Designing receivers from the ground up – new series

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Design competition winners:

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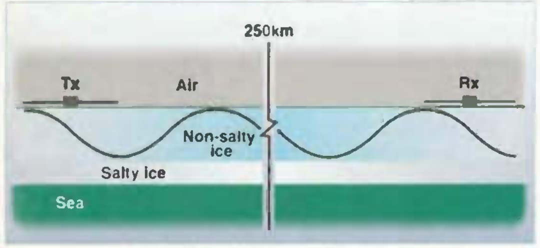
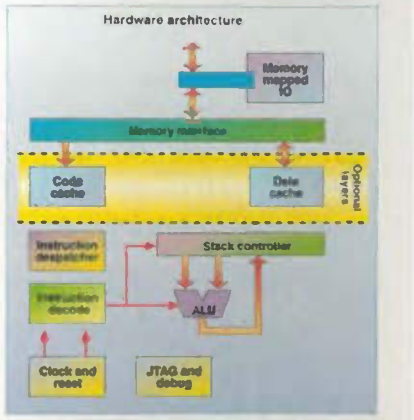
CONTENTS



Researchers have used the non-salty layer of Arctic ice as a waveguide to communicate over 100 miles. Find out how on page 572.

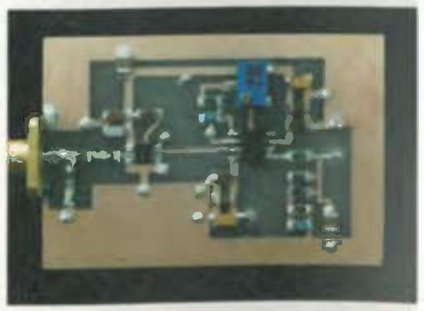
Cyril Bateman's RF millivoltmeter exploits an innovative true RMS-to DC converter chip to measure signals at well over 100MHz – page 592.

Illustration: Hashim Akib



Steve Bush reports on how Java is booming in embedded applications – page 584.

Ian Hickman explains how to design power transformers in this month's beginners' corner – page 643.



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Has DAB really arrived?

In recent years, *Electronics World* has carried several articles anticipating the arrival of digital audio broadcasting, and reporting on tests and demonstrations. DAB has now officially arrived, so what is it like in reality? To sum up, I find DAB disappointing.

I am not necessarily criticising the DAB system as a piece of technology. It may well be adequate for its purpose, although one could question whether there is really a need for yet more pop music stations even if they are aimed at differing audiences. The car-radio re-tuning problem has already been solved by RDS. My concern is the UK implementation of DAB, particularly for local services.

Audio quality is at its best similar to FM – DAB has a wider frequency response, but the bit censor creates some roughness from time to time. At worst it is bettered by AM. The system seems to have particular trouble with the spoken voice, and complex orchestral passages; I usually still prefer to listen on FM. DAB copes OK with pop music though.

Things could be improved by using higher bit rates, but this would reduce the number of stations in a multiplex and so upset the economic arguments for DAB. As you might expect, the BBC generally has better audio quality than commercial stations.

The biggest problem is severe co-channel interference. There are at present only seven channels, and two are used for national multiplex. Each major conurbation then needs to use about four out of the remaining five channels for regional and local multiplex. The result is that frequencies are being re-used much closer than good engineering practice would allow.

My local multiplex in Birmingham is being clobbered so badly by co-channel interference from Manchester that my tuner cannot resolve it at all. The Wolverhampton multiplex is audible but badly corrupted, despite ample signal strength, due to interference from Liverpool or Bristol.

My local topography seems to screen me from nearby transmitters – apart from Sutton Coldfield, but that does not carry the local multiplex – but long-distance signals may arrive at a slightly higher angle so hop over the local hills.

The DAB service is soon to be rolled out to new areas. If all channels are already in use in the Manchester, Birmingham and London areas, and already interfering with each other, what will happen in, say, Stoke-on-Trent or Oxford? Will I eventually lose the only local multiplex I can hear – Coventry – when this channel gets re-used rather closer than it is at present?

Local screening is not uncommon, but at least on FM or TV you may be able to tune to an alternative transmitter carrying the same station on a different channel. At present I can receive a wide variety of FM stations, so if I get bored with Radio WM I can listen to Radio Leicester or Radio Stoke; analogue switch-off with DAB will eventually remove this choice.

