised, must be machined with very close tolerance.

An interesting consequence of this experiment is that the Earth – a rotating magnet – should have similar properties, **Fig. 2.** It should have an electric field of equal polarity at the magnetic poles – the pink area – and the opposite polarity around the equator, the greenish belt. The sliding contacts could very well be the stream of charged particles coming from the Sun, thus creating a huge homopolar generator.

Generating a higher voltage

The low voltage, typical of this machine, can be increased by simply adding a magnet. It should be placed with its

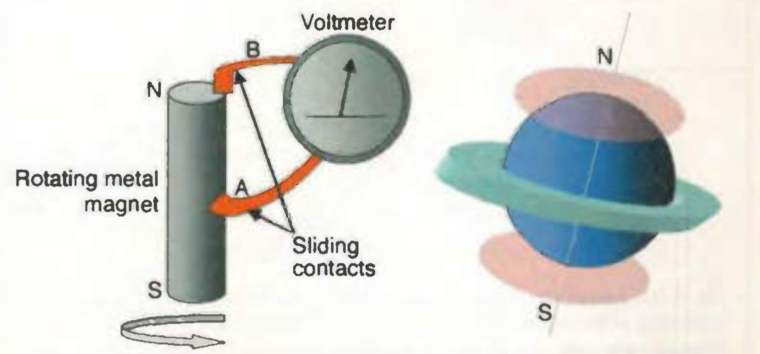


Fig. 2. The easiest way to verify Faraday's discovery is with a plain metal magnet. Its rotation will generate a voltage between either end and the centre of the magnet. Maybe the Earth behaves in the same manner.

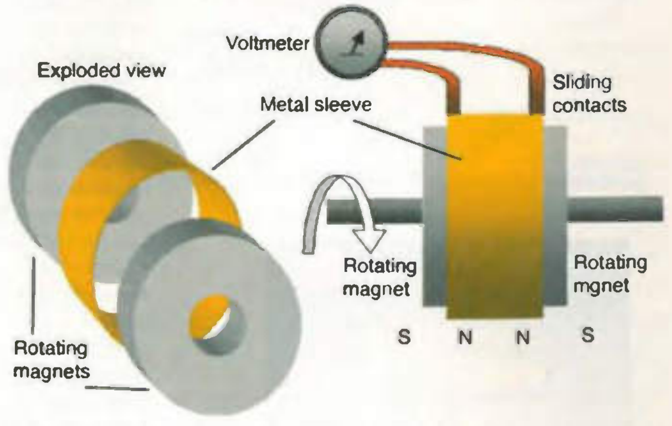


Fig. 3. At least double the voltage is available if two magnets are assembled with like poles facing each other. The distance between the poles is critical and some adjustment is necessary to find the best positioning.

south pole facing the copper disc of Fig. 1 and arranged so that the disc is in the middle. This solution will double the available voltage. An additional experiment was conceived with the purpose of providing a different mechanical configuration, Fig. 3.

Two ring magnets of the same type used in the first experiment were assembled so that like magnetic poles face each other. The distance was about 20mm and a copper sleeve was placed in such a way as to slightly overlap the like poles of the two magnets.

Voltage output was taken right at the edge of the sleeve as shown in Fig. 3. Using this set-up you will measure twice the voltage that you measured before. The additional benefit of this solution is that both sliding contacts are mechanically the same: this could be an advantage from a mechanical point of view as there is no need to connect electrically to a centre point.

In all the above experiments, I observed no decrease or variation of the voltage in spite of the magnetic screening. This screening comprised steel or ferrite tubing, which was applied to the wires connecting the brushes.

In order to avoid eddy currents, the brushes were reduced in size until they were just two needles. Doing this resulted in no change in the output voltage.

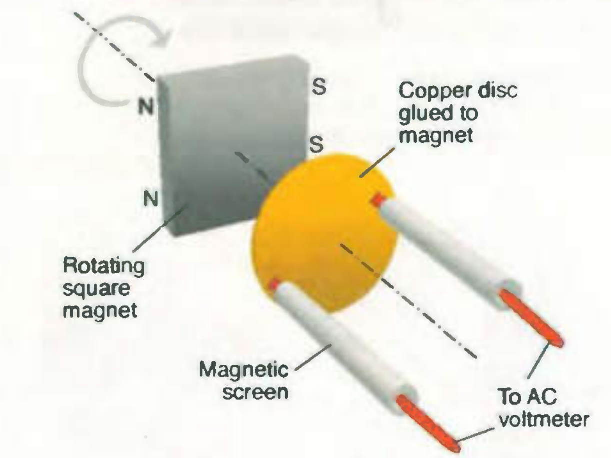


Fig. 4. AC generation is possible using both poles of a magnet. Magnetic screening of the wires connected to the brushes is essential in order to eliminate the voltage induced by the changing magnetic field.

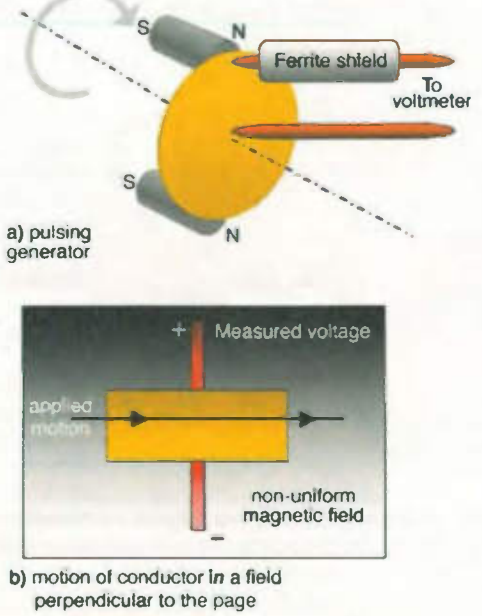


Fig. 5. In a), the same voltage as the configuration of Fig. 1 is generated if only the disc is rotated. When both the disc and the magnets are rotated you will measure a pulsing homopolar voltage. Fig. 1 is therefore only a special case of a more general law involving the movement of a conductor in an inhomogeneous magnetic field, Fig. 5b).

An AC generator

The fact that the Faraday homopolar generator is a DC machine is an inconvenience that limits its flexibility. **Figure** 4, is a suggestion of a machine generating an alternating current.

A square magnet, with north and south poles on the sides, is rotated and the voltage is taken with the collector brushes placed at the edge of the disc. If one of the brushes is placed in the centre you will have half the voltage.

One major problem with this generator is that there is also an induced voltage on the collector wires. This means that if you want to measure the homopolar voltage, you should try to get rid of the induced voltage.

I found getting rid of the induced voltage rather difficult. Screening the collector wires with ferrite tubing helped, but it didn't solve the problem fully. I was not able to totally eliminate the induced signal that was interfering with the signal generated with the homopolar effect.

There seems to be a 90° phase difference between the two signals but further investigation proved difficult due to the low level of the signal and the precarious mechanical set up I was using.

I tried a variation of the above AC generator using the rotor of a bicycle dynamo with an eight-pole magnet. Its outer circumference was covered with a layer of copper. The brushes were placed at an angle of 45°, or odd multiples of it, one from the other, this is the same angle as one magnetic pole relative to the other.

The unwanted signal from the induced voltage is even more difficult to control with this arrangement. Careful magnetic screening and positioning of the collector wires was necessary in order to detect the homopolar voltage.

As expected, the output voltage depends on the speed of the rotor and the intensity of the magnetic field. What was not expected was the fact that the largest voltage is present when one of the brushes is placed where the magnetic field is most inhomogeneous while the other brush is located where the field is mostly homogeneous or absent.

A special case is the machine of Fig. 3, where both brushes are placed within an inhomogeneous field but they both refer to a zero potential, which is a point in the middle between the two brushes. Actually I found that there is an ideal distance between the magnets of Fig. 3 where the voltage is at its highest. It is likely that the magnetic lines are more or less deformed, hence the resulting field is more or less homogeneous.

A non-uniform field is necessary

The results of all the above experiments seem to obey the law whereby a conductor generates a voltage across it when it moves in a non-uniform magnetic field, Fig. 5b).

I repeated the basic Faraday experiment, Fig. 5a), but this time with one cylindrical magnet providing a magnetic field that does not extend on the whole disc. Note that two magnets are shown in order to have a mechanically balanced system.

The sliding contact is next to the magnet. Rotation of the disc will give the same voltage as in Fig. 1. Rotation of the disc and the magnets however does not give the same voltage but a pulsing unidirectional voltage – once you manage to remove the induced signal.

It is now the wire connected to the sliding contact that is immersed in a non-uniform magnetic field that generates the homopolar voltage.

The original Faraday experiment, Fig. 1, with the disc and the magnet moving together, is a very special case. It could lead you to think that it is the sliding contact that is actually moving in an inhomogeneous magnetic field and becomes the source of a continuous voltage, because the field is always present.

In Fig. 5b) you could move the conductor and the magnet originating the field together, but you would still have a DC voltage. This is the same as having a stationary conductor/magnet and the sliding contacts moving in a non-uniform magnetic field.

It is now clear where the voltage is generated. Any magnetic shielding of the wire will still preserve the non-uniformity at the contact point, so it will make no difference. Surely it will make a difference when the mechanical forces acting on the various parts are considered? But as yet we have no satisfactory answers. Further experimentation is necessary.

The Faraday homopolar generator could be a viable alternative to generators currently in use. However, any serious implementation would require superconducting magnets and specially designed brushes. Until then, the machine is bound to remain a laboratory curiosity or, possibly, something to be totally ignored because of its apparently inexplicable features. ■

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NEWPRODUCTS

Please quote *Electronics World* when seeking further information

Miniature FM radio modules

Quasar has launched the first miniature FM radio transmitter and receiver modules to offer communication rates of 128kbit/s at distances of up to 400m. The new QFMT5-434 transmitter and QFMR5-434 receiver modules simplify the implementation and reduce the component count of one-to-one and multi-node wireless links and provide 'drop-in' replacement of lower speed modules. Applications include building security and car alarms, remote industrial process monitoring, computer networking and electric gate openers, with small size and low power consumption making the modules ideal for battery-powered, portable designs.

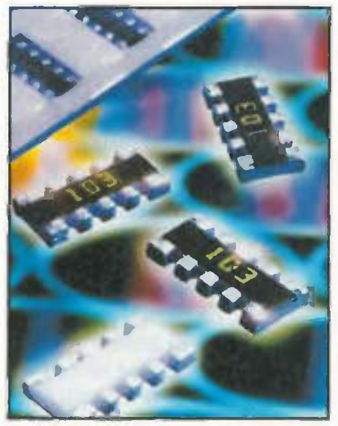
Supplied in a fully shielded miniature SIL package, the QFMT5-434 transmitter features no adjustable components, offers good resistance to shock and vibration and delivers a temperature-compensated RF output from –20°C to 55°C. The QFMR5-434 high-sensitivity receiver is based on a single conversion FM Super-heterodyne design with double-RF filtering that includes a SAW front end for maximum sensitivity. Analogue and digital outputs allow for direct connection to a microprocessor or additional processing circuitry, while a received signal strength indicator (RSSI) provides a DC output voltage that is proportional to the RF input signal.

Quasar
Tel: 01273 898000
www.rfsolutions.co.uk

Chip-resistor network

The latest chip resistor networks from Rohm Electronics integrate eight resistors into a single ultra-miniature package to help designers save space and reduce component count while simplifying both circuit layout and final assembly.

Ideally suited for voltage pull-up in bus-based systems, the



MNR15 is a 10-pin chip resistor network that measures just 3.2x1.6x0.5mm and replaces eight conventional 0402 resistors. Resistance values range from 56Ω to 100kΩ and tolerances are ±5% in all cases.

Rohm's chip resistor network is rated for a maximum overload voltage of 50V. The operating temperature for the devices is –55 to 125°C, while rated power is 0.032W at temperatures up to 70°C.

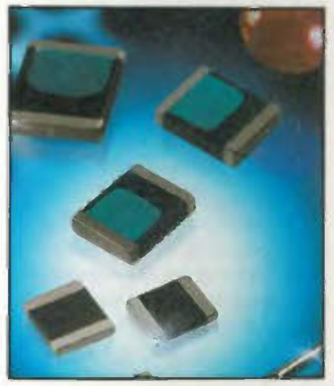
Rohm Electronics
Tel: 01908 282666
www.rohm.co.uk

Surface-mount varicons for transient suppression

A range of Keko varicon devices, designed to provide circuit protection against surge voltages and electrostatic discharge effects in electronics, automotive and telecommunications equipment.

Varicons are a new generation of multilayer zinc-oxide based protective devices that combine the variable-resistance properties of varistors with variable capacitance.

The Keko range includes the standard ZV Series, covering



Oscilloscope sends e-mail

Yokogawa's DL1740 SignalExplorer is a compact 4-channel digital oscilloscope combining a sample rate of 1GS/s, an analogue bandwidth of 500MHz and a memory length of 1Mword.

Despite its high-range performance, the instrument, is small enough to occupy an A4 size footprint on a test bench and weighs only 5.5kg.

A 500MHz bandwidth is achieved via a dedicated single-chip input amplifier. The fast screen update rate of 30 times a second – even at 1Mword record length – results from using a dedicated high-speed data-processor set known as the Data Stream Engine.

The combined effect of these new circuits is a reduction in component count, very low power consumption and minimal heat sink requirements.

The DL1740 uses a 6.4-inch TFT display. The repetition frequency of waveform data can be determined by using brightness, thus emulating analogue oscilloscopes.

The instrument is equipped with USB compliant interface as standard, allowing keyboards and USB printers to be connected. The ability to use a keyboard has improved the instrument's ease of use and screen images from the colour display can be easily output to a colour printer. The Waveform Viewer software provided allows the user to exchange waveform data and screen images between the oscilloscope and a PC.

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

DC voltages from 3 to 170V in chip sizes from 0603 to 2220 with maximum single-pulse (8/20μs) surge currents up to 1200 A; the ZVE Series for ESD suppression, with low capacitance values and short response times in chip sizes from 0603 to 1210; and the low-profile medium-voltage DV Series, for voltages from 60 to 300V RMS in chip sizes 3225 and 4032 with maximum single-pulse (8/20μs) surge currents up to 1200A.
VTM (UK)
Tel: 01494 738608
www.vtm.co.uk

Circular connectors

TTI has introduced two ranges of circular connectors from Amphenol. The first of these, the C16-1 series connectors are designed to meet the high requirements of industrial applications under harsh environmental conditions. Available as straight or right-angled versions, the connectors offer 3+PE (400V/16A) and 6+PE (250V/110A) contact arrangements for power and signal applications.
With housings made from high grade plastic material, the connectors are protected against accidental unlocking by a threaded coupling. The range includes versions with screw, solder and crimp terminations. The Amphenol C091 range is a miniature circular for signal transfer applications. Offering a number of contacts from 2 to 8, 12 and 14, the connectors feature a metal threaded or bayonet coupling and offer complete EMI shielding in the fully mated and locked position.
TTI
Tel: 01494 460000

One piece and high density interfaces

Samtec has introduced the FSI and SEI one piece and MSB and CSB high density interfaces. The one piece interfaces are on 1mm pitch with up to 100 contacts and a mated board spacing of 3mm (FSI) or up to 30 contacts for a mated board space of 1.65mm (SEI). These interfaces have phosphor bronze contacts and are for MCM interfaces, high shock and vibration applications. The high density interfaces have up to 500 i/os on 1.27mm pitch with profiles down to 3mm between mated boards. Tiger-eye beryllium copper contacts are used. Solder ball attachments are available.
Samtec
Tel: 001236 739292
www.samtec.com

Computing platform for customised and embedded systems

Sun Microsystems has announced a processor board to complement the Netra family of servers. The Netra AX1105-500 PCI board helps embed the Ultrasparc IIe processor and the Solaris operating environment into applications where a traditional server cannot be used due to environmental or space requirements. The board is for commercial and industrial products and systems in telecoms, aerospace, medical imaging and banking applications. It delivers 500MHz performance in a profile that can support 1U, 2U or ATX chassis designs. The motherboard is a standard ATX form factor board, powered by the 64-bit Ultrasparc IIe processor. The board also supports Wind River's real time OS VxWorks for embedded systems.
Sun Microsystems
Tel: 01276 416379
www.sun.com

Audio analysing DMM

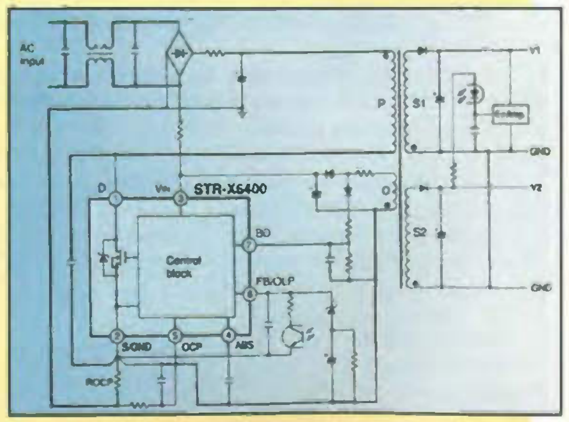
Keithley Instruments has announced the Model 2015-P audio analysing DMM for audio quality testing in telecoms applications in the 300 to 4000Hz human voice spectrum. It can characterise a signal spectrum without a separate computer or analysis software, can report frequency and amplitude of the highest value within a specified frequency band, and can identify additional peaks in descending order of magnitude. Once a peak frequency component is identified, it can give the magnitude and frequency of maximum components above and below the peak value. If desired, the instrument can also report the magnitude of a specified set of frequency components or determine the difference in amplitude between two spectral components. Audio measurements include total harmonic distortion, THD+noise, signal-to-noise plus distortion and frequency spectrum analysis. Filter types include CCITT weighted, C message, A weighted, CCIR and CCIR-ARM audio shaping. Selectable high and low-pass filters can be combined to form a band-pass filter. It also provides a dual-output 20Hz to 20kHz sine wave generator. The second output inverts the first output for testing differential input circuits for common mode or noise cancellation, or it can be used as a pulse train for synchronisation of other instruments with the model's source output.
Keithley Instruments
Tel: 0118 929 7500
www.keithley.com

ADSL line-driver hybrid

Advanced Power Components (APC) has released an ADSL line interface hybrid component for high-density CO and DSLAM line card applications. Developed specifically for use with Alcatel Microelectronics' 20450 Quad ADSL Central Office chip set, APC78147 is a dual channel device containing all of the analogue components required between the AFE chip / line driver and the line connector. The device is suitable for high volume pick and place SMD assembly and offers a significant reduction in the per square inch PCB area required per port, increasing the number of channels possible per line card. Performance has been validated across the full industrial temperature range from –40 to +85 C.The APC78147 is available for sampling from April 2001.
APC
Tel: 01634 290588
www.apc-plc.co.uk

Multi-mode off-line switch-mode PSU IC

The STR-X6400 Series from Allegro MicroSystems is a multi-mode off-line switch-mode power-supply integrated circuit which automatically changes its operating mode to ensure maximum efficiency over different output loads. The new circuit can operate in any one of four modes: quasi-resonant; quasi-resonant plus bottom cycle skip, PRC (pulse ratio control with fixed off-time); and burst mode. It detects the output load and selects the appropriate mode automatically, achieving an efficiency of up to 90% with 120W output
The design of the new circuit means that additional standby power is unnecessary,with the standby input being one-third less than on existing SMPS power ICs. A hysteresis function is included to stabilise the bottom cycle operation, and a step-drive function is used to control the turn-on surge current of the MOSFET.
Allegro MicroSystems
Tel: 01932 253355
www.allegromicro.com