Once you admit that it is not possible to re-use every channel in every region, then the four colour map theorem suggests that you probably need four sets of channels – i.e. 16 channels, for one regional and three local multiplex – for the major regions, plus some more for intermediate towns and cities. So there could be a need for over 20 channels just to provide a reliable regional and local service, rather than the five currently available.

As a side-effect this would allow some 'out of area' reception, which the Radio Authority might frown upon because it complicates commercial considerations by increasing consumer choice.

In some locations it may even be found that the national Single-Frequency Networks fail due to long-distance propagation. The much-vaunted multipath enhancement properties of DAB only work for nearby transmitters carrying the same multiplex. Over 75km path difference even the same multiplex counts as co-channel interference. 75km is not a huge distance on Band 3 for a well-sited transmitter.

DAB adopted vertical polarisation, presumably to help car and portable receivers operating near ground level. But then it uses low power transmitters, so these receivers won't get enough signal in any compromised location. Vertical polarisation means that Yagi antennas don't have a null on the side, so a fixed installation cannot easily eliminate co-channel interference.

I fear that unless DAB is properly re-engineered it will acquire the same reputation as Digital Terrestrial Television is gaining – unreliability and poor signal quality. DAB needs more channels even for the existing service, and many more when new areas join in. Transmitter numbers and powers need to be increased if car and portable receivers are to work reliably. DAB must not be used to restrict listener choice.

The Government and the Radio Authority need to re-think, and turn DAB from a 'pilot' into a 'service'. Otherwise 'Digital' won't mean 'new and exciting' but rather, 'expensive and flaky'. ■

Dave Kimber G8HQP, Davant Technology Ltd

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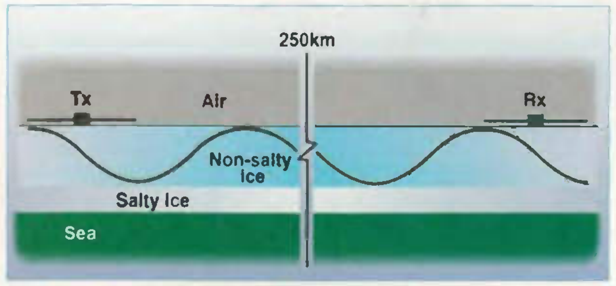
Researchers use ice as a waveguide

A US researcher has used Arctic ice – over a hundred miles of it – as a waveguide.

On a recent trip to the Arctic Paul Mileski, an electronics engineer and scientific diver with the US Naval Undersea Warfare Center, found time to take a detailed look at using ice as a giant two-dimensional waveguide and got some surprisingly good results.



"Surprising because sea ice has quite a high salt content," he said. "We got a range of over 150 miles [240km] on 10 to 100mW and the receiver [antenna] was a clip-lead from a multimeter."

Mileski has worked out what is going on. "The way it behaves in a classic sense is as a conductive surface on top of which is a dielectric slab. The relatively low loss dielectric supports TEM (transverse electric-magnetic) waves bounded by the sea water."

Coupling radio waves into the ice has been done in two ways. Mileski has found that an insulating wire with an exposed conductive end can be lowered through a narrow hole in the ice. "We were getting very successful results with a length of wire terminated in sea water with a meshing network [to match it] on top," he said.

This technique requires a long drill, so he favours the second method; roll out a long insulated dipole across the ice. Mileski finally settled on two 150m wires to couple the 600m wavelength (500kHz) radio signal into the ice. This dipole does not act like a classic antenna, with a null in line with its length, but rather as a coupling structure that has a peak in line and a null at right angles.

Voice and 75bit/s data were sent and the signal successfully leaps gaps in the ice 30m wide. One problem is noise generated by charged show particles hitting the wire. Covering the wire with snow cuts down the noise.

What use is the whole thing? "It would be terrific for emergency communications," said Mileski.

Optical demultiplexer finds a new application

A device commonly used to demultiplex different wavelengths sent over fibre optic lines might ultimately be used for high-speed time-domain transmission in optical fibres says engineers from Purdue University.

"This is the first time that anybody has realised this technology could be modified for a different function," said Andrew Weiner, a professor of electrical and computer engineering.

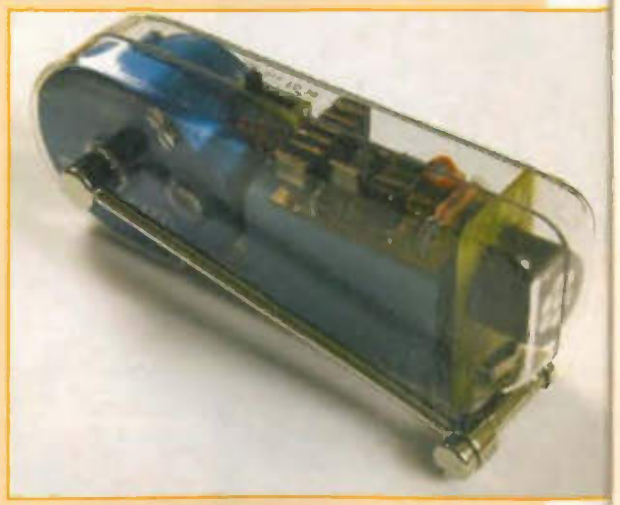
A modified commercial 'arrayed waveguide grating' can, according to Purdue, turn a single pulse of laser light into a rapid-fire burst of 21 pulses, each separated by only 2ps – ten times faster than the transmission speed of each channel in state-of-the-art commercial optical communication systems, claimed Weiner.

"We realised that, rather than just being used to separate wavelengths, these arrayed waveguide gratings might have another application that nobody seemed to recognise," Weiner said. "If you send a pulse of light into one of these devices you can get a burst of pulses coming out."

Energy store wind up

Wind-up technology may get a boost due to developments at Atkin Dèsign and Development. Howard Atkin, a self-confessed efficiency nut, runs the company. "My background is in squeezing the very last bit of efficiency out of a product. Sometimes for no real reason except that it feels right," he said.

Rather than store energy in a spring like Trevor Bayliss' radio designs, Atkin is using supercapacitors charged from a generator.



For commercial reasons, Atkin will not discuss any details of the gear-generator-store combination except to say that it is "scalable, from very small to very big. Half a watt to 50 watts", he said.

He also said that the generator is nothing like the brush-type motor that Bayliss uses.

Atkin won a DTI Smart feasibility award in 2000, which he has used to develop various ideas in the concept. "I have used the best universities in the country to develop parts and a specialist company to optimise the gears," he said. "These have been developed with a system approach. If you are not careful you can end up with a lot of good components that don't work together."

The next phase is to make the power source ready for production. "I have parts which are the best that can be built, made from gold, frankincense and myrrh. These need to be value-engineered to get them ready for production."

Fuzzy logic helps Mars rovers wander on their own

Future NASA Mars landers could navigate autonomously using a combination of fuzzy logic and neural network techniques. From the left, the pictures show an original image, ground-landscape separation, object location, then a three-image panorama of the route chosen.

NASA is looking at neural networks and fuzzy logic to control its next generation of planet rovers.

"We want to tell the robot to think about any obstacle it encounters just as an astronaut in the same situation would do," said Dr Ayanna Howard, artificial intelligence scientist at the Jet Propulsion Laboratory in California. "Our job is to help the robot think in more logical terms about turning left or right, not just by how many degrees."

The ultimate goal for Howard and colleagues is "putting a robot on Mars and walking away, leaving it to work without direct human interaction," she said.

The prototype operates on images from on-board cameras and chooses its own route to operator-selected locations. On-board processing attempts to take into account the general lie of the land, possible pitfalls like ditches and rocky patches, and immediate obstacles.

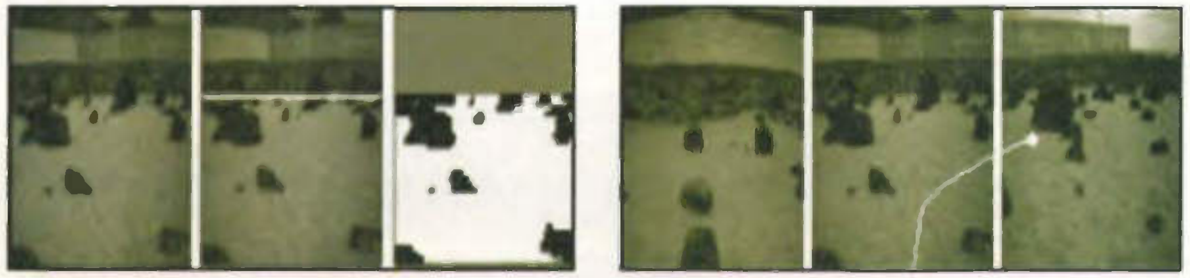
This surveying involves initially identifying linear features to locate edges in the landscape. These edges allow the local terrain to be separated from distant objects, as well as spotting potential ditches and cliff edges. Then the local terrain is analysed for rocks, the size – 'small' or 'large' – of the rocks and their location relative to each other.