110W quad output AC/DC power supply

Artesyn Technologies has launched a compact quad-output ac-to-dc power supply that is ideal for low-power applications involving peak current demands.
The new NLS110-9602 has a power-output capability of 80 watts – which can be boosted to 110 watts with 20ft³/min of forced air cooling – and is power factor corrected to comply with the EN61000-3-2 standard for harmonic current content. The unit provides +5.1V, +24V, +12V and –12V outputs with maximum continuous ratings of 10A, 4.5A, 5A and 1A respectively. However, the power supply is also capable of delivering very high peak currents.
The NLS110-9602 is an open-frame, low-profile design, with a small 7.0 by 4.25 inch footprint. The power supply features a universal input that accommodates any voltage in the range 90 to 264V AC at any frequency from 47Hz to 440Hz. The 5.1V DC output is equipped with overvoltage protection, and all four outputs are comprehensively protected against short-circuit conditions, with full auto-recovery. An optional Power Fail Detect signal is available.
Artesyn's NLS110-9602 power supplies fully comply with the rigorous EN55022-A radiated noise standards, and meet all applicable immunity standards, including EN61000-4-2, -3, -4, -5 and -6 levels 3.
Artesyn Technologies
Tel: 353 24 25572
www.artesyn.com

Quartz device for UMTS and CDMA

Epson has introduced the TG-2820CB temperature compensated crystal oscillator for CDMA and UMTS applications. Frequency stability is ±4.5 x 10^{-6} over a temperature range of –30 to +80°C. Available in frequencies between 12 and 19.8MHz, it has 1.5mA maximum consumption at a supply level of 2.8V. Footprint is 3.2 by 2.5mm and height 1.5mm. It comes in an SMD ceramic package.
Epson Electronics
Tel: 0049 089 14005277
www.epson-electronics.de

Pan and tilt surveillance camera kit

A range of mini dome cameras with automatic pan and tilt are available from Stortech Electronics. The PT range of surveillance cameras comes in a variety of camera options and is ideal for internal use such as retail store monitoring or where close commercial surveillance might be required.
Once installed, the controller enables a full 360° panning action and a 90° tilt from the horizontal.
Stortech Electronics
Tel: 01279 419913

Efficient white-LED driver

Powered by multiple or single battery cells, the ZXSC300 converter circuit provides a programmable constant-current drive for white LEDs and maintains an efficiency of 94%. The combination of a ZXSC300 controller IC and an FMMT617 low saturation transistor feeds single or multiple white LEDs, connected in series or parallel, from a wide range of input voltages. Helping to maximise valuable battery life, the high efficiency conversion produced by the circuit is sustained as cell voltage descends to as low 0.8V. The ZXSC300 converter suits a broad range of portable equipment using white LEDs in backlighting and illumination roles, from mobile phones to flashlights.
Zetex
Tel: 0161 622 4444
www.zetex.com

Power supplies for embedded systems

Diamond Systems has introduced two Jupiter-MM dc-to-dc power supply modules for PC/104 embedded systems in air and ground vehicle, process control, military, aerospace and medical equipment applications. The supplies accept input voltages from 8 to 30V DC, making them compatible with 12 and 24V systems. Output power is provided directly onto the PC/104 bus connectors, as well as on an auxiliary connector for external tapping. The MM-LP-XT has a +5V output and maximum power output of 25W. The MM-SIO-XT has ±5 and ±12V outputs plus two serial ports. Maximum power output is 50W. This version has up to 10A that can be provided on the +5V output or the 50W of total power can be distributed across the four outputs. The board is jumper configurable to support RS232, RS422 and RS485. Efficiency is 92 per cent.
Diamond Systems
Tel: 001 510 456 7800
www.diamondsystems.com

Quad LVDS line driver with flow-through pin out

Maxim Integrated Products has introduced the MAX9123 quad, low-voltage differential signalling line driver with flow-through pin out. It accepts four TTL or CMOS input levels and translates them to LVDS output signals that conform to the Ansi TIA and EIA-644 LVDS standard. Applications include digital copiers, laser printers, cell phone base stations, telecoms switches and network switches and routers. The flow-through pin-out reduces crosstalk by separating the inputs and outputs. Two TTL and CMOS inputs can be used to set all four outputs to a high-impedance state and reduce power consumption to 16mW typical.
Maxim
Tel: 0118 9303388
www.dbserv.maxim-ic.com

Short-circuit protection for relay

The AQY210KS Photomos relay from Matsushita Electric Works has a short-circuit function to protect sensitive equipment, such as modems, telephones and measurement instruments, from high current loads. When the load current exceeds specifications, it goes into an open state and cuts off the input current within 50μs. Because there is almost no

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heating, the relay remains intact. To make the circuit available again, the input current is reset. The relay comes in a four-pin small-outline package measuring 4.3 by 4.4 by 2.1mm. It can handle a continuous load current of 120mA at voltages up to 350V AC, with a channel resistance of typically 23Ω. The relay is suitable for switching low-level analogue signals and a typical application would be in modem circuit protection.
Matsushita
Tel: 01908 231555
www.matsushita.co.uk

32-bit RISC microcontroller

The 32-bit RISC microcontroller is based on a SH2 CPU core and features 160Kbyte of on-chip, full speed Flash memory. The SH7018F offers a compact peripheral set and modern 0.35μm manufacturing processes.
It is suitable for consumer, industrial, office-automation and PC peripheral applications and is also targeted throughout the distribution network.
The SH7018F's SH2 CPU offers sixteen 32-bit wide general registers, an on-chip multiplier and a five-stage pipeline, which enables it to achieve 20MIPS Dhrystone performance at an operating frequency of 20MHz (50ns instruction cycle time).
With 160Kbytes of on-chip, high speed and single voltage flash memory, the CPU is able to execute instructions with no wait states. The peripheral set comprises of a three channel 16-bit timer unit (MTU), an additional 16-bit free running counter (CMT), an 8-bit interval timer (TIM2), a watchdog, one USART and an analogue-to-

digital converter (ADC). There are also a maximum of 62 I/O pins plus 8 input-only pins, some of which provide external external interrupt capability. Seventeen of the I/O pins run at 5V, while the logic and other I/O pins run at 3.3V.
Hitachi
Tel: 01628 585163
www.hitachi-eu.com

CMOS SRAM with low data retention voltage of 1.5V

NEC's latest 1Mbit Static RAM (SRAM) is designed using full CMOS process memory allowing it to operate with supply currents as low as 25mA (max) and with supply voltage 1.8 - 3.6V, data is retained at voltages as low as 1.5V. Offering high-speed access times down to 70ns and operating in an extended, –25 to +85 degrees C temperature range, the CMOS SRAM is an ideal memory for a wide variety of industrial applications and mobile phones.

Designed using a 0.32-micron rule, in a 128K-word by 8-bit organisation, the SRAM is available in 3 varieties (B, C and D versions) to suit a variety of industrial applications: B offers the fastest access times of 70ns with supply voltages from 2.7V; C offers access times down to 100ns and supply voltages down to 2.2V; D offers the lowest supply voltage of 1.8V with access times down to 120ns. C and D versions retain data with a supply voltage as low as 1.5V.

The CMOS SRAM is available in a 32-pin STSOP (I) (8x13.4mm) package and to enable extended capacity, the SRAM comes with two chip enable inputs/CE1 and CE2.

Compared to DRAM, SRAM offers faster access times and does not require refresh circuitry but previously has been relatively power hungry - limiting the applications of SRAM in telecoms circuits which rely on the circuit's residual power.
NEC Electronics
Tel: 01908 691133

Cold-cathode quad lamp inverter

Taiyo Yuden has announced an inverter to drive four cold-cathode fluorescent tubes for LCD backlighting. The LS680 is 10mm high, 160mm long and 34mm wide and provides an input range of 10.8 to 13.2V DC and a total output of 16W. With a built-in brightness control, the module provides dimming capability by PWM or DC voltage. It uses entirely surface-mounted components.
Taiyo Yuden
Tel: 01494 464642
www.yuden.co.jp

Rail-to-rail i/o op-amps

The LT1672, LT1673 and LT1674 are Linear Technology's single, dual and quad rail-to-rail input and output, precision operational amplifiers that draw 1.5 and 2μA maximum supply current. These decompensated amplifiers have a 12kHz gain bandwidth product and 4μV p-p input voltage noise from 0.1 to 10Hz. They do not stress the supply during start up. The supply current is controlled during power on and is limited to about 2μA typical. Supply voltage range is from 2.2 to 36V and protection against reverse battery connection is to 18V, making them suitable for use in 12V lead acid battery applications. The over-the-top feature lets the inputs function

correctly when one or both are above the positive supply, making them suitable for high side current sensing applications. They let a low voltage system sense a higher voltage supply. Maximum input offset voltage is $375\mu V$, offset voltage drift 0.4 to $2\mu V/°C$, open loop gain 215V/mV, CMRR and PSRR 90dB minimum and input offset and bias current 20 and 250pA respectively.
Linear Technology
Tel: 01276 677676
www.linear-tech.com

Clock generators suit embedded communications

AMI Semiconductor has introduced a family of one- and two-phase-lock loop clock ICs. They are for fixed-frequency systems such as digital cameras, computer peripherals, printers, LAN hubs and Gigabit Ethernet hubs. Several are derived from common platforms. The platforms can be customised for frequencies, pin outs and features in the final processing layers, similar to a gate array. Each device uses an on-board crystal oscillator to generate a reference frequency, which is used by the PLLs to generate precision output frequency ratios (ratiometric derivation). The crystal oscillator is suitable for any fundamental-mode crystals in the 5 to 30MHz range. For video systems, which cannot tolerate unpredictable artifacts, dual-PLL models are available that generate three output clocks from the crystal reference by driving the frequency through an array of post-dividers. These three outputs are locked to reduce phase noise, jitter and EMI from harmonic stacking.
AMI Semiconductor
Tel: 0049 3513 15323
www.amis.com

Various voltages from one package

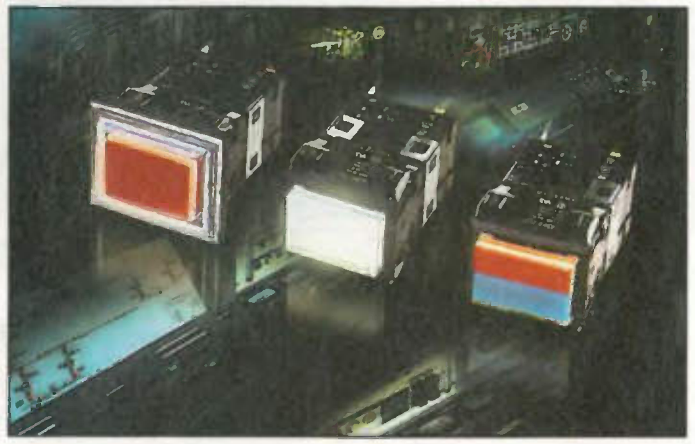
AVX has improved its Bestcap electrochemical capacitor to provide various voltage options from one package. The standard ranges of 3.5, 4.5 and 5.5V with custom ranges from 1.5 to 10V are available. Capacitances are between 50 and 560mF, ESR from 25 to $230m\Omega$ and maximum leakage current 5 to $40\mu A$ depending on size. Package options are 28 by 17 by (2.9 to 6.0)mm and 48 by 30 by (2.9 to 6.0)mm with moulding wall construction down to 0.7mm.
AVX
Tel: 01252 770000
www.avxcorp.com

Push-button switch with integral lamp

Omron has introduced a split screen lighted push-button switch for process control applications. The A3PN IP40-rated switch is available in rectangular and square models, measuring 32mm deep. The DPDT switch, which uses an LED, is available in seven solid colours including red, white, blue and pure white. Options include a horizontal split screen and an indicator only version (M2PN series). Rated voltage is 5, 12 or 24V DC. A 110V DC model with voltage-reduction circuit is also available. Operating range is –10 to +40°C. Options include switch-guard, seal cover and character plate.
Omron
Tel: 020 8450 4646
www.omron.com

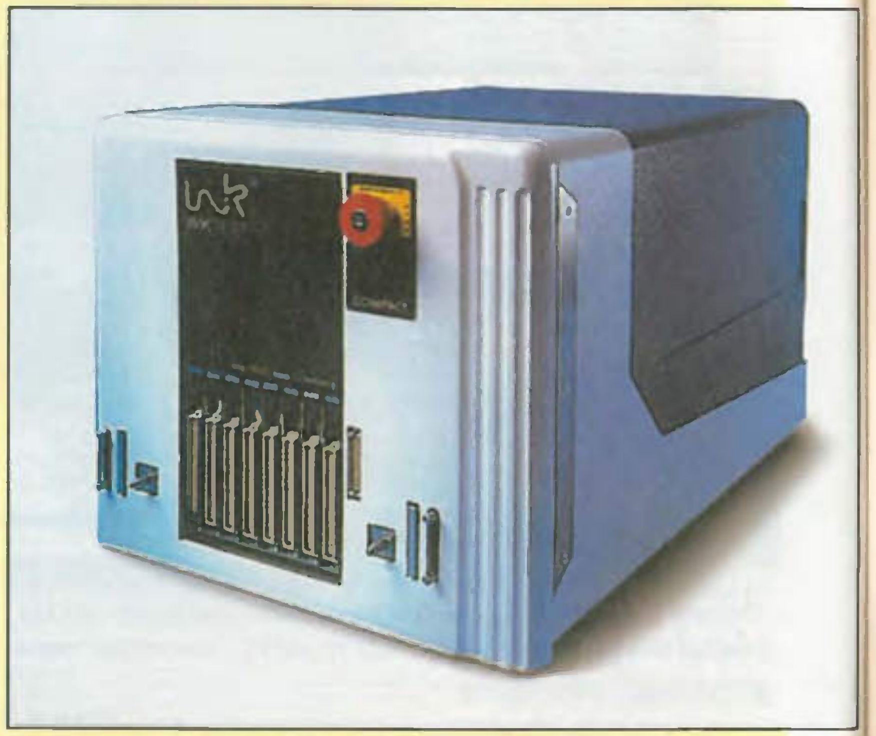
Blade and beam down to 0.4mm pitch

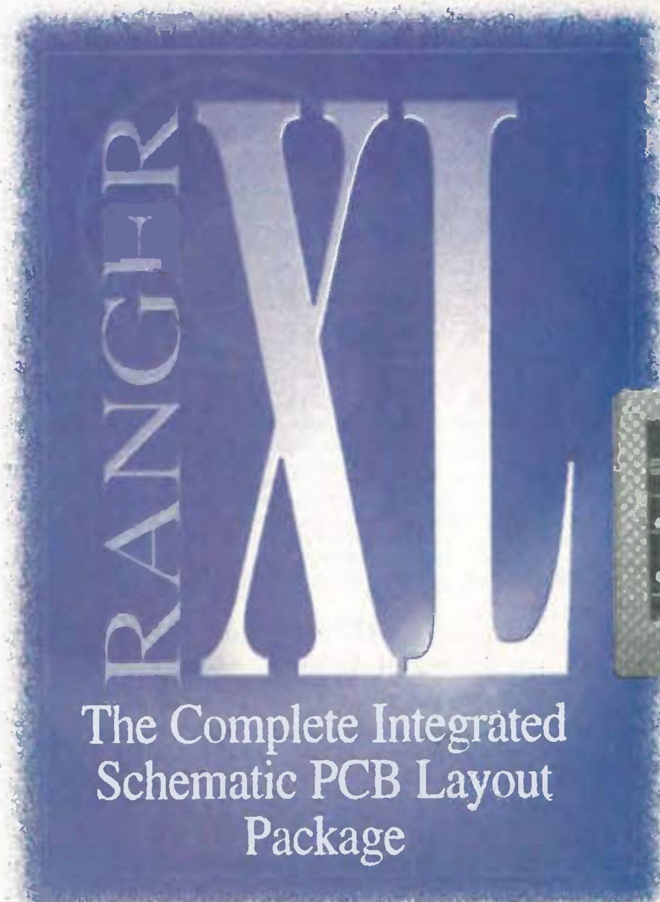
Connectors from Deltron Roxburgh for board stacking systems have pitches down to 0.4mm on a blade-and-beam construction. The Samtec BTF headers and BTS sockets have



Eight-slot test system for function test

WK Test has introduced the 5908 Compact tester, an eight-slot member of its 5900 family of functional and combinational testers for high-volume contract manufacture and consumer applications. It accepts all the firm's modular test instruments, so it can be configured for other test requirements. A Windows NT interface provides a multitasking environment. Features include network integration and data-logging. Remote control over the LAN, WAN or Internet is possible using optional utilities. The system supports up to two variable and four fixed power supplies. Each variable supply provides up to 400W. The eight identical system slots can accommodate any of the firm's more than 40 modular test instruments, which provide functions such as analogue and digital measurements, analogue waveform simulation, digital simulation, data acquisition, microprocessor and ROM emulation and communications simulation. Custom modules are also available. It comes in free-standing or rack-mount versions. The free-standing unit measures 400 by 600 by 360mm.
WK Test
Tel: 01243 825811
www.wktest.com





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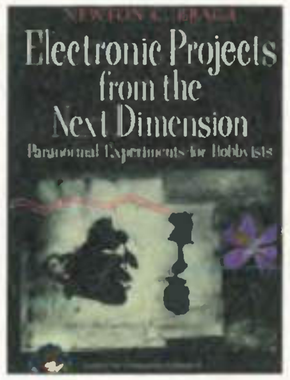

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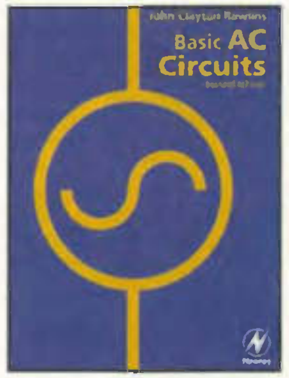

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Fairchild Semiconductor
Tel: 01793 856856
www.fairchildsemi.com

Lithium-cell monitor IC

Texas Instruments has announced a battery monitor IC for Li-ion- and Li-polymer-based applications in mobile hand sets, Internet audio appliances and digital cameras powered by a single lithium cell. The device measures critical parameters in a battery pack, such as charge, discharge, self-discharge and temperature and employs an auto-calibrating VFC (voltage-to-frequency converter) for continuous charge/discharge integration. With information from the measurement IC, a host system can calculate remaining battery capacity and predict remaining system run or standby times. The bq2023 battery monitor communicates measurement information to the host controller using the SDQ serial interface, a single wire command-based protocol compatible with the Dallas Semiconductor 1-Wire. The part also includes 224 bytes of flash memory and a 64-bit ID ROM register. The non-volatile memory integration allows a system designer to add a fuel gauge to a system while replacing serial EEPROMs or ID chips, which are currently used in battery packs to store battery pack characteristics or identification and addressing codes. It is packaged in an 8-pin TSSOP. With a 25mΩ sense resistor, the resolution of the converter is better than 0.12mAh.
Texas Instruments
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www.ti.com

SM shielded-drum core inductors

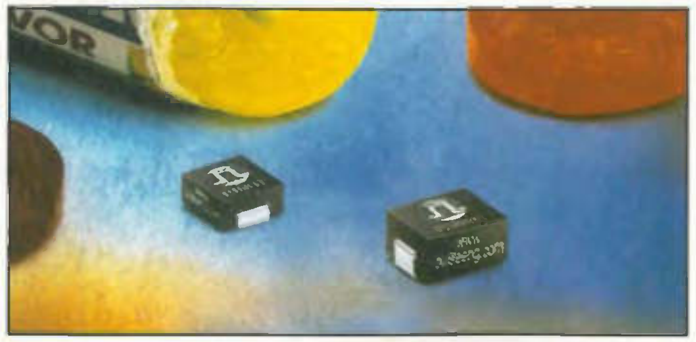
Available from Ultimate Renaissance are two SMT shielded drum core power inductors claimed to give an EMC performance comparable with a toroidal core. Applications are in switching power supplies and dc-to-dc converters. The Pulse P1172 and P1173 have inductances from 2.7 to 47μH and DC rating from 2.5 to 10.0A. Package outline is 12.0mm square footprint and 8.0mm height. Weight is 4.5g. They are available on tape and reel.
Ultimate Renaissance
Tel: 01793 439310
www.ur-home.com

0805-profile SMT LED for PCBs

Ledtronics has introduced the Surfled 0805-profile LED for SMT processes. It measures 2 by 1.25mm, comes in six colours and several related hues, is available on EIA standard 8mm tape on 2.1m reels, and has a viewing angle of 140°. It operates from 20mA. When used with the firm's Optiled light pipes, it simplifies the spatial relationship between the PCB and remote indicator illumination points by channelling all the LED-generated light to where the light is required. There are two versions – high efficiency and super bright. Applications include back lighting computer panels, hand-held instruments, and datacoms and telecoms status indicators.
Ledtronics
Tel: 001 310 534 1505
www.ledtronics.com

Surface-mount power inductor

Pulse has announced two surface-mount power inductors for use with Volterra's low-voltage power delivery semiconductors. With a profile height of only 0.125 inches (3.17mm), the compact SMT power inductors are designed for high-current, low-voltage, low-profile dc-to-dc power converter applications which use latest generation microprocessors. The package used by Pulse for these devices is the industry's only flat-top, self-leaded design with a fixed clip, which is ideal for easy pick-and-place applications and higher current ratings. The inductors are currently available in tape and reel packaging and are magnetically shielded for reduced EMI. ■
Pulse
Tel: 01483 401 700



High-impedance RF probe

Having completed his 50Ω true-RMS-reading RF millivoltmeter, Cyril Bateman went on to design this add-on probe to provide a high-impedance, low-capacitance input.