Analysed terrain data feeds into a layer of behavioural models, one for the main aim, to get to point X, one for terrain based navigation, finding likely routes, and one for local collision avoidance. The behavioural models then vie to persuade a final analysis block where to drive the rover.

At the moment, the whole thing runs on a 333MHz Pentium II Linux-based computer in a Compact PCI chassis. Data comes from six high-mounted CMOS NTSC video cameras through dual four-input framegrabber boards. The whole lot sits on a prototype rover with a trailer.

Example fuzzy table. This one looks for areas suitable to drive through.

small rock	large rock	separation	roughness
few	few		smooth
many	few		rough
many		far	rough
	many	close	rocky



A combination of fuzzy logic and neural networks enables robot pioneers to detect the obstacles in an unfamiliar terrain (left, a sequence of one image being processed), assess the relative safety of various alternative routes, and plot a path to its destination (right, a three-image panorama), all without real-time human guidance.

California design software firm guilty of stealing source code

In a criminal trial in California design software firm Avant! has been found guilty of stealing source code for place and route software from Cadence.

Avant! and seven defendants pleaded 'no contest' to charges including conspiracy, trade secret theft and securities fraud. This allowed the US District Court to find the firm guilty and order a fine of \$27m.

"We think this is a good settlement for the company," said Clayton Parker, head of corporate marketing at Avant!. "Putting this matter to rest was the best thing to do."

Gerry Hsu, Avant!'s chairman, has personally been fined \$2.7m, while six other former and current employees received fines varying from \$1 00 000 to \$2.7m. The firm is expected to pay the fines.

However, five of the six will face jail terms, including Mitch Igusa who is alleged to have sent the source code from Cadence to Avant!. He is facing up to six years imprisonment for his actions. Hsu managed to avoid being sent down, and will remain as chairman of the firm. The court has yet to decide 'restitution' damages.

Ray Bingham, president and CEO of Cadence, said: "This means that Avant! and seven individuals have accepted criminal responsibility for stealing, conspiring to steal, or concealing the theft of Cadence chip design code and will be held accountable."

Now that the criminal trial is out of the way, a civil lawsuit brought by Cadence can go ahead. Although pleading 'no contest' is not technically an admission of guilt, and should not impact the civil trial, it is bound to have some effect. Moreover, the burden of proof in the civil trial is not so great as in the criminal case.

Cadence will be asking for as much as a billion dollars in damages for the theft. Parker said the products that allegedly used the source code, ArcCell and Aquarius, brought the firm \$208m in revenue and \$125m in profit. "The company has substantial cash reserves," Parker said. This is believed to be \$250m in cash and securities.

Avant! does not expect the outcome of the trial will affect the day-to-day running of the firm. "We don't expect to lose any customers. They understand this dispute is about old tools," added Parker.

40nm-thick interconnects

Researchers at Infineon Technologies have produced metal lines just 40 to 50nm thick embedded in a dielectric film.

The team used the damascene technique to make the wires. After patterning the substrate using a phase-shifted mask, a groove was etched in the material. The substrate was then covered with metal, and the excess polished away using chemical mechanical polishing (CMP).

Moreover, an electrical assessment of the wires showed they have suitably low resistance for use as local interconnect between transistors.

According to the International Technology Road map for Semiconductors, such line widths are not required until after the year 2011.

"Pioneer 10 lives on"

Scientists operating a radio telescope in Madrid have contacted pioneer 10, lost last August, at a range of 11 billion kilometres.

"Pioneer 10 lives on," declared project manager Dr Larry Lasher of NASA. "The fact that we can still stay connected with the spacecraft is fantastic. We are overjoyed."

The hunt for the craft has been going on continually since it disappeared.

"We have been listening for the Pioneer 10 signal in a one-way downlink non-coherent transmission mode since last summer with no success," Lasher said. "We therefore concluded that in order for Pioneer 10 to talk to us, we need to talk to it."

Pioneer 10 was launched on March 2, 1972 and was the first Earth spacecraft to pass through the asteroid belt and obtain close-up images of Jupiter. It also discovered that Jupiter is predominantly liquid.