When measuring in-circuit signal voltages, it is essential that the method used presents a high impedance, low capacitance load to the circuit under test. Measuring in-circuit direct voltages presents little difficulty. Making AC in-circuit measurements is quite a different matter – especially at high frequencies.

If you examine the techniques used to measure alternating voltages, you will find that most methods have limited frequency range. Average-responding millivoltmeters involving rectifying techniques[1] can be used up to perhaps 10MHz. Most RMS-responding meters though, based on ICs using the 'implicit' solution method,[2] are limited to 1 or 2MHz. There's more on this in the panel entitled 'RMS'.

Because of frequency-range restrictions, many designers are forced to rely on their oscilloscope to measure AC voltages. Many oscilloscopes measure to at least 100MHz. Because of capacitive loading though, most test probes inhibit measurements of low level signals at high frequencies.

My Coline M12SW oscilloscope probe has a bandwidth of 250MHz and 1.4ns rise time. When set to divide by ten, it applies 18pF loading to the circuit. Set to times one, its applies 60pF to the test circuit. The probe's bandwidth reduces to just 10MHz and rise time increases to 35ns. By 100kHz, square-wave signals are visibly distorted.

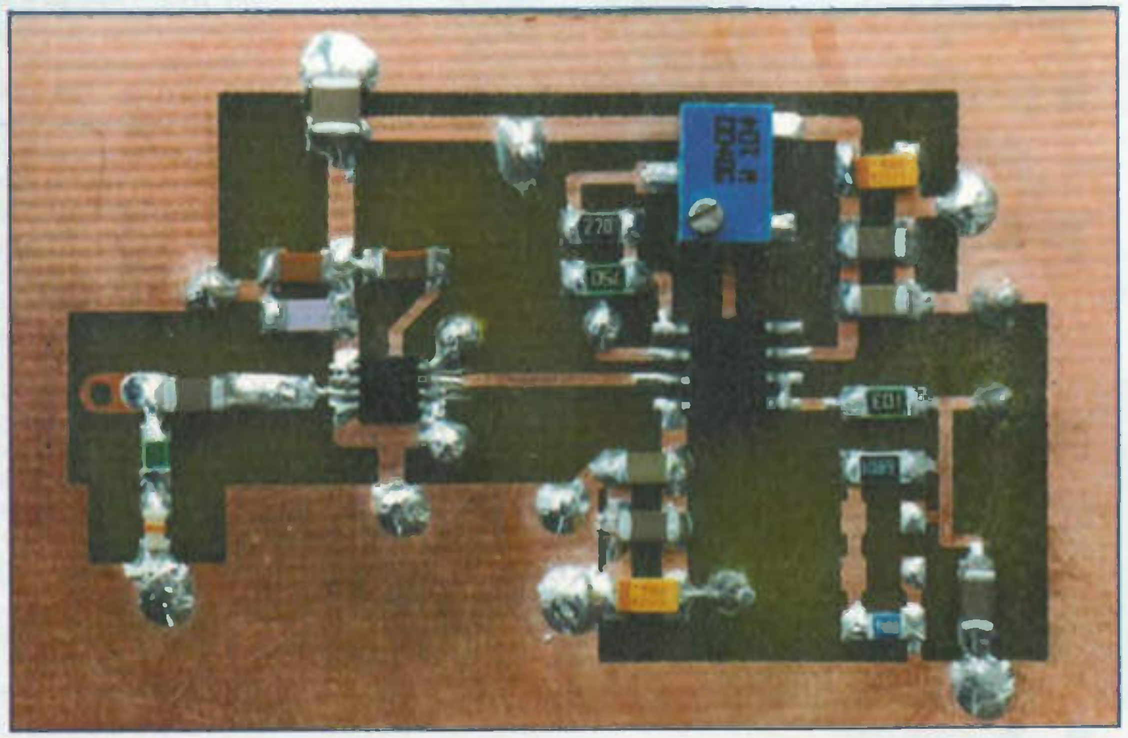


Fig. 1. Using the AD8361 to accurately measure true RMS from 10kHz to 100MHz in a 75Ω system. This is a double-sided printed board originally designed for my 50Ω meter. It provides an excellent match to 75Ω with R_1 changed to 110Ω, R_2 to 0Ω and L_1 to 12nH.

While at low frequency this probe exhibits greater than 1MΩ of impedance, by 10MHz this reduces to 800Ω when set to divide by ten, and 265Ω for divide by one.

Many scope probes impose much heavier loading. An HP9100 probe for example, set to times one, applies a 150pF load to the circuit under test.

Measuring signal levels less than 500mV, you may be forced to set the probe to divide by one in order to obtain an adequate display. Many modern ICs become unstable or decidedly peaky when loaded with such capacitance. At best the probe's capacitive load attenuates the measured signal.

In order to measure low-level signals the probe's capacitance to ground must be reduced. In practice this means using either a 50Ω measurement system, or a low-capacitance active probe.

In the August 2001, issue I described a simple but accurate circuit with a 50Ω input impedance.[3] It is able to measure RMS voltage to 100MHz. Equipped with a suitable high-impedance, low-capacitance 'front end', this design will solve many measurement problems, Fig. 1.

Commercially-available active probes can be very expensive, so I decided to investigate the design of a suitable 'front end' impedance-converting, probe.

High-Z probe

Much modern circuit design uses supply voltages of ±5V or less, resulting in maximum peak signal levels of around 3V. A target input voltage of 2V RMS would suit most circuits and maximise the dynamic range of my AD8361 based true RMS meter.[3]

Having reviewed a number of design alternatives, including discrete FET followers and high impedance ICs, I decided to seek an IC solution. For ease of use, this would need to be able to drive a metre-long coaxial cable, connecting the probe to my RMS meter. To match my RF RMS voltmeter design's performance, I wanted a flat response to 100MHz.

Choosing the ICs

Following a review of ICs available from mainline distributors, I decided to base my probe design on the Maxim[4] MAX4005. This is a unity-gain, very-high-frequency, JFET input IC with an internal 75Ω precision thin-film output resistor.

This chip provides a ±0.1dB gain flatness of 60MHz and ±0.2dB to 80MHz, a 350ps rise time and 950MHz bandwidth. Its 10pA input bias current and 2.2pF input capacitance are ideal for use in a high-input-impedance, low-capacitance probe.

Attenuating the 2V input to 1.3V into the MAX4005 and using a correctly-terminated 75Ω coaxial cable, would provide the required 650mV maximum input to my AD8361 RMS meter.[5] This level ensures that its optimum dynamic range is achieved.

Measuring RMS voltage

Following a number of circuit simulations, as described in my last article,[3] changing the value of the two input resistors and L_1 provided an excellent input VSWR for a 75Ω system, the 650mV input being slightly attenuated to provide 620mV maximum into the AD8361 IC. Adjusting the resistor values in the output attenuator to the DVM easily corrected the output voltage, Fig. 2.

Since only component value changes were needed, the same PCB as used for my 50Ω input design would provide a 75Ω version for this new high input-impedance meter.

I assembled a PCB using these revised components, fitted it with temporary cables and connectors and tested it, Fig. 3. Using a 75Ω generator, the circuit performed slightly better at both high and low frequencies, than it did when arranged for a 50Ω system.

Probe design

Satisfied that this 75Ω input version of my RMS meter circuit would perform when supplied from a high input impedance 75Ω output probe, I could now finalise my design.

The MAX4005 is a single-ended unity-gain buffer with a high impedance JFET input. It is intended to drive a 75Ω load so includes a built in precision thin-film output resistor. In a correctly-terminated coaxial

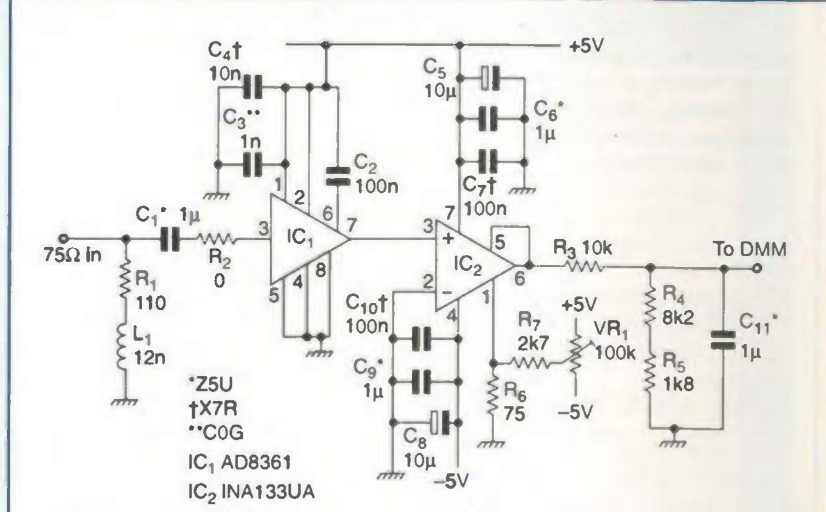


Fig. 2. The final schematic for my 75Ω true RMS meter printed board. The INA133 in-amp is used to correct no-signal offsets and as output buffer. This version needs a ±5V supply.

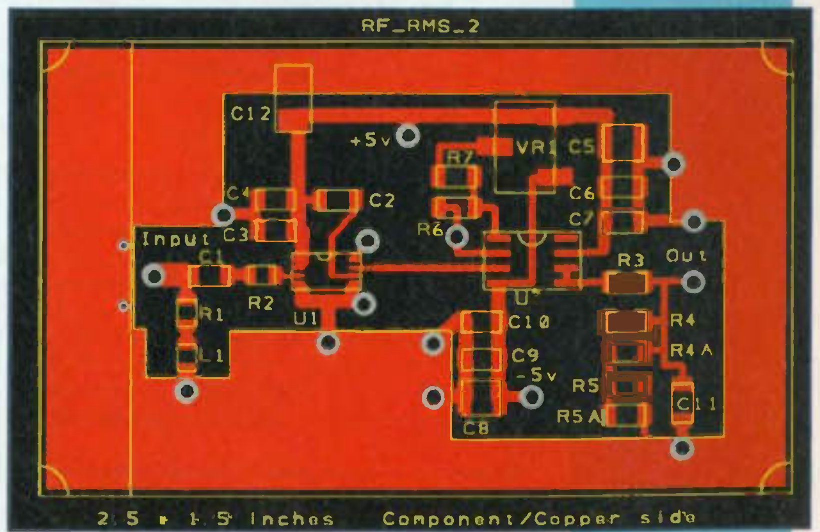
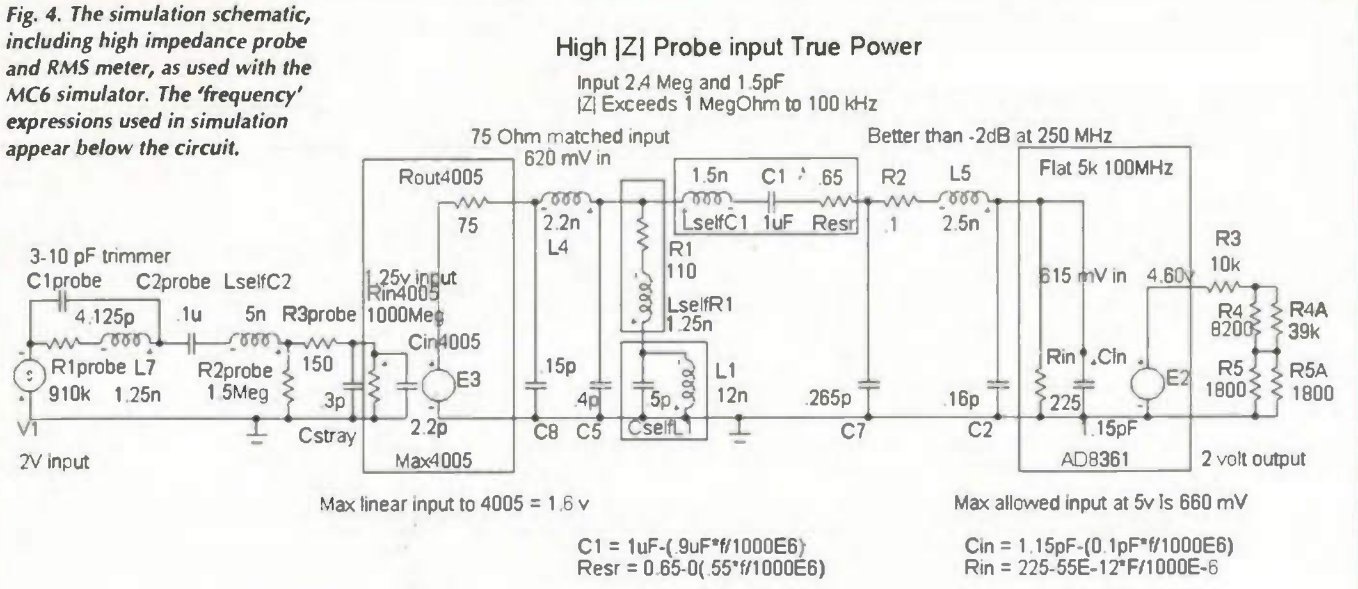


Fig. 3. Layout of the complete printed board, size 63.5 by 38mm, adapted for 75Ω RF RMS measurement. Accepting a maximum 650mV input, this board is able to accurately measure true RMS from 10kHz to 100MHz. Apart from the power and output pins, a full underside ground plane is used.

Fig. 4. The simulation schematic, including high impedance probe and RMS meter, as used with the MC6 simulator. The 'frequency' expressions used in simulation appear below the circuit.



cable, this resistor minimises reflections. Double termination then reduces the chip's signal gain to 0.5.

The device provides an extremely linear output, typically within 0.5% for input signals from –1.3 to +1.6V. When powered from ±5V, it can accept a maximum input of ±2.5V. With the above attenuation, the maximum safe input to this probe then becomes 7V pk-pk.

Gain flatness is within ± 0.1dB to 60MHz and ±0.2 dB to 80MHz. The amplifier's slew rate is 1000V/μs. With an input capacitance of 2.2pF and input current less than 1nA it is ideally suited to my front-end probe design.

This IC is only available in an SO8 surface-mount package and it has an unusual pin-out. To support its high slew rate, it provides two pins each for its positive and negative supplies. Furthermore pins four and five are connected to V_{CC} while pins one and eight connect to V_{EE}. It is important to remember this arrangement, otherwise it is only too easy to reverse connect the supplies when testing the probe.

Circuit design details

Unfortunately, Maxim[4] cannot supply a macromodel for its MAX4005. With the aid of the Microcap MC6 simulator, I used a voltage-dependent source together with two resistors and a capacitor to represent this IC in my simulations. Since the MAX4005's –3dB bandwidth extends to 950MHz and I needed operation only to 100MHz, this seemed to a provide a workable model.

A major design objective was to minimise probe input capacitance to ground. Ground planes and earth traces would be kept as far away as possible from the probe input and the IC.

Minimal stray capacitances to ground were modelled. With due allowance for inductance of circuit tracks and components, simulations to define the input attenuator circuit commenced.

I attempted to include a range switching relay to increase possible inputs to 20V. However, I found that the added inductances and capacitances needed to realistically represent the relay switch contacts complicated the design. As a result, I settled for a non-switched probe working to 2V AC, Fig. 4.

Ultimately a simple and small input attenuator was devised which in my simulations, performed exceptionally well. To minimise input capacitance though, I was forced to use an input trimmer capacitor rather than the more usual trimmer to ground. When set to 4.1pF this provided flat simulations to above 100MHz, Fig. 5.

Now it was time to etch and test a prototype probe.

Probe circuit board

I wanted to keep the probe size as small as practical for two reasons. These were ease of use when measuring in densely packaged circuit boards, and ease of housing in a suitable, small hand-held case. One housing possibility was a plastic tooth-brush case from Sainsburys, coated internally with screening spray.

The high slew-rate capability of the MAX4005 requires a low-impedance power supply. Since the power supply used would be at least a metre away from the probe, extensive on-board decoupling capacitance was essential. This occupied much of the final PCB design.

To minimise circuit strays and inductances, the probe input circuit had to be as compact as possible, requiring physically small components. Initially, the input trimmer presented a special problem. Most surface-mounted styles were simply much too large. In the end I settled on a small type CTZ2 from AVX, Farnell part No 578-370.

On tests, this trimmer, C_1 in Fig. 6, performed exceptionally well, but being rated to only withstand 100g pressure, it needed a delicate touch and an extremely small trimming tool to adjusting it.

A 38 by 23mm board was designed using 10μF tantalum, 100nF X7R and 1nF COG ceramic decoupling capacitors. Fitted with a temporary 75Ω load, a prototype board was assembled for tests, Fig. 7.

To ensure accurate measurements, an SMA stripline connector was fitted as input to the probe. This was then tested with both sine and square waves supplied from a

Measuring true RMS

The AD536, the first low cost IC to provide accurate RMS measurements, was designed by Barrie Gilbert more than twenty years ago. Using the 'implicit' solution of the RMS equation, feedback is used to perform the square-root calculation. This IC provided a 60dB dynamic range at low frequencies and a 1% accuracy with a 7:1 crest factor.

Now he has provided another unique RMS measuring IC, but this time based on the 'explicit' solution. The AD8361 provides extremely wide bandwidth, measuring from audio to 2.5GHz and up to 30dB dynamic range. This new technique can measure high crest factor waveforms, with an accuracy similar to a thermal bridge.

The AD8361 'TruPwr' detector is an RMS-to-DC converter that provides true RMS-responding measurements for complex waveforms with a nominal conversion gain of 7.5. It has a claimed dynamic range of 30dB and a ±0.25dB linear response up to 2.5GHz. It can measure from very low frequencies but not DC, up to 2.5GHz.

The device includes two identical squaring cells whose outputs are balanced by the action of a high-gain error amplifier. It calculates the conversion from RMS to DC automatically, by first squaring the input signal in the input cell. This input cell has a nominal low frequency input impedance of 225Ω. The input pin is biased to 0.8V so the input signal must be DC blocked by an external capacitor.

The current output from this squaring cell is averaged using an internal resistor and 27pF capacitor. At frequencies below 240MHz, additional external filter capacitance is needed. This averaged voltage is applied to one input of the error amplifier.

The second squaring cell is used to close a negative feedback loop around the error amplifier. This second cell is driven by a fraction of the quasi-DC output of the AD8361. When the voltage at the input of this second squaring cell is equal to the RMS value of the input signal, the loop is stable and the AD8361 output represents the RMS value of the input.

Scaling errors in both squaring cells cancel and they track with temperature resulting in stable measurements over the temperature range.

With a +5V supply, the AD8361 is linear with inputs up to 660mV, which with a nominal conversion gain of 7.5 results in a DC output of 4.95V. In practice, at 100MHz, this conversion gain is limited to between 6.5 and 8.5 times.

To allow for AD8361 offset errors, if the maximum possible linear input signal is used it is advisable to slightly increase the power supply voltage, by 100 or 200mV to avoid output compression, taking care not to exceed its maximum permitted supply of 5.5V.

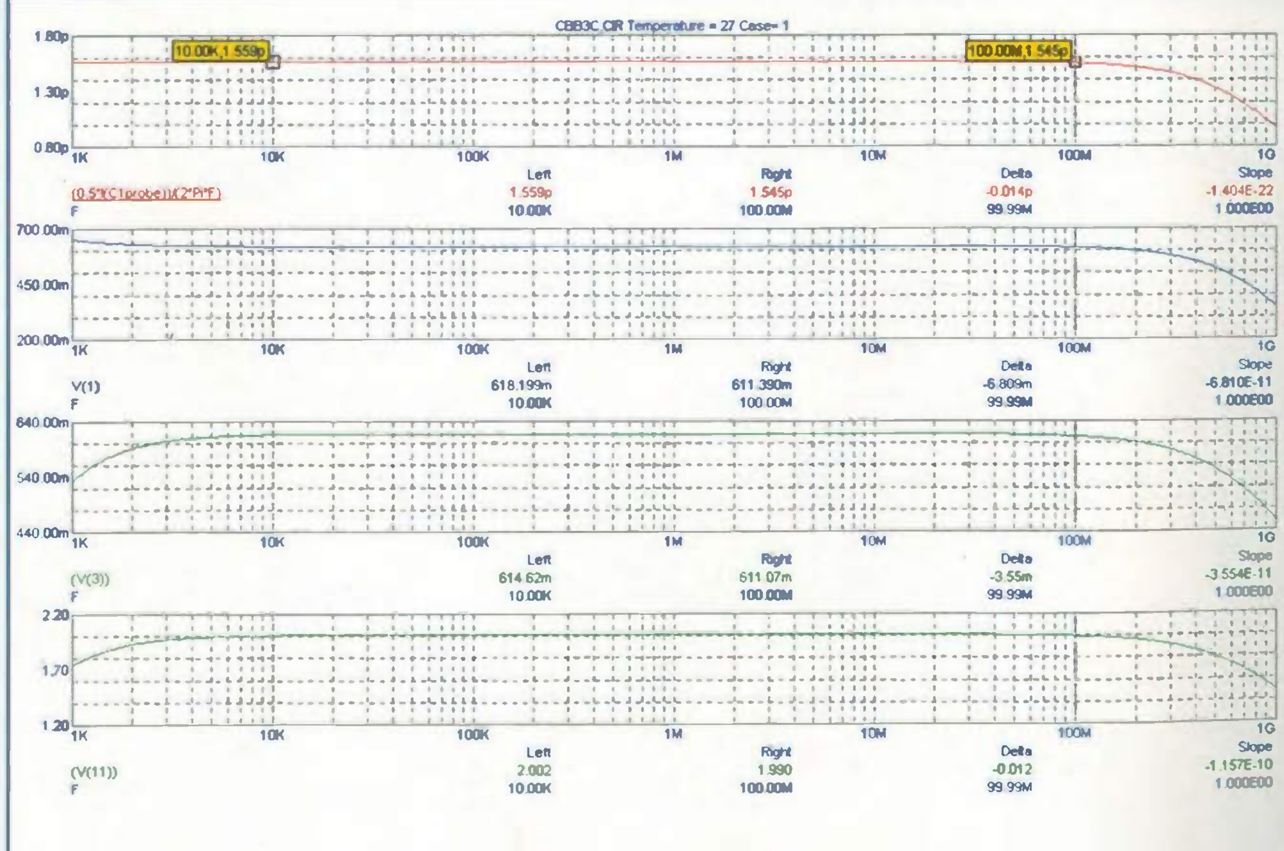


Fig. 5. Simulation results for a 2V input to the circuit of Fig. 4. Up to 100MHz, all simulation plots agree closely with the measured results. The top curve shows input capacitance, the other three the probe output, AD8361 input and the final attenuator output.

Calibrating the probe

The high impedance probe PCB, loaded with a temporary 75Ω terminating resistor, should be adjusted as a stand-alone probe before you connect it to the true-RMS meter PCB.

This is necessary because the input VSWR of the 75Ω RMS board rises slightly due to the impedance of its 1μF capacitor as frequency is reduced below 5kHz. The base reference voltage where the trimmer capacitor setting has little or no effect must be taken at 1kHz or lower frequency. The probe's uncorrected flat frequency response extends down to 100Hz.

Two methods can be used to determine the correct setting for the trimmer capacitor. The best method is to use a precise known sine wave at two frequencies, 1kHz and 1MHz. However this is only possible if you have a good AC voltmeter available to set and measure the probe input and output voltages.

Attach a short length of 75Ω coaxial cable to the high impedance probe output then to the voltmeter high impedance input via a 75Ω through termination located on the meter's input socket. Note the probe's output voltage at 1kHz then using exactly the same input voltage, adjust C_{1probe} to obtain this reading at 1MHz.

A practical alternative, is to use a good 1kHz square wave, available as the calibrator on your oscilloscope front panel.

Attach a short length of 75Ω coax to the output of the high impedance probe then to the oscilloscope input via a 75Ω through termination located on the oscilloscope input socket. Apply the high-impedance probe test prod to this calibrator and while monitoring the high-impedance probe's output on your oscilloscope, adjust C_{1probe} for the best possible square wave.

Remove the temporary coaxial cable and reconnect the probe output to the RMS meter PCB using your permanent 75Ω coaxial cable cable.

Combined assembly

For accuracy, a 50Ω signal generator and 50Ω connecting cables as needed, should be terminated with a 50Ω through load. Voltage at the output of the 50Ω through-terminating load is monitored using a known voltmeter and adjusted to the required 500mV.

Attach the high-impedance probe input prod and the known voltmeter to this 50Ω through load using a 'T' connector.

Do not adjust C_{1probe}, which has already been set for the correct response.

Measure the DC voltage V_1 at the junction of U_2 and R_3, then adjust the attenuator value by adding resistors R_{4A} and R_{5A} as appropriate.

The required total value R_x, which is the net sum of R_4, R_{4A}, R_5, and R_{5A} can be easily calculated:

$$R_x = \frac{0.5 \times 10000}{V_1 - 0.5} \text{ in ohms.}$$

Typical net values for R_4 and R_5 for AD8361 at nominal and maximum gains are,

AD8361 gain=7.5	$R_4 \| \| R_{4A}$=6.75kΩ	$R_5 \| \| R_{5A}$=900Ω
Nearest std values	R_{4A}=39kΩ	R_{5A}=1.8kΩ
AD8361 gain=8.5	$R_4 \| \| R_{4A}$=5.98kΩ	$R_5 \| \| R_{5A}$=200Ω
Nearest std values	$R4_A$=22kΩ	R_{5A}=220Ω

Apply signal voltages of 0.25, 0.5 and 1.0 and verify attenuator output accuracy at these voltages.

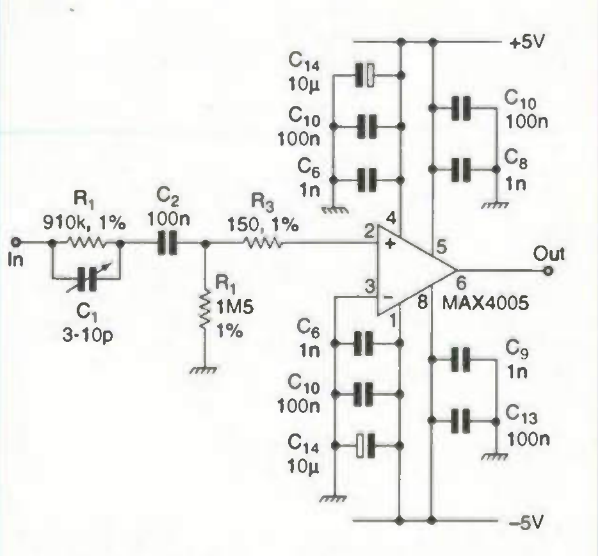


Fig. 6. The final probe's schematic as built and photographed. Note the MAX4005 uses four power pins in an unusual configuration. Pins 4 and 5 supply +5V, pins 1 and 8 supply –5V. Because the power supply is a metre distant, substantial on-board decoupling capacitors are needed.

correctly-terminated 50Ω system.

I monitored sine wave input and output voltages using my HP331A below 1MHz and my HP8405A vector voltmeter for higher frequencies. Square wave input and output voltages were monitored using a pair of carefully-matched and adjusted Coline M12SW oscilloscope probes, both set to divide by ten.

Using a 1kHz square wave, trimmer C_{1probe} was first adjusted for optimum square wave response by observing both the probe input and output voltages simultaneously on my dual-beam oscilloscope.

Increasing this square wave input to 5MHz revealed identical input and output probe traces, indicating this probe's response was at least as good as my matched pair of Coline M12SW scope probes at this frequency.

Sine wave measurements from 100Hz to 100MHz using the HP331A and HP8405A instruments showed a near flat response to 100MHz for this probe on its own, when terminated in 75Ω, **Fig. 8**.

Coaxial input prod

In order to measure in-circuit voltages, I wanted my probe to accept the prods and adaptors provided for my Coline M12SW probes. The Coline probe body terminates in a 4.8mm diameter tube and 0.7mm probe tip.

This 4.8mm tube was no problem. I already possessed some lengths of thin wall 4mm ID brass tubing obtained from a local aeromodelling supplier. Incidentally 'Paper Mate' ball pen refills are also almost exactly the required external diameter.

My wife's sewing box provided two needles of suitable length and diameter. Together with two PTFE insulators from insulating stand-offs, I had the required parts.

To accept the Coline spring hook adapters, I needed a prod body at least 30mm long. A few calculations suggested that a coaxial prod of this length, assembled using these parts, would exhibit some 1.65pF of capacitance.

In the event, to ensure a sufficiently tight fit of the PTFE into my tube I had to insert a 5mm length of 1.4mm outside-diameter tubing into the central hole of each PTFE insulator. This would slightly increase the calculated capacitance.

When assembled, my finished test prod measured 1.9pF. Added to the probe PCB capacitance of 1.5pF this would result in a total probe input capacitance of some 4pF.

Final tests

This prod was attached to my prototype probe PCB and the above sine wave and square wave tests repeated. I was satisfied with the results. To facilitate a secure test prod mounting, I increased the PCB length to 44.5mm. I then assembled and calibrated the final probe. **Fig. 9**.

The final probe was attached to my 75Ω true-RMS measuring PCB using RG179 B/U coaxial cable, ready for full tests of the completed assembly. This PTFE insulated coaxial cable, which is available in cut lengths, is easily soldered directly to a PCB without damaging the cable.

Testing using 500mV sine waves, I found performance from 5kHz to 100MHz was better than ±0.25dB. At 200MHz it was nearly –0.5dB and better than –3.0dB to 300MHz.

Using a 500mV reference, dynamic range measured at 1MHz was better than 0.25dB to 1.9V. Measured at 10MHz with a 500mV reference, it was –0.28dB at 250mV, better than –2.0dB to 100mV and better than –4dB at 50mV.

The particular AD8361 IC used exhibited a no-signal offset of 110mV. This was corrected to 3.4mV using an INA133 instrumentation op-amp. I would expect to

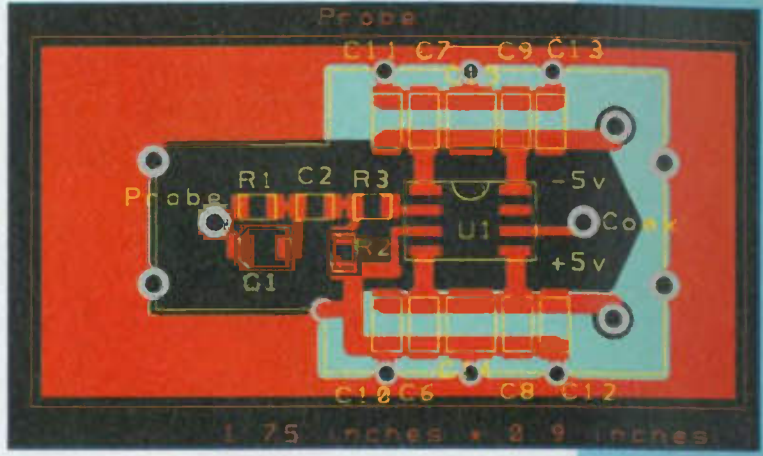
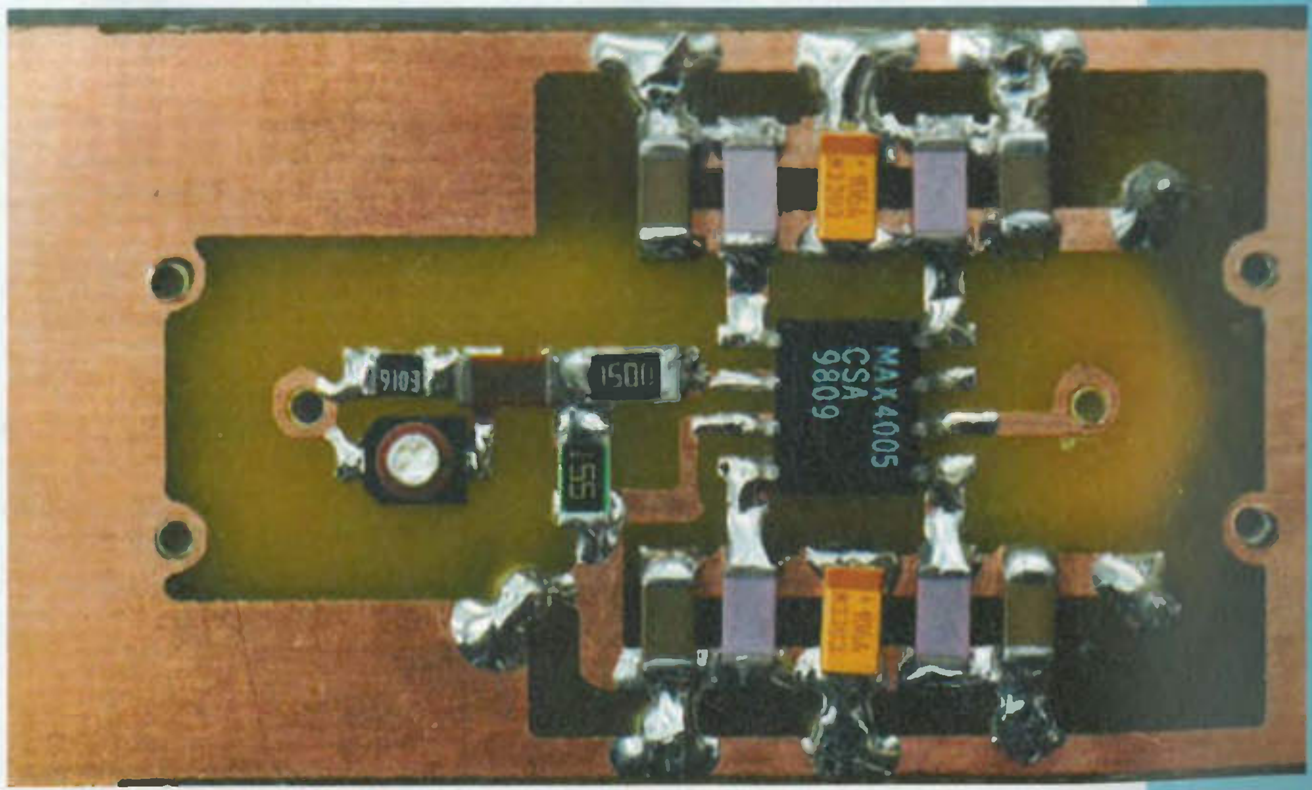


Fig. 7. The final double-sided probe PCB as built and tested using upper and lower ground planes. To facilitate secure mounting of the coaxial test prod, board length has been increased from 38 to 44.5mm. To minimise capacitance, both ground planes are distanced from the high-frequency paths.

Fig. 8. Assembled and pre-tested probe PCB, awaits fitting of the coaxial test prod and connection to the RMS meter assembly and final tests. The small 3-10pF trimmer capacitor, C_{1probe}, appears immediately below the left most, 910kΩ, 0805 size input resistor. The underside ground plane is visible as the darker area.



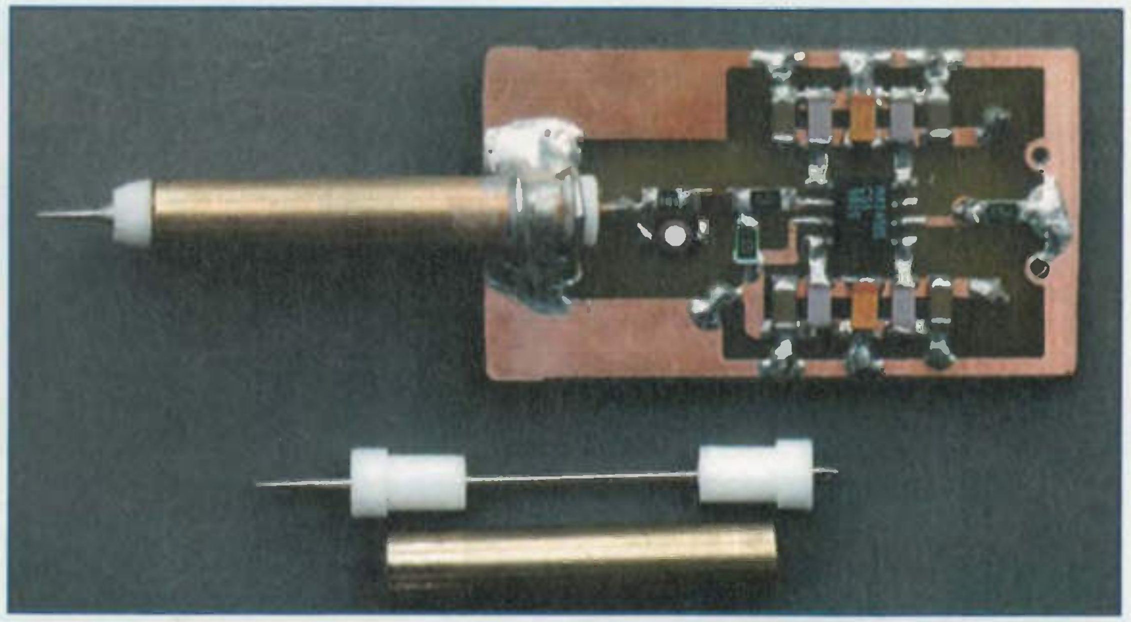


Fig. 9. Exploded view showing component parts used to assemble a low capacitance coaxial test prod, able to accept adaptors from my 250MHz Coline oscilloscope probes.

attain this dynamic range or better with almost all samples of the AD8361.

Altogether this probe and revised RMS board has produced a most satisfactory and low cost high performance RMS responding in-circuit volt meter. Its

Table 1

Voltage error by frequency for probe and 75Ω RMS meter.