In 1983, it became the first man-made object to pass the orbit of Pluto. The spacecraft continued to gather valuable scientific data in the outer regions of the solar system until its mission ended in March 1997, 10 billion kilometres from Earth. At that distance it took nearly ten hours for radio signals to bridge the gap.

Since then the spacecraft has been used by the Deep Space Network in a concept study of chaos theory.

Pioneer 10 carries the famous gold plaque with an image of a man and a woman and information about Earth. The spacecraft headed towards the constellation of Taurus and will pass by one of the constellation stars in 2 million years time.

Point and shoot chemical analyser

Oak Ridge National Laboratory (ORNL) in Tennessee has developed a point-and-shoot chemical analyser. It uses Raman technology, hitting the sample with a helium-neon laser beam and using acousto-optic tuneable filters and a photo sensor to analyse the light that returns. This light shows the vibration energies, which are unique to each compound in the sample.

"The recent development of acousto-optic tuneable filters has made the use of Raman technology practical for applications in a variety of environments," said an ORNL.

RF imaging for medical use

Toshiba's Cambridge Research Laboratory is spinning-out TeraView, a company to exploit terahertz frequency RF imaging in the 'terahertz gap' between 10^{11} and 10^{13}.

"We are the first company that is solely dedicated to commercialising this part of the frequency range," claimed Don Arnone, TeraView's CEO. The company has first round funding from venture capitalists and Toshiba, which now has a minority shareholding.

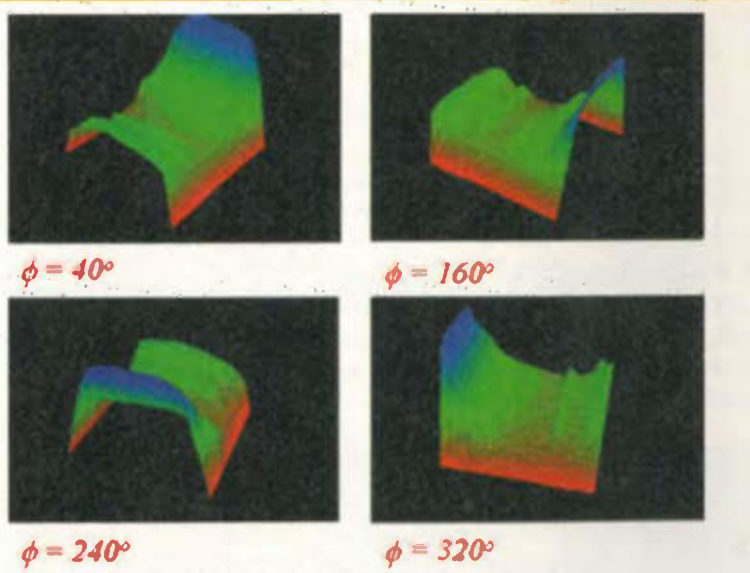
"The team will initially be around five scientists plus me and Professor Michael Pepper as scientific director," said Arnone. "We will recruit six to ten additional people with a view to hitting some milestones in 12 to 18 months time."

On these milestones, which are mostly prototype, some value engineering and some initial medical trials, aimed at skin cancer and dental work, will rest TeraView's second round of funding, he said.

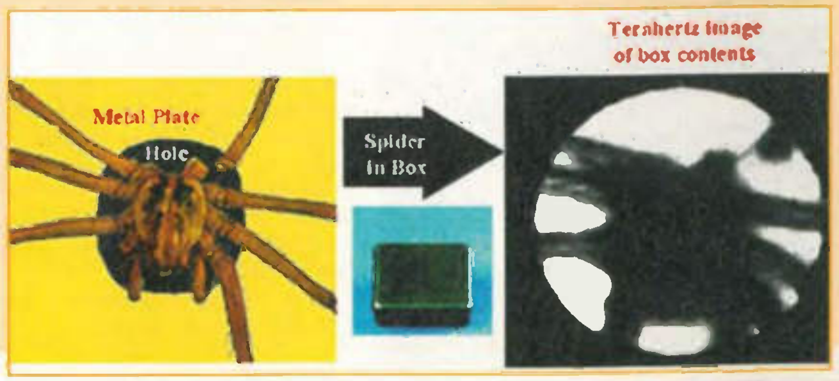
"Terahertz radiation has two remarkable properties," said Arnone. "The transparency of many common materials in this frequency band allows us to image a wide variety of items, including skin, teeth