Frequency (Hz)	DC out (mV)	Relative (dB)
1k	386	–2.25
5k	486	–0.25
10k	496	–0.07
50k	497	–0.05
100k	501	+0.01
500k ref.	500	0.0
1M	499	–0.01
5M	501	+0.02
10M	500	0.0
25M	500	0.0
50M	504	+0.12
75M	509	+0.15
100M	512	–0.206
150M	506	+0.104
200M	473	–0.48
250M	428	–1.35
300M	380	–2.38
350M	335	–3.48
400M	291	–4.7

Note: 0dB input was set to 500mV

input impedance is high and its capacitance very low.

For the amplitude results, I used an HP331A voltmeter up to 1MHz and an HP8405A vector voltmeter for higher frequencies.

For the above tests, the AD8361 no-signal output offset of 110.5mV was reset to +3.4mV by adjusting the

Dynamic range performance of the probe

Input at 1MHz	Output DC	Error (dB)
1.9V	1.937V	+0.167
1.5V	1.521V	+0.12
1V	1.003V	+0.026
Input at 10MHz		
1V	1.003V	+0.026
750mV	748mV	–0.023
500mV reference	500mV	0.0
250mV	242mV	–0.28
100mV	81mV	–1.83
50mV	32mV	–3.88
25 mV	13.7mV	–5.22

INA133 reference voltage.

The 10MHz input amplitude for each measurement was monitored using an HP8405A vector voltmeter, while at 1MHz an HP331A voltmeter was used.

My next article will detail how this high-impedance, low-capacitance probe design can be used to provide a unity-gain, high-impedance, low-capacitance, oscilloscope probe, usable from 100Hz to 100MHz. ■

References

1. Bateman, C., 'Measure AC millivolts to 5MHz', *Electronics World*, p. 281, April 2000.
2. RMS to DC converters ease measurement tasks. http://www.analog.com
3. Bateman, C., 'Measuring RF millivolts', *Electronics World*, p. 592, August 2001.
4. 950 MHz FET-Input Buffer with 75Ω output, http://www.maxim-ic.com
5. 'Revolutionary RF IC Performs RMS-DC Conversion', *Microwaves and RF*, September 1999.

CIRCUITIDEAS

Fact: most circuit ideas sent to *Electronics World* get published

The best circuit ideas are ones that save time or money, or stimulate the thought process. This includes the odd solution looking for a problem – provided it has a degree of ingenuity.

Your submissions are judged mainly on their originality and usefulness. Interesting modifications to existing circuits are strong contenders too – provided that you clearly acknowledge the circuit you have modified. Never send us anything that you believe has been published before though.

Don't forget to say why you think your idea is worthy.

Clear hand-written notes on paper are a minimum requirement: disks with separate drawing and text files in a popular form are best – but please label the disk clearly.

Send your ideas to: Jackie Lowe, Cumulus Business Media, Anne Boleyn House, 9-13 Ewell Road, Cheam, Surrey SM3 8BZ

Increasing the signal level handling of a digital potentiometer

While building a stereo amplifier I decided to make use of two DS1804 digital potentiometers from Dallas Semiconductor. The idea was to prevent the routeing of low-level analogue signals from the rear phono connector panel to the volume control on the front of the box, so minimising the noise that could be picked up *en route* to the amplifier inputs.

The voltage restriction on the potentiometer means that the signals on any of the analogue terminals H, L, or W cannot go more than 0.5V below ground (pin 4) or 0.5V above the supply (pin 8).

Even after dividing down the line-level signals from my CD and tuner, I experienced a problem characterised by a low-level buzzing coming through the speakers with the wiper in position 0, i.e. at ground potential, in response to peaks in the music level. Rather than try to DC bias the analogue terminals above ground and run the risk of signal degradation, I used two diodes to drop the supply and create an artificial ground for the DS1804, giving an extra 0.6V or so of signal handling (0V and the analogue ground on L are tied together at the power supply).

To prevent overdriving of the INC and UP/DN inputs, I used two more diodes. The resistors are needed on these lines to ensure the device sees correct logic-low levels.

Encoder output is debounced by a 74HC123 dual monostable, set to around 0.5s pulse width.

An extra feature of the circuit shown is the driving of the CS input, pin 7. When the AC supply is active, this line is pulled low by the BC337 and the potentiometer reacts to the states of the control lines.

When the unit is powered down, CS goes high because the 0.47μF capacitor discharges before the 5V supply drops. The current wiper position is transferred to a non-volatile store on-chip. This is automatically retrieved at next power-up, so restoring the last setting of the volume.

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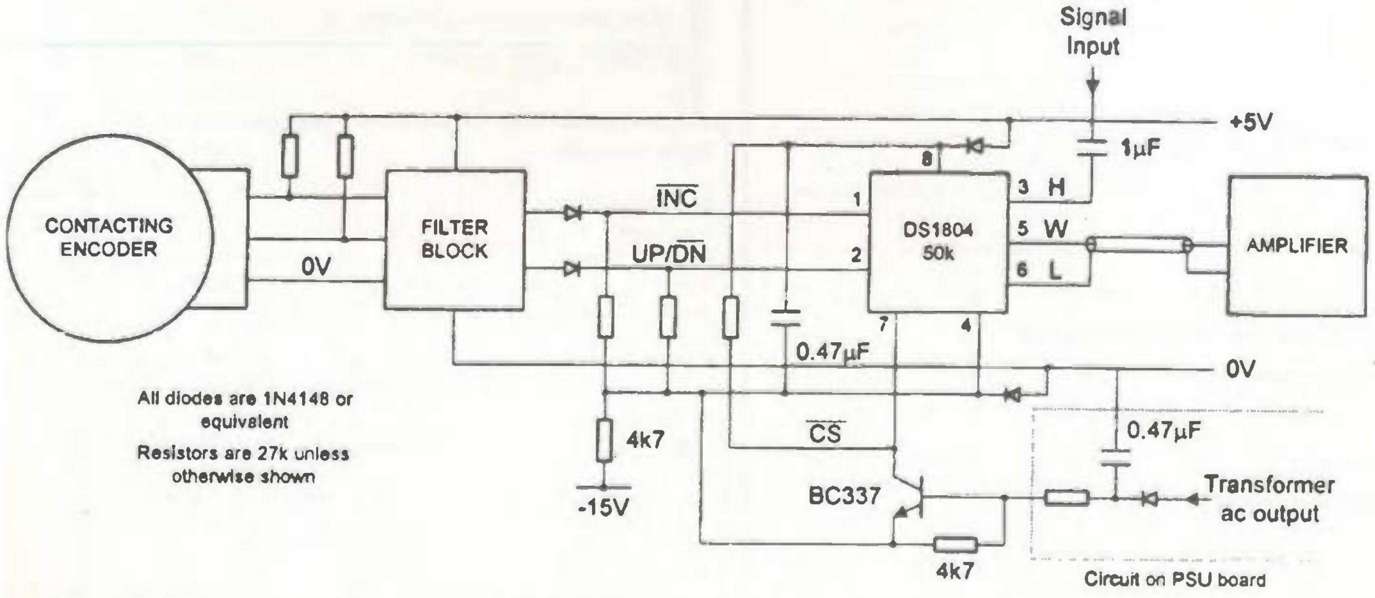

Enhanced digital volume control handles audio. For the other channel, simply connect pins 1, 2, 4, 7 and 8 to the same lines, feed the signal via another capacitor to pin 3, and connect pins 5 and 6 to a second amplifier.

Stepper motor sequencer

There is no shortage of imaginative circuitry or dedicated stepper motor drivers, but for basic 4 pole, unipolar 2 phase excitation, a 4018 is perhaps as cheap and easy as it gets. A glance at a data book index would not suggest suitability but it is a slightly unusual counter, as the circuit configuration timing diagram illustrates.

Although not used in this simple circuit, initial starting condition can be set up with the jam inputs. For operational characteristics the reset has been show in the timing diagram, with all the jams low.

Outputs can be gated for reversal but in such applications a dedicated driver might be more appropriate.

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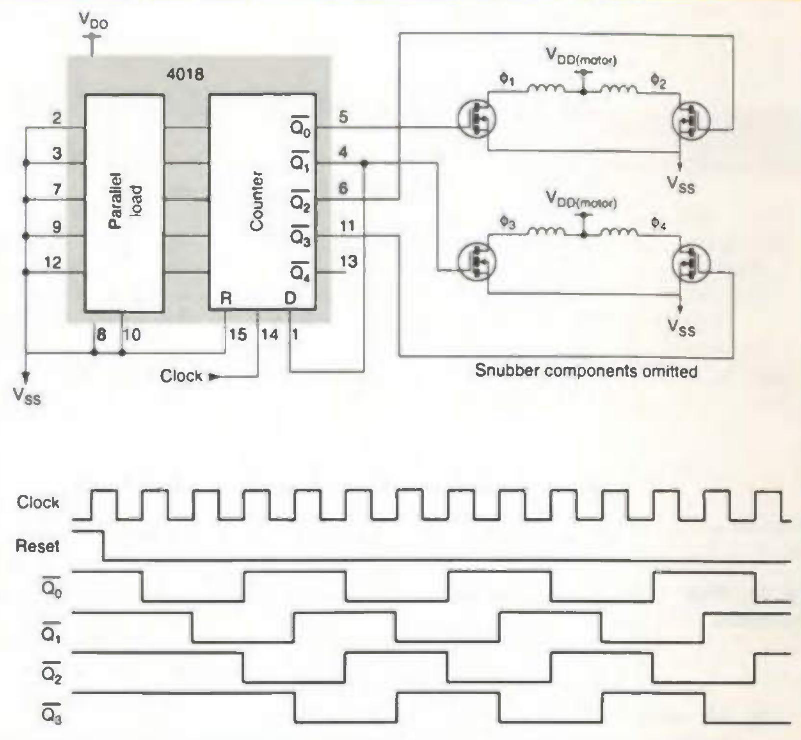

Attenuators using common resistors

During testing of sensitive amplifiers, it is sometimes useful to insert an extra 20 or 40dB attenuation of the signal coming from the source. In case you do not have a handy ready-made coaxial attenuator, the pads shown in **Fig. 1** can be rapidly constructed and fitted in a tin box.

For a 50Ω system, with no added attenuation, the source and load are both 50Ω, and hence there is an automatic drop of 50% of the source EMF at the load. The power in the load is one half or –3dB relative to the total power supplied by the source.

For an additional loss of 20dB, the values in Fig. **1a)** preserve an accurate 50Ω impedance on the input and output sides, and give the correct additional attenuation: thus there is an overall loss of ×20 in source EMF.

Alternatively, the end resistors can be replaced with single 62Ω, and the link a single 240Ω. The impedance is then 0.7% high, but the attenuation error is rather larger: the signal ends up almost 4% high. Similarly, the 40dB pad of Fig. **1b)** provides accurate 50Ω impedances, and an overall loss of ×200 in source EMF. With the values shown, the errors are less than 0.05%, though tolerances of common resistors will tend to swamp this.

Bear in mind that the front-end resistor may have to dissipate appreciable power. I use Philips MRS25 type with 0.6W rating, allowing 5.5V rms across 51Ω.

At high frequencies, stray capacitance and aerial coupling from input to output become a limiting factor: e.g. 1pF at 1MHz contributes a shunting effect of –j160kΩ. Hence trying to make a 60dB pad all in one go with a 15kΩ+10kΩ cross-link is not very successful except at moderate frequencies.

A rapid form of construction is to splice the components into a 50Ω BNC-BNC cable, or use a BNC plug-cum-cable plus a BNC socket. With a little care and insulating tape, a fairly well screened result can be achieved by adding strips of braid all around.

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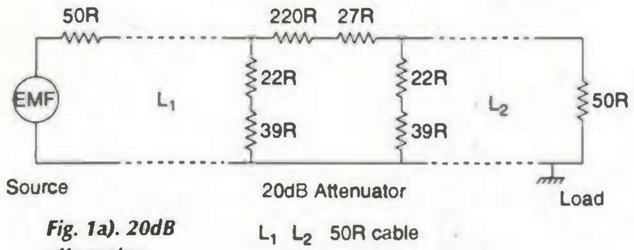

Fig. 1a). 20dB attenuator.

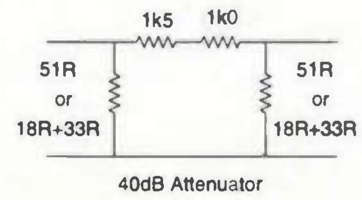

Fig. 1b). 40dB attenuator.

Cascode driver improves SMPS

Using a cascode power stage for switch mode power supplies operating from high supply voltages – direct from the mains for example – offers potential advantages.

Faster switching is achieved. This is because the limited current capability of the controller chip has to drive the Miller capacitance of the lower MOSFET transistor (see circuit diagram) through a collector voltage swing of only about 15V rather than the 400V of a single-transistor circuit. This reduces switching losses.

Additionally, the Miller capacitance of a 50V-rated transistor will generally be less than that of a similarly rated 500V device. A clamp circuit is necessary in order to minimise the turn-off drain voltage swing at this point and this may conveniently be arranged to provide power for the controller chip.

Feeding only the gate bias circuit of the upper transistor of the cascode pair from the start-up resistors gives a fast turn-on during power up, even though a relatively large capacitor is provided to maintain supply to the control chip between supply cycles. Also, there is the capability to supply auxiliary circuits like slope compensation amplifiers with a moderate parasitic drain.

The start-up resistors have low power ratings so they may be small and inexpensive. Note that two resistors are used, not for dissipation reasons, but safety. Small resistors are not rated to stand off mains – be sure the resistors used have an adequate voltage rating.

The circuit operates as follows. On start-up R_1+R_2 pull up C_2 and Tr_2 acts as a voltage follower to provide the few milliamps required to charge C_4+C_1 to the start up volts of the controller chip. Most controller chips of this type incorporate an internal zener function. The current into this chip is limited by Tr_1, which clamps the gate volts of Tr_2 if the current when it detects 0.7V or so across R_3.

In operation Tr_3 is pulsed on as required by the controller chip. In addition to driving the output via Tr_2, when on, it clamps the gate volts of Tr_2 to about 10V via D_4 and D_2. When it turns off, the rise in volts on the drain of Tr_2 charges C_3 via the drain/source capacitance of Tr_2 and D_3. This is a fast diode with a high transient current rating but only about 20V reverse rating. This charge supplies operating current to the controller chip via R_6,R_7 which keep fast high currents out of the controller.

If Tr_2 has a large source/drain capacitance and the design features a high switching rate it may be necessary to limit the current flowing into the controller chip by adding a zener of say 15V across C_4 and increasing R_5 to keep the current within the chip makers specification.

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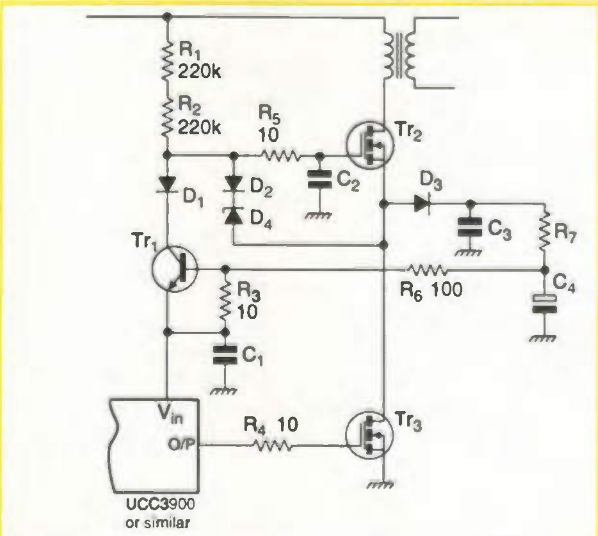
Circuit Ideas editor's note: this circuit is suggested by its author, but has not been built and tested.

Cascode connection used in a switch mode power supply.



Valve-based, time-aligned, active loudspeaker crossover filter

£75 winner

A recent article[1] described a perfectly time-aligned active filter system for audio. The main characteristic of the system was that any input signal can be faithfully reproduced at the output.

The circuit described here is based on the same principle but implemented using valves, which have enjoyed resurgence in recent years. The complete circuit – excluding the power supply – is shown in **Fig. 1**.

No global feedback is employed. The buffered input signal is fed to a high-pass filter, which is a simple second-order *RC-RC* filter. The low-pass filter is built from a difference amplifier with inputs from the buffered input signal and the output from the high-pass filter.

Note that other high-pass filters may be used without adjustment. A volume-pot is fitted within the low-pass section to cater for any difference in amplifier and/or speaker sensitivity.

The 1kΩ variable resistor V_R is used to balance any mismatch between the two halves of the 12AU7. Gain of the low-pass section is approximately four. Both simulation and experimental results agree with the description given in the reference.

A typical simulation output is shown in **Fig. 2**.

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Reference

1. 'Adaptable active speaker system', *Electronics World*, Feb. 2000.

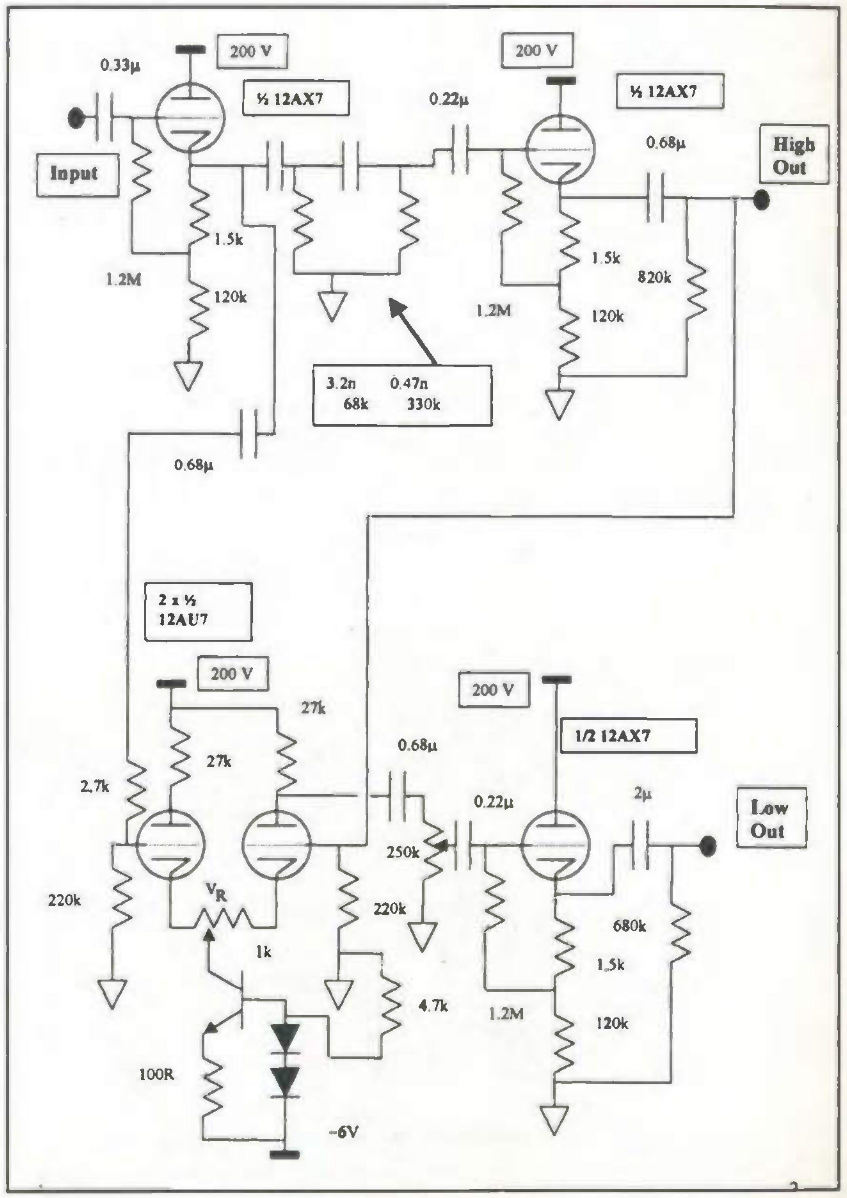


Fig. 1. Valve-based active crossover filter with time alignment.

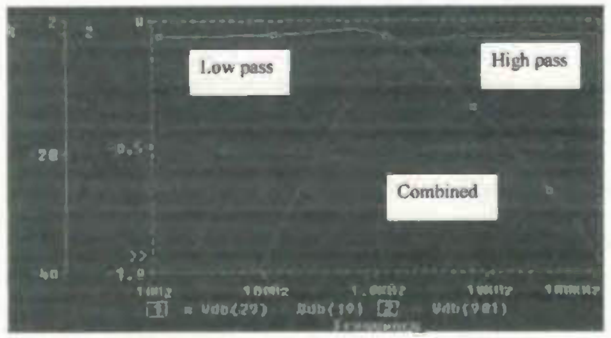


Fig. 2a) Frequency response of the filter. The high-pass section is –3dB at 1.5kHz, –6dB at 0.96kHz. The low-pass side is +1.5dB at 0.6kHz and –3dB at 2.2kHz.

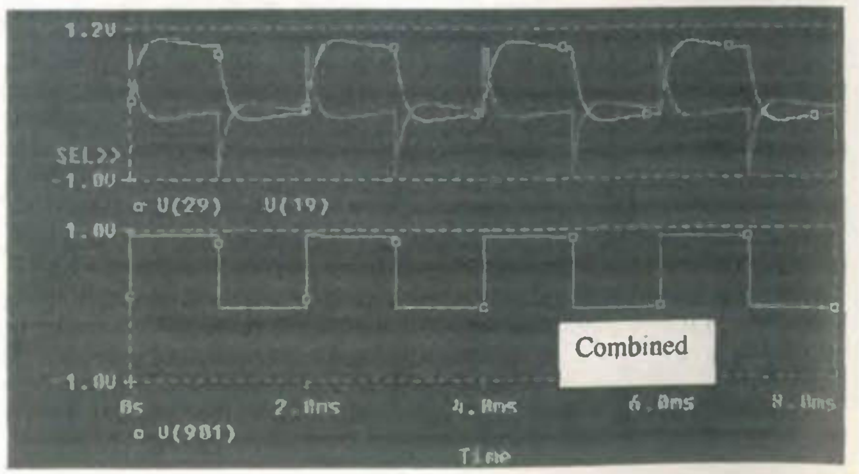
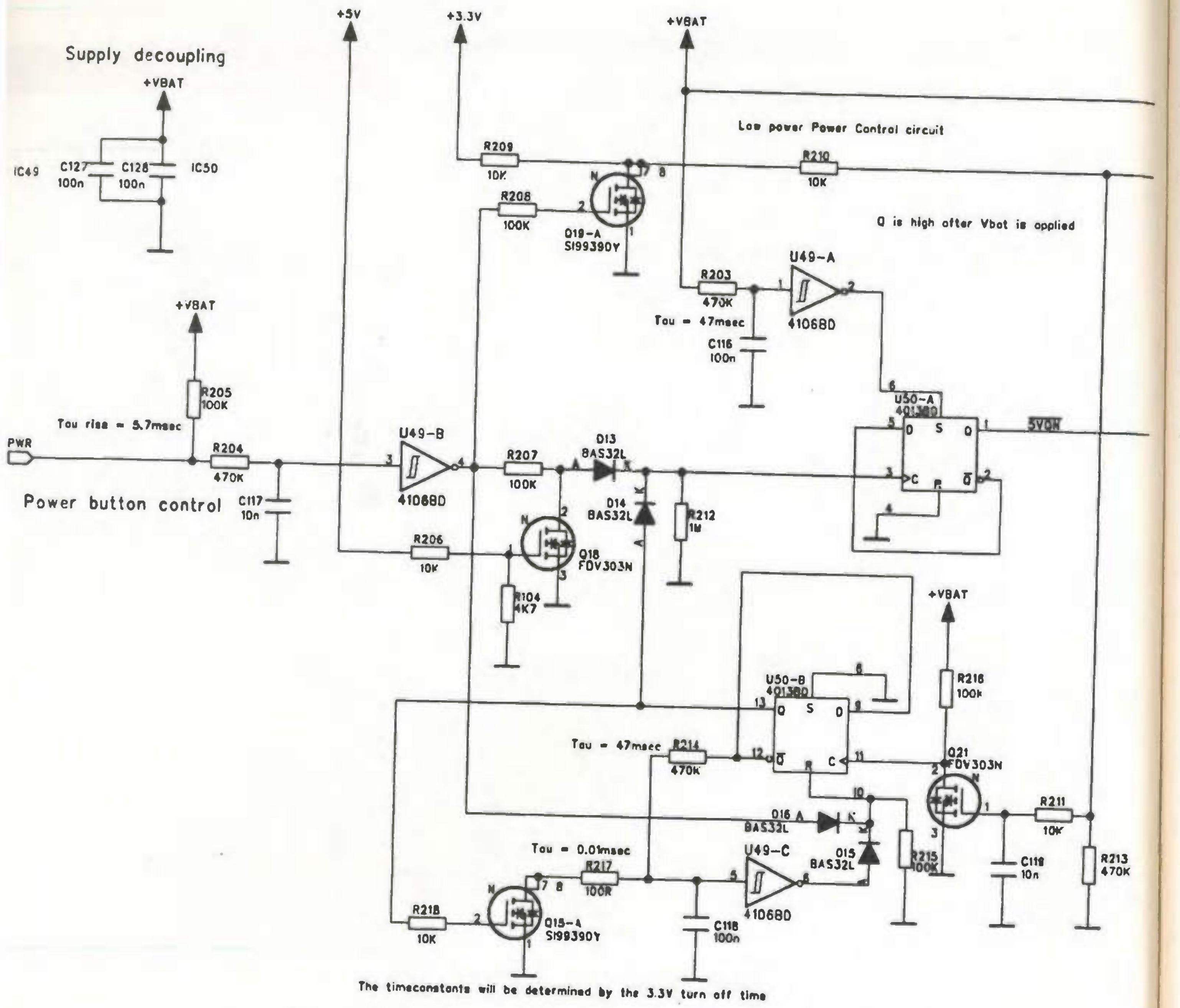


Fig. 2b) Rectangular signal response.



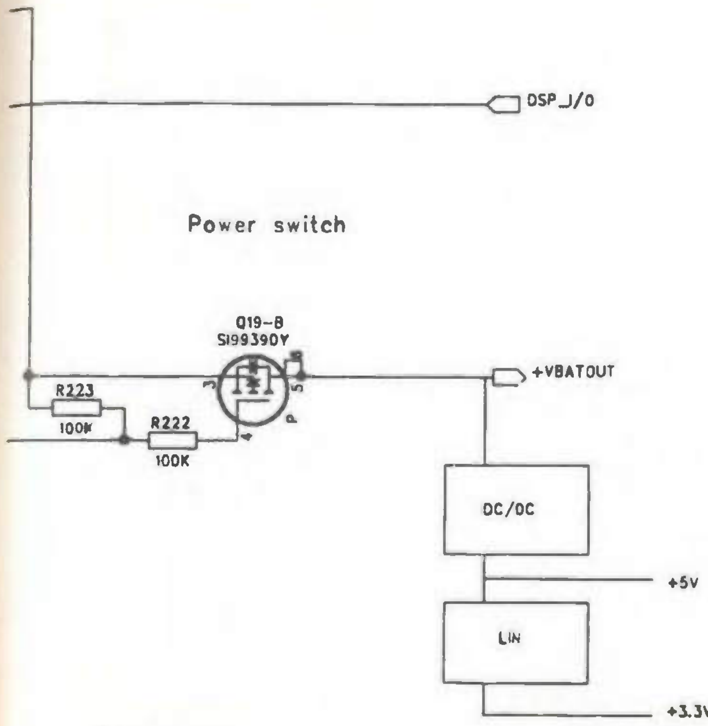
Intelligent power switching

This circuit lets the power for an instrument be controlled by a push button. The operator simply turns the instrument on by pressing the power button on the keyboard. When the instrument is on and the user again presses the power button, the host DSP system or controller has the control of the circuit. This ensures that whatever the DSP is doing can be ended and that the DSP will turn the instrument off.

Power button control. The functions of this circuit are explained under the following conditions:

1. There is no battery connected to the instrument
2. The battery is connected to the instrument
3. The power button is pressed the first time
4. The power button is pressed the second time.

1. When no battery is connected all signals are at zero volt, i.e. low.
2. When the battery is connected, it is essential to have the two D-type flip-flops set into correct states from start. Flip-flop U50-A is set by U49-A because of the *RC* circuit at its input. This ensures that its output and S_1 is high for around 50ms before it goes low. When U50-A is set, the Q output of U50-A is high and the /Q output of U50-A is low. As a result, U50-A's D input is also low. This puts transistor Q_{19}-B into its off state.
The second flip-flop, U50-B, is reset by U49-B through diode D_{16}. This ensures that its R input is high for approximately 5ms. Once it is reset Q of U50-B is low and /Q is high, leaving U50-B's D input high also.
Now both flip flops are in known states. The signal DSP_I/O, which is an I/O signal from the DSP, is kept low during this process through R_{213} since the +3.3V is off.
3. When the power button is activated the first time the signal PWR will be shorted to GND while the switch is closed. This causes the signal at the output of U49-B to go high for that period. Then U50-A is clocked at its C input. This causes the outputs of U50-A to invert so transistor Q19-B turns on, resulting in the +5V supply going on. This +5V supply is connected at +$VBAT_{OUT}$. The +3.3V supply also turns on. When the +5V rail has come up, then Q_{18} turns on, ensuring that the signal path from U49-B to the C input of U50-A can not be altered by the user. This prevents the user from activating U50-A again.
The I/O line PB05 on the DSP is automatically configured as an input when the device starts up, after the +3.3V is running. As long as the user presses the power button, this signal will be low since Q19-A is on. The DSP will not continue its program before the user has released the power button and PB05 goes high. Though U50-B will be reset once again, this will not alter its output. The system is running.
4. When the power button is pressed again, PB05 goes low again resulting in a GPIO interrupt. This eventually causes the DSP to initiate a shutdown. This activation will of course also toggle the C input of U50-B, but its output will not change. The user releases the power button and now the DSP can start the shutdown process. First the DSP has to ensure that any vital processes are ended and then store any critical data in nonvolatile memory. Then it programs PBO5 as output and sets it low. This causes the C input of the U50-B to invert the output, which both will toggle the C input of U50-A and invert its output, shutting down the system. But at the same time Q15-A and U49-C will cause a pulse to reset U50-B again, leaving the circuit in the same state as before the user pressed the power button the first time.



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Logic level selects direct or inverse-acting power control curve

On/off, proportional, PID, etc., controllers are used either in direct acting or inverse-acting modes. The best example of the inverse-acting mode is a heating process, while the refrigeration process is an example of a direct-acting mode of control.

Almost all the digital controllers can be configured to act as either direct-acting or inverse-acting. But in the case of analogue controllers you have to change the hardware to change their configuration from direct acting to inverse acting or *vice versa*.

Therefore analogue controllers designed for a heating process (or refrigeration process) cannot be used for refrigeration (or heating) without a change in the hardware.

This circuit, **Fig. 1**, has been designed and developed to be programmed via a logic level to function either as direct or inverse acting. It can be connected in between the controller and the power unit, after fixing the voltages V_1 and V_2, which define the maximum and minimum voltage values of the controller output and the power unit input.

Figure 2 shows the variation of the output voltage with 0 and 1 logic levels in response to the input voltage V_{in}. Mathematically this can be put as:

$V_{out}=V_{in}$ for logic 1 (switch S open)
$V_{out}=V_1+V_2-V_{in}$ for logic 0 (switch S closed)

We chose V_1 and V_2 as 10V and 0V respectively for a controller whose output varies from 0V to 10V for 0% to 100% power. You can choose the range as 0-10V, 0-5V, or any other range by selecting V_2 and V_1.

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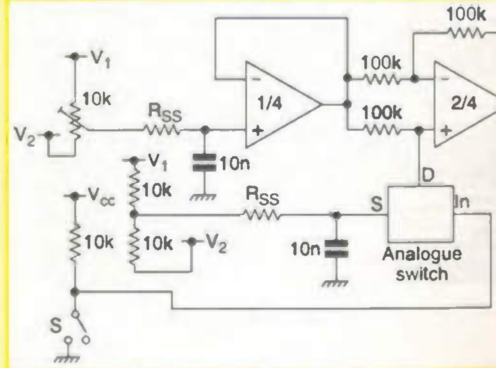


Fig. 1. S_1 determines if V_{out} moves in the same sense as V_{in}, or the reverse.

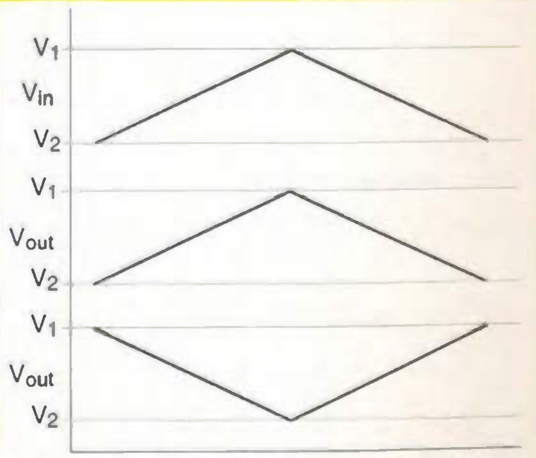


Fig. 2. Illustrating the circuit's action. Top is the input, next is the output with logic 1 applied and bottom the output with logic 0.

Inverting current conveyor from an OPA1662

An OPA2662 operational transconductance amplifier can be used as a non-inverting current conveyor, or CCII(+), without any modification. In this case, $Y=B$, $X=E$ and $Z=C$, where *XYZ* denote terminals of the current conveyor and *BEC* denote pins of the amplifier, **Fig. 1a)**.

In many applications though, an inverting current conveyor, or CCII(–), is also required. To construct one, we used both parts of an OPA2662 dual integrated circuit, Fig. **1b)**.

Relationships between currents I_x and I_z and voltage U_y where R_x and R_z were 100Ω are shown in **Fig. 2a)** for CCII(+) and in Fig. **2b)** for CCII(–). The following coefficients can be determined from these curves:

$I_z=3.1I_x$ for CCII(+) and
$I_z=2.9L_x$ for CCII(–).

Both current conveyors were used in the circuit realising a frequency-dependent negative conductance, as in **Fig. 3**. This FDNC circuit was described in *Electronics World*, May 2001. The negative conductance $-\varpi^2D$ appearing at the input terminals follows,

$$D=\frac{C_1C_2R}{k_1k_2}$$

Here, k_1=3.1 and k_2= 2.9. In **Fig. 4**, the points represent measurements and continuous lines show data determined from $-\varpi^2D$.

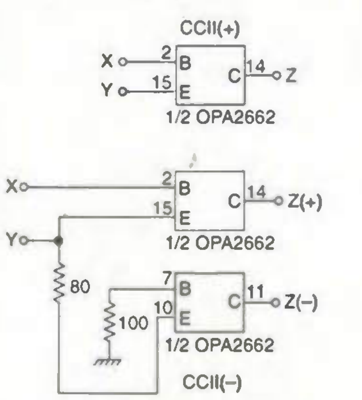


Fig. 1. Current conveyors, with the CCII(+) in diagram a), and the CCII(–) in diagram b).

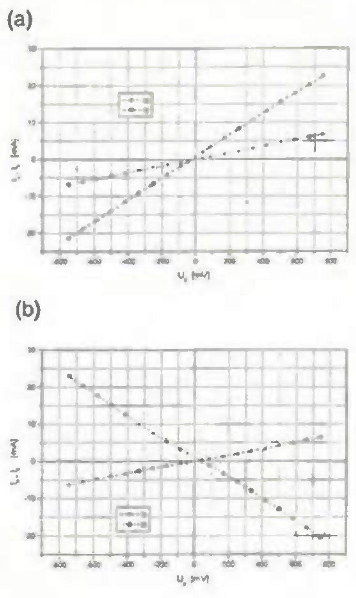


Fig. 2. Currents I_x and I_z versus voltage U_y for CCII(+) and CCII(–) in plots a) and b) respectively.

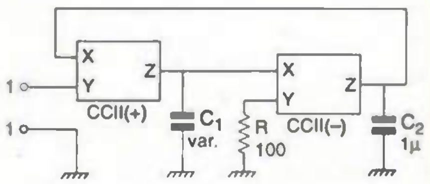
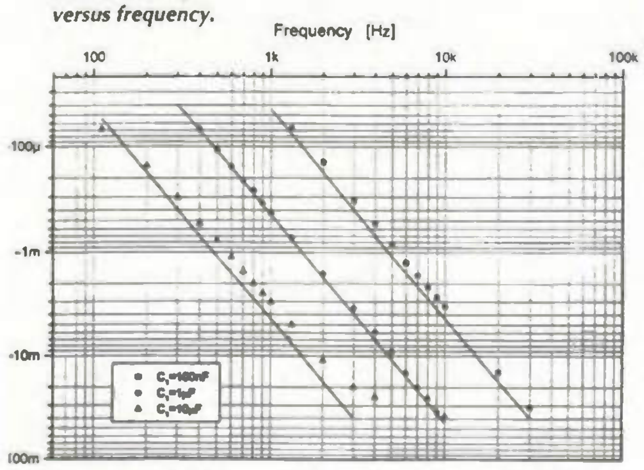


Fig. 3. Circuit realising a frequency-dependent negative conductance, where $Y(s)=s^2D$ on terminals 1-1.

Fig. 4. Negative conductance of the circuit from Fig. 2 versus frequency.



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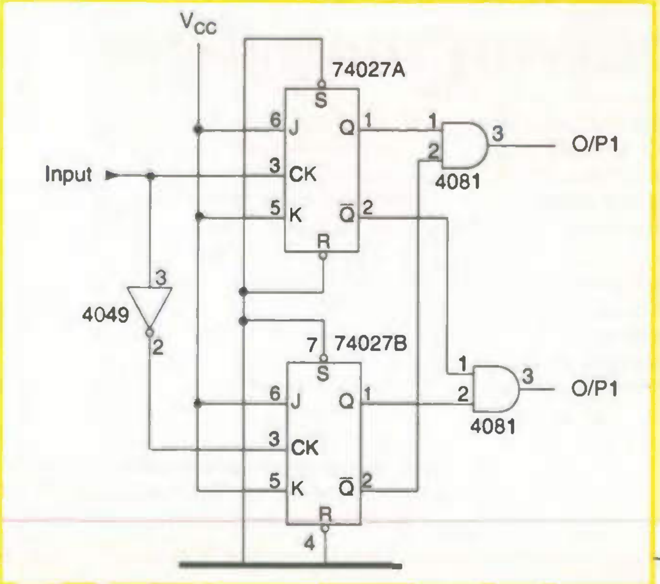
Reference

1. L.Tomawski, M. Mafika, M. Slawiec, 'FDNCs using current conveyors,' *Electronics World*, May 2000, p. 380.



Alternate pulse router

Originally devised to facilitate the calibration of mechanical speed traps, this simple circuit has proved useful in several other applications such as in driving pulse transformers.

The circuit routes alternate pulses to a second output.

The 4049 inverter at the input can be replaced with another type of inverter, for example, a transistor circuit.

Similarly, other devices can replace the 4027 bistable device and the 4081.

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Simple circuit for measuring capacitor ESR or low resistance

An unconventional but simple circuit using a 555 timer IC and five other components, plus a +5V DC supply and a frequency counter, enable approximate comparison/measurement of capacitor ESR values. Such low values cannot be measured by most capacitance bridges or meters.

The circuit consists of a CMOS version 555 astable oscillator with 50:50 duty cycle output. It is modified by the addition of R_2 and r. Small variations in the value of r – for example the ESR of the timing capacitor – provide a large change in output frequency.

Output frequency of the oscillator is given by:

$$f=\frac{1}{2CR\ln(2-3k)}$$

Here, C is capacitance, R is (R_1+R_2+r), r is ESR of capacitor C and factor k is,

$$k=\frac{R_2+r}{R}$$

For correct functioning, the factor k must not exceed 0.333. If it does, the 555 circuit will oscillate at a very high frequency, governed mainly by the 555 IC propagation delay.

As the value of k increases through the range from zero to 0.31, the oscillator frequency increases exponentially, by a factor of ×10. A further increase in the value of k from 0.310 to 0.331 gives a further ×10 increase in output frequency.

Using R_1=4.7kΩ, ±1% and R_2=2.2kΩ, ±1%, then for ESR=r=0, k is nominally equal to 0.3188.

If the ESR, r now increases to say 100Ω, k increases by around 3% to 0.3286. The circuit now gives an output frequency that is approximately three times the value of the frequency when r=0.

Note that an increase in resistance, r (ESR) gives an exponential increase in output frequency.

Test method. Calibrate the circuit by using a known good low-ESR capacitor of the same capacitance value as the device under test, plus a low value resistance box. Alternatively, use a selection of 1 to 150Ω 1% tolerance resistors.

Plot a calibration graph of output frequency versus r value, over the range zero to 150Ω. Connect the capacitor under test into the circuit and determine the ESR value from the output frequency measured, by means of the graph.

In order to give good resolution for low values of ESR, of, say, <10Ω, and to overcome the effect of frequency variations due to C value variations, add a fixed resistor in the range 10 to 100Ω ±1% tolerance, in series with the capacitor under test.

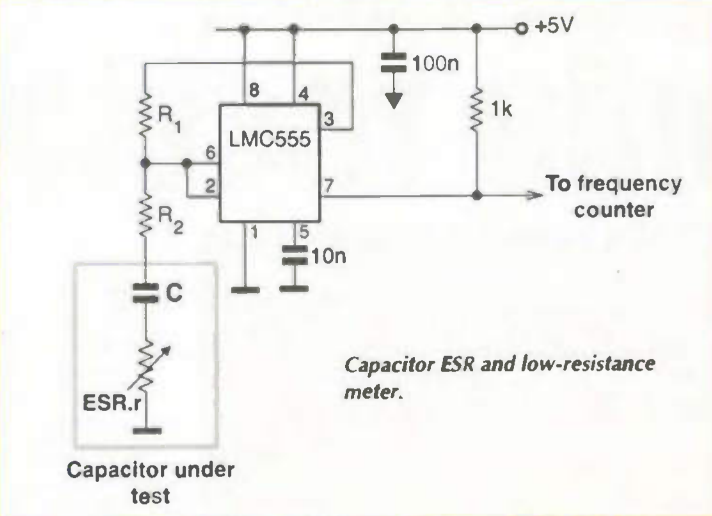
Determine the value of r as above, then subtract the additional resistor value, to determine the internal capacitor ESR value.

Precautions. Use a well regulated DC supply and decoupling capacitors, to avoid frequency instability.

Remember that capacitor ESR is both frequency and temperature dependent.

Other uses.

1. High resolution milli-ohm low resistance measurement. Use a fixed, high grade, low-ESR capacitor e.g. polystyrene or polypropylene dielectric and connect resistor under test close to the oscillator circuit.
2. Oscillator with grounded frequency control.
3. Oscillator where increase in resistance is to give increase in frequency, e.g. PTC thermistor measurement, control or temperature compensation.
4. High resolution temperature measurement, with 100Ω PRT sensors up to +100°C.



Capacitor ESR and low-resistance meter.

£75 winner

Typical results

With C=0.1µF, R_1=4.7kΩ, R_2=2.2kΩ.

r (Ω)	k factor	Freq. out (kHz)
0	0.3188	16.37
10	0.3198	17.5
20	0.3208	18.8
50	0.3237	24.3
100	0.3285	47.1
110	0.3295	58.2
120	0.3305	75.3
130	0.3314	108
140	0.3324	173
150 (max)	0.3333	230

Note – with r greater than 100Ω, a resolution of better than 1mΩ can be obtained.

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Designing radio receivers III

Joe Carr explains what the noise floor is, and reveals how you can interpret the various ways that receiver makers express static measurements of performance. This is the third in a set of four articles that cover designing receivers from the ground up.

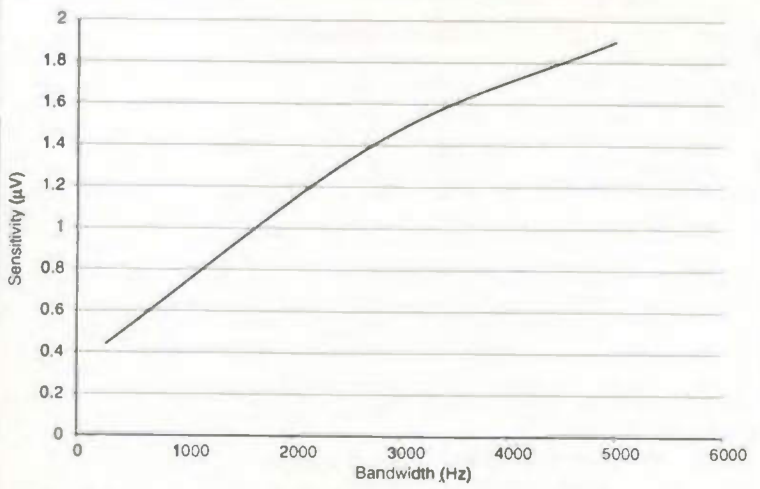

Fig. 1. Some receiver makers specify sensitivity only at lower bandwidths to make the set's performance look better than it is.

In my previous article I looked at the various stages of a superheterodyne receiver, and introduced the subject of the performance parameters of the receiver. This article investigates the receiver's noise floor and describes how static measurements of performance are made.

The noise floor of a receiver is a statement of the amount of noise produced by the receiver's internal circuitry. It directly affects the sensitivity of the receiver.

Typically, the noise floor is expressed in dBm. The noise floor specification is evaluated as follows: the more negative the better. The best receivers have noise floor numbers of less than –130dBm, while some very good receivers offer numbers of –115dBm to –130dBm.

The noise floor is directly dependent on the bandwidth used to make the measurement. Receiver advertisements usually specify the bandwidth. But be careful to note whether or not the bandwidth that produced the very good performance numbers is also the bandwidth that you'll need for the mode of transmission you want to receive.

If, for example, you are interested only in weak 6kHz wide AM signals, and the noise floor is specified for a 250Hz CW filter, then the noise floor might be too high for your use.

Receiver performance under static conditions

There are two principal static levels of performance for radio receivers, namely sensitivity and selectivity.

Sensitivity refers to the level of input signal required to produce a usable output signal, and is variously defined. The selectivity refers to the ability of the receiver to reject adjacent channel signals – again, variously defined.

Let's take a look at both of these factors. Keep in mind though, that in modern high-performance radio receivers, the static measures of performance, although frequently cited, may also be the least relevant compared with the dynamic measures. This is especially so where environments with high interference levels are involved.

Sensitivity

Sensitivity is a measure of the receiver's ability to pick up – i.e. 'detect' – signals. It is often specified in microvolts.

A typical specification might be '0.5µV sensitivity'. The question to ask is: "relative to what?" The sensitivity number in microvolts is meaningless unless the test conditions are specified.

For most commercial receivers, the usual test condition is the sensitivity required to produce a 10dB signal-plus-noise-to-noise ($S+N/N$) ratio in the mode of interest. For example, if only one sensitivity figure is given, you have to find out what bandwidth is being used. The 5 to 6kHz for AM, 2.1 to 3kHz for single sideband, 1.8kHz for radioteletype or 200 to 500Hz for CW.

Indeed, one of the places where 'creative specification writing' takes place for commercial receivers is in advertisements. Some will enthusiastically cite the sensitivity for a narrow bandwidth mode – CW for example – while the other specifications are cited for a more commonly used wider bandwidth mode such as SSB.

In one particularly egregious example, an advertisement claimed a sensitivity number that was applicable to the 270Hz CW mode only, yet the 270Hz CW filter was an expensive option that had to be specially ordered separately!

The amount of sensitivity improvement is seen by evaluating some simple numbers. Recall that a claim of 'x-µV' sensitivity refers to some standard such as 'x-µV to produce a 10dB signal-to-noise ratio in y-Hz bandwidth.'

Consider the case where the main mode for a high frequency (HF) shortwave receiver is AM for international broadcasting. The sensitivity is 1.9µV for 10dB SNR, and the bandwidth is 5kHz.

If the bandwidth were reduced to 2.8kHz for SSB, then the sensitivity improves by the square root of the ratio, or $\sqrt{(5/2.8)}$. If the bandwidth is further reduced to 270Hz – i.e. 0.27kHz – for CW, then the sensitivity for 10dB SNR is $\sqrt{(5/0.27)}$.

The 1.9µV AM sensitivity therefore translates to 1.42µV for SSB and 0.44µV for CW. If only the CW version is given, then the receiver might be made to look a whole lot better than it is, even though the typical user may never use the CW mode. Note the differences in **Fig. 1**.

The sensitivity differences also explain why weak SSB signals can be heard under conditions when AM signals of similar strength have disappeared into the noise, or why the CW mode has as much as 20dB advantage over SSB, *ceteris paribus*.

In some receivers, the difference in mode – AM, SSB, RTTY, CW, etc. – can conceivably result in sensitivity differences that are more than the differences in the bandwidths associated with the various modes. The reason is that there is sometimes a 'processing gain' associated with the type of detector circuit used to demodulate the signal at the output of the IF amplifier.

A simple AM envelope detector is lossy because it consists of a simple diode, such as a 1N60, and an RC filter, which is a passive circuit without amplification. Other detectors, including product detectors for SSB and synchronous AM detectors, have their own signal gain. As a result, they may produce better sensitivity numbers than the bandwidth suggests.

Another indication of sensitivity is minimum detectable signal, or MDS, which is usually specified in dBm. This signal level is the signal power at the antenna input terminal of the receiver required to produce some standard $S+N/N$ ratio, such as 3dB or 10dB, **Fig. 2**.

In radar receivers, the MDS is usually described in terms of a single pulse return and a specified $S+N/N$ ratio.

Also, in radar receivers, the sensitivity can be improved by integrating multiple pulses. If N return pulses are integrated, then the sensitivity is improved by a factor of N if coherent detection is used, and $\sqrt{N}$ if non-coherent detection is used.

Modulated signals represent a special case. For those sensitivities, it is common to specify the conditions under which the measurement is made. For example, in AM receivers the sensitivity to achieve 10dB SNR is measured with the input signal modulated 30 percent by a 400 or 1000Hz sinusoidal tone.

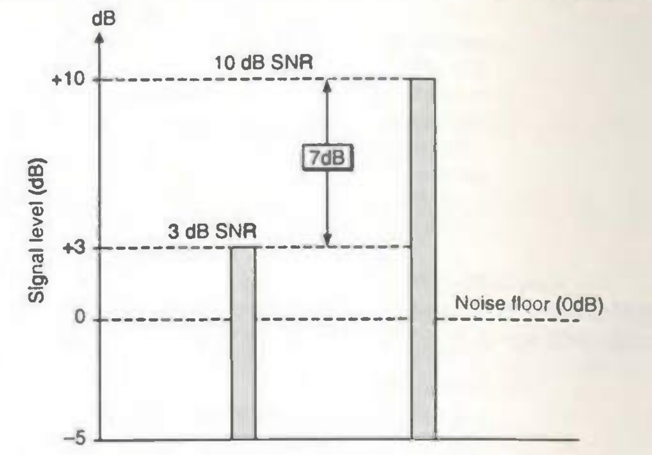

Fig. 2. Sensitivity may be expressed as a minimum detectable signal, or MDS. This signal level is the signal power needed at the antenna input terminal of the receiver to produce some standard S+N/N ratio, such as 3dB or 10dB.

An alternate method is sometimes used for AM sensitivity measurements, especially in consumer radio receivers. Here, SNR may be a little hard to measure with the equipment normally available to technicians who work on those radios. This is the 'standard output conditions' method.

Some manuals will specify the audio signal power or audio signal voltage at some critical point, when the 30% modulated RF carrier is present. In one automobile radio receiver, the sensitivity was specified as "X µV to produce 400mW across 8Ω resistive load substituted for the loudspeaker when the signal generator is modulated 30% with a 400Hz audio tone."

The cryptic note on the schematic showed an output sine wave across the loudspeaker with the label "400mW in 8Ω (1.79 volts), @ 30% mod. 400Hz, 1µV RF." What is missing is mention of the level of total harmonic distortion, or THD, that is permitted.

The sensitivity is sometimes measured in essentially the same way. However, the signal levels will specify the voltage level that will appear at the top of the volume control, or output of the detector/filter, when the standard signal is applied. Thus, there are two ways seen for specifying AM sensitivity: 10dB SNR and standard output conditions.

There are also two ways to specify FM receiver sensitivity. The first is the 10dB SNR method discussed above. In other words, this is the number of microvolts of signal at the input terminals required to produce a 10dB SNR when the carrier is modulated by a standard amount.

The measure of FM modulation is deviation expressed in kilohertz. Sometimes, the full deviation for that class of receiver is used, while for others a value that is 25 to 35 percent of full deviation is specified.

The second way to measure FM sensitivity is the level of signal required to reduce the no-signal noise level by 20dB. This is the 20dB quieting sensitivity of the receiver. If you tune between signals on an FM receiver, you will hear a loud 'hiss' signal – especially in the VHF/UHF bands.

Some of that noise is externally generated, while some is internally generated. When an FM signal appears in the passband, that hiss is suppressed, even if the FM carrier is unmodulated. The quieting sensitivity of an FM receiver is a statement of the number of microvolts required to produce some standard quieting level, usually 20dB.

Pulse receivers, such as radar and pulse communications units, often use the tangential sensitivity as the measure of performance. This is the amplitude of pulse signal required to raise the noise level by its own RMS amplitude, **Fig. 3**.

Selectivity

Although no receiver specification is unimportant, if one had to choose between sensitivity and selectivity, the proper choice most of the time would be to take selectivity.

Selectivity is the measure of a receiver's ability to reject adjacent-channel interference. Or put another way, it's the ability to reject interference from signals on frequencies close to the desired signal frequency.

In order to understand selectivity requirements, you must first understand a little bit of the nature of radio signals. An unmodulated radio carrier theoretically has an infinitesimal – i.e. near-zero – bandwidth, although all real unmodulated carriers have a very narrow, but non-zero, bandwidth because they are modulated by noise and other artifacts.

As soon as the radio signal is modulated to carry information, however, the bandwidth spreads. Even an on/off telegraphy (CW) or pulse signal spreads out either side of the carrier frequency an amount that is dependent on the sending speed and the shape of the keying waveform.

An AM signal spreads out an amount equal to twice the highest audio modulating frequencies. For example, a communications AM transmitter will have audio components from 300 to 3000Hz, so the AM waveform will occupy a spectrum that is equal to the carrier frequency, *F*, plus and minus the audio bandwidth – or *F* ±3000Hz in the case cited.

An FM carrier spreads out according to the deviation. For example, a narrow-band FM landmobile transmitter with 5kHz deviation spreads out ±5kHz, while FM broadcast transmitters spread out ±75kHz

An implication of the fact that radio signals have bandwidth is that the receiver must have sufficient bandwidth to recover all of the signal. If it doesn't, information may be lost and the output is distorted.

On the other hand, allowing too much bandwidth increases the noise picked up by the receiver and thereby deteriorates the SNR. The goal of the selectivity system of the receiver is to match the bandwidth of the receiver to that of the signal. That is why receivers will use 270 or 500Hz bandwidth for CW, 2 to 3kHz for SSB and 4 to 6kHz for AM signals. They allow matching the receiver bandwidth to the transmission type.

The selectivity of a receiver has a number of aspects that must be considered: front-end bandwidth, IF bandwidth, IF shape factor, and the ultimate – or distant frequency – rejection.

Front-end bandwidth

The 'front-end' of a modern superheterodyne radio receiver is the circuitry between the antenna input terminal and the output of the first mixer stage.

The reason why front-end selectivity is important is that it prevents out-of-band signals from afflicting the receiver. Transmitters located nearby can easily overload a poorly designed receiver.

Even if these signals are not heard by the operator, they can desensitise a receiver, or create harmonics and intermodulation products that show up as 'birdies' or other types of spurious interference on the receiver. Strong local signals can take up a lot of the receiver's dynamic range, and thereby make it harder to hear weak signals.

In some crystal video microwave receivers, that front-end might be wide open without any selectivity at all, but in nearly all other receivers there will be some form of frequency selection present.

Two forms of frequency selection are typically found. A designer may choose to use only one of them in a design. Alternatively, both might be used in the design, but separately, under control of the operator. Or finally, both might be used together. These forms can be called the resonant-frequency filter, **Fig. 4a)** and bandpass filter, Fig. **4b)**, approaches.

The resonant-frequency approach uses *LC* elements tuned to the desired frequency to select which RF signals reach the mixer. In some receivers, these *LC* elements are designed to track with the local oscillator that sets the operating frequency. That's why you see two-section – or three-section – variable capacitors for AM broadcast receivers with two different capacitance ranges for the sections. One section tunes the LO and the other section tunes the tracking RF input. In other designs, a separate tuning knob – 'preselector' or 'antenna' is used.

The other approach uses a sub-octave bandpass filter to admit only a portion of the RF spectrum into the front-end. For example, a shortwave receiver that is designed to take the HF spectrum in 1MHz pieces may have an array of RF input bandpass filters that are each 1MHz wide – for example 9 to 10MHz.

In addition to the reasons cited above, front-end selectivity also helps improve a receiver's image rejection and 1st IF Rejection capabilities.

Image rejection

An image in a superheterodyne receiver is a signal that appears at twice the IF distance from the desired RF signal. It is located on the opposite side of the LO frequency from the desired RF signal.

In **Fig. 5**, a superheterodyne operates with a 455kHz (i.e. 0.455MHz) IF, and is tuned to 24.0MHz (F_{RF}). Because this receiver uses low-side LO injection, the LO frequency F_{LO} is 24.0–0.455, or 23.545MHz.

If a signal appears at twice the IF below the RF – i.e. 910kHz below FRF – and reaches the mixer, then it too has a difference frequency of 455kHz, so will pass right through the IF filtering as a valid signal. The image rejection specification tells how well this image frequency is suppressed. Normally, anything over about 70dB is considered good.

Tactics to reduce image response vary with the design of the receiver. At design time, the best approach is to select an intermediate frequency that is high enough so that the image frequency will fall outside the passband of the receiver front-end.

Some HF receivers use an IF of 8.83MHz, 9MHz, 10.7MHz or something similar. For image rejection these frequencies are considerably better than 455kHz receivers in the higher HF bands. However, a common design trend is to do double conversion.

In most such designs, the first IF frequency is considerably higher than the RF, being in the range 35 to about 80MHz. A frequency of 50MHz is common in HF receivers, 70MHz in microwave receivers.

The high IF makes it possible to suppress the VHF images with a simple low-pass filter. If the 24.0MHz signal (above) were first up-converted to 50MHz (74MHz LO), for example, the image would be at 124MHz. The second conversion brings the IF down to one of the frequencies mentioned above, or even 455kHz.

The lower frequencies are preferable to 50MHz for bandwidth selectivity reasons because good quality crystal, ceramic or mechanical filters in the lower frequency ranges are easily available.

Rejecting the first IF

The 1st IF rejection specification refers to how well a receiver rejects radio signals operating on the receiver's first IF frequency. For example, if your receiver has a first IF of 50MHz, it must be able to reject radio signals operating on that frequency when the receiver is tuned to a different frequency.

Although the shielding of the receiver is also an issue with respect to this performance, the front-end selectivity affects how well the receiver performs against first IF signals.

If there is no front-end selectivity to discriminate against signals at the IF frequency, then they arrive at the input of the mixer unimpeded. Depending on the design of the mixer, they then may pass directly through to the high gain IF amplifiers and can be heard in the receiver output.

IF bandwidth

Most of the selectivity of the receiver is provided by the filtering in the IF amplifier section. The filtering might be *LC* based – especially if the principal IF is a low frequency like 50kHz – or a ceramic resonator, a crystal filter or a mechanical filter. Of these, the mechanical filter is usually regarded as best for narrow bandwidths, with the crystal filter and ceramic filters coming in next.

The IF bandwidth is expressed in kilohertz, and is measured from the points on the IF frequency response curve where gain drops off –3dB from the mid-band value, **Fig. 6**. This is why you will sometimes see selectivity specified in terms such as '6kHz between –3dB points.'

The IF bandwidth must be matched to the bandwidth of the received signal for best performance. If a too-wide bandwidth is selected, then the received signal will be noisy, and signal-to-noise ratio deteriorates. If too narrow, then you might experience difficulties recovering all of the information that was transmitted.

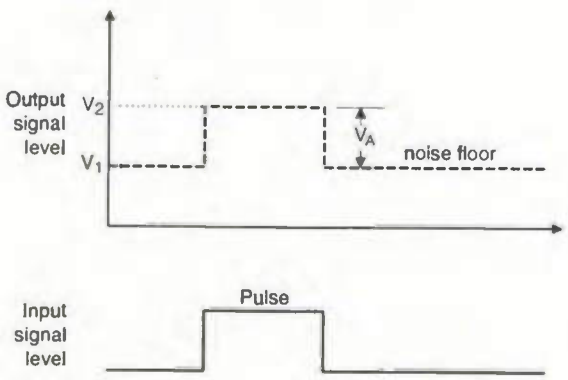

Fig. 3. Pulse receivers, such as those found in radar systems, often use tangential sensitivity as a measure of performance. This is the amplitude of pulse signal needed to raise the signal above the noise level by its own RMS amplitude.

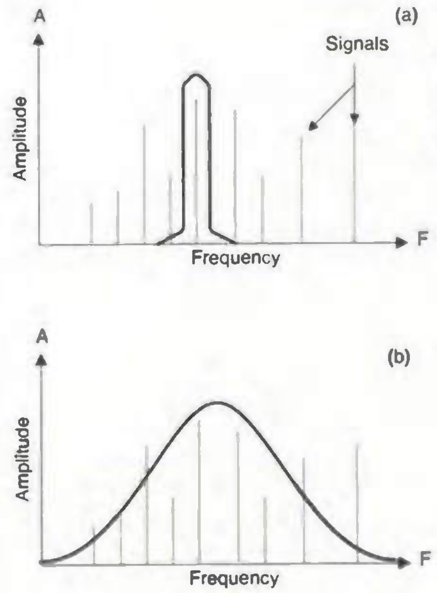

Fig. 4. Front-end bandwidth. Some receives use a resonant-frequency filter, a), others a band-pass filter, b).

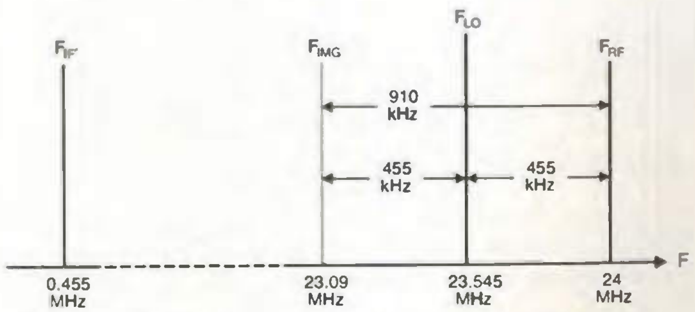

Fig. 5. In a superhet, if a signal that's the same amount below the local oscillator frequency as the wanted signal is above it, interference is caused. A receiver's ability to reject this unwanted signal is its image rejection capability.

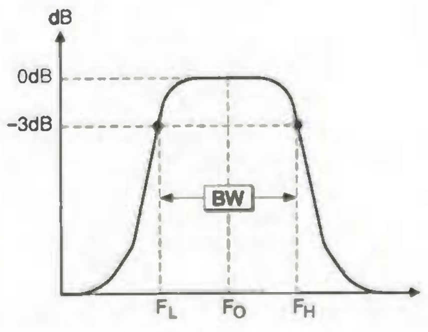

Fig. 6. Intermediate-frequency bandwidth is expressed in kilohertz, and measured from the points on the intermediate frequency response curve where gain drops off –3dB from the mid-band value.

For example, an AM broadcast-band radio signal has audio components out to 5kHz, so the signal occupies up to 10kHz of spectrum space ($F\pm5$kHz). If a 2.8kHz SSB IF filter is selected, then it will tend to sound 'mushy' and distorted.

IF passband shape factor

The shape factor is a measure of the steepness of the receiver's IF passband. It is taken by measuring the ratio of the bandwidth at –6dB to the bandwidth at –60dB, **Fig. 7a)**. The general rule is that the closer these numbers are to each other, the better the receiver.

Anything in the 1:1.5 to 1:1.9 region can be considered high quality, while anything worse than 1:3 is not worth looking at for 'serious' receiver uses. If the numbers are between 1:1.9 and 1:3, then the receiver could be regarded as being middling, but useful.

The importance of shape factor is that it modifies the notion of bandwidth. The cited bandwidth – 2.8kHz for SSB for example – does not take into account the effects of strong signals that are just beyond those limits, Such signals can easily 'punch through' the IF selectivity if the IF passband 'skirts' are not steep. After all, the steeper they are, the closer a strong signal can be without messing up the receiver's operation.

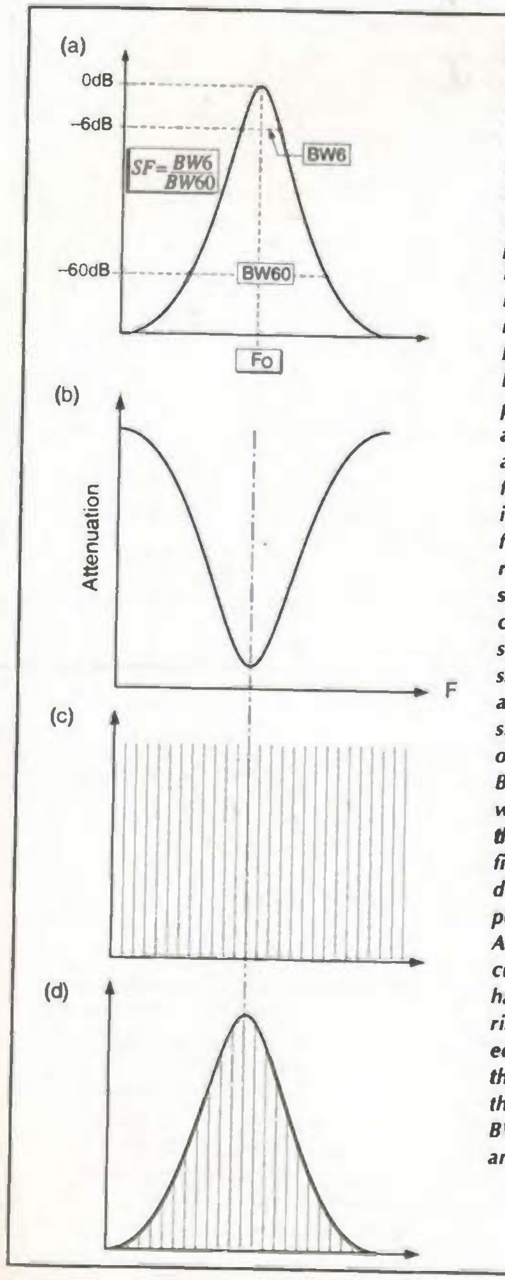

Fig. 7a). Shape factor, SF, is a measure of the steepness of a receiver's intermediate-frequency passband, a). It is important as it modifies the notion of bandwidth. Inverse curve b) plots the filter's attenuation against frequency. In c) is a range of frequencies reaching the IF stage. If the IF curve is shallow, as shown, high-amplitude signals just outside the BW60 limit in a) will pass through the filter and degrade performance. An ideal transfer curve would have vertical rise and fall edges cutting through both the BW60 and BW6 points, and a flat top.

The situation is illustrated in Fig. **7b)**. This curve inverts Fig. 7a) by plotting attenuation versus frequency. Assume that equal amplitude signals close to F_O are received, Fig. **7c)**, the relative post-filtering amplitudes will match Fig. **7d)**. Thus, selecting a receiver with a shape factor as close to the 1:1 ideal as possible will result in a more usable radio.

Ultimate rejection

A receiver's distant-frequency or 'ultimate' rejection specification tells you something about the receiver's ability to reject very strong signals that are located well outside the receiver's IF passband. This number is stated in negative decibels, and the higher the number the better.

An excellent receiver will have values in the –60 to –90dB range, a middling receiver will see numbers in the –45 to –60dB range, and a terrible receiver will be –44 or worse.

Stability

The stability specification measures how much the receiver frequency drifts as time elapses or temperature changes.

Local-oscillator drift sets the overall stability of the receiver. This specification is usually given in terms of short-term drift and long-term drift – for example from local-oscillator crystal aging. Short term drift is important in daily operation, while the long term drift ultimately affects general dial calibration.

If the receiver is controlled by a variable-frequency oscillator, or VFO, or it uses partial frequency synthesis, which combines VFO with crystal oscillators, then the stability is dominated by the VFO stability. In fully synthesised receivers, the stability is governed by the master-reference crystal oscillator.

If either an oven-controlled crystal oscillator (OCXO), or a temperature compensated crystal oscillator (TCXO), is used for the master reference, then stability on the order of 1 part in 108°C is achievable.

For most users, short-term stability is most important – especially when tuning SSB, ECSS or RTTY signals. A common specification value for a good receiver will be 50Hz/hour after a three hour warm-up, or 100Hz/hour after a 15 minute warm-up. The smaller the drift the better the receiver.

The foundation of good stability is at design time. The local oscillator, or VFO portion of a synthesiser, must be operated in a cool, temperature stable, location within the equipment, and must have the correct type of components. Capacitor temperature coefficients are selected in order to cancel out temperature-related drift in inductance values.

Post-design-time changes can also help, but these are less likely to be possible today than in the past. The chief cause of drift problems is heat. In the days of vacuum-tube oscillators, the heat of the internal heating filament produced lots of heat that in turn created drift.

A related phenomenon seen on low-cost receivers, or certain home-brewed receivers of doubtful merit, is mechanical frequency shifts. Although not seen on most modern receivers – even some very cheap designs – it was once a serious problem on less costly models. This problem is usually seen on VFO-controlled receivers in which vibration to the receiver cabinet imparts movement to either the inductor, L, or capacitor, C, element in an LC VFO.

Mechanically stabilising these components will work wonders.

AGC range and threshold

Modern communications receivers must be able to handle signal strengths over a dynamic range of about 1 000 000:1. Tuning across a band occupied by signals of wildly varying strengths is hard on the

ears and hard on the receiver's performance. As a result, most modern receivers have an automatic gain control (AGC) circuit that smooths out these changes.

Automatic gain control will reduce gain for strong signals, and increase it for weak signals It can be turned off on most HF and VHF/UHF communications receivers.

The AGC range is the change of input signal (in dBμV) from some reference level, such as 1V EMF, to the input level that produces a 2dB change in output level. Ranges of 90 to 110dB are common.

The AGC threshold is the signal level at which the AGC begins to operate. If set too low, then the receiver gain will respond to noise, and irritate the user. If set too high, then the user will experience irritating shifts of output level as the band is tuned.

AGC thresholds of 0.7 to 2.5μV are common on decent receivers, with the better receivers being in the 0.7 to 1μV range.

Another AGC specification sometimes seen deals with the speed of the AGC. Although sometimes specified in milliseconds, it is also frequently specified in subjective terms like 'fast' and 'slow.'

This specification refers to how fast the AGC responds to changes in signal strength. If set too fast, then rapidly keyed signals like on-off CW, or noise transients will cause unnerving large shifts in receiver gain. If set too slow, then the receiver might as well not have an AGC. Many receivers provide two or more selections in order to accommodate different types of signals. ■

Joe's fourth and final article on receiver design looks at measuring of performance under dynamic conditions.

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Douglas Self

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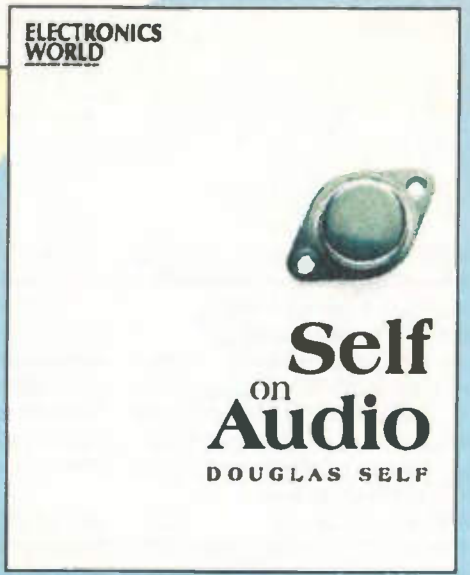

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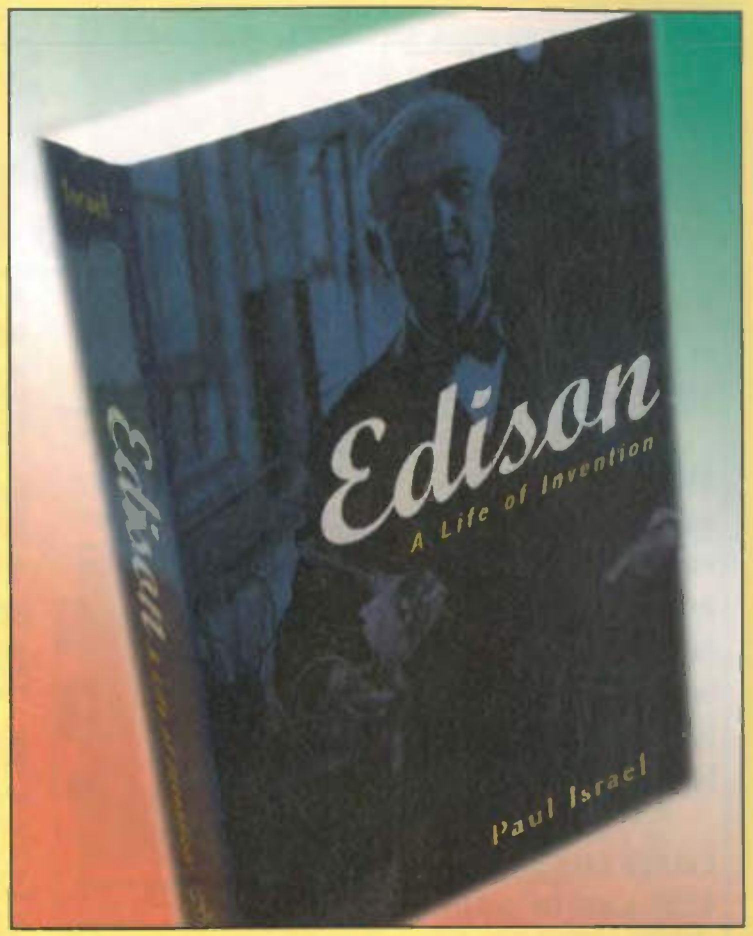

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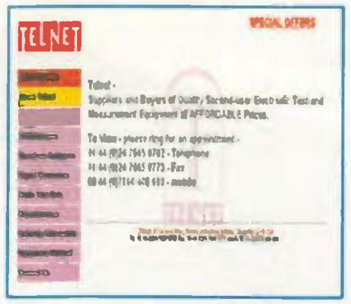

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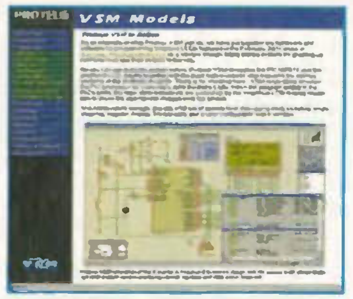

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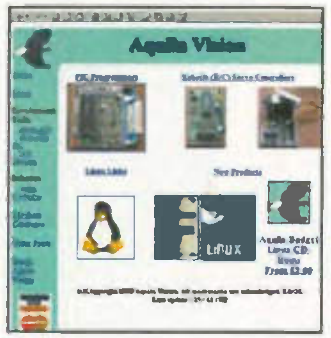

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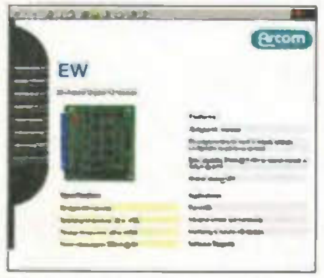

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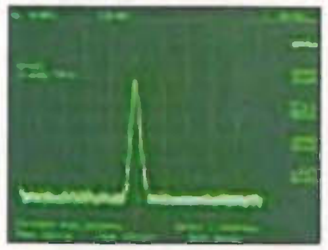

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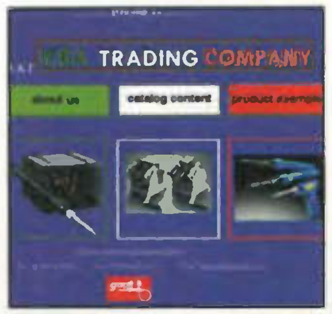

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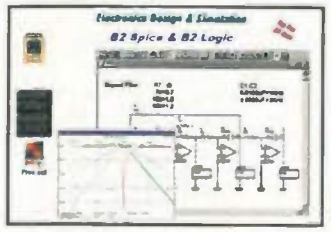

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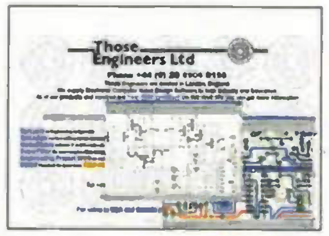

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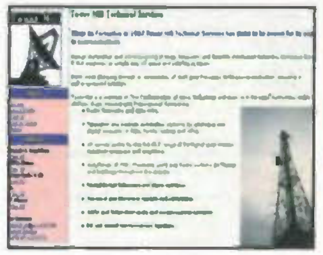

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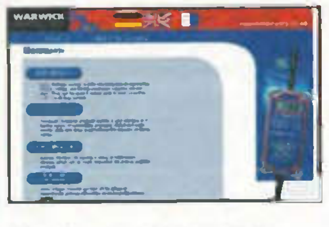

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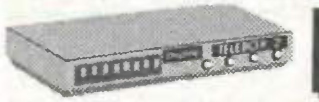

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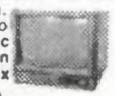

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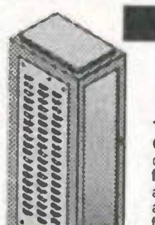

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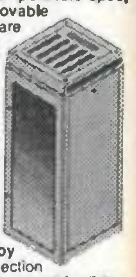

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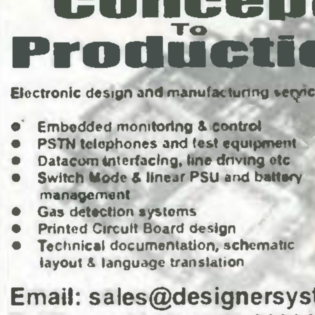

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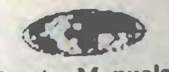

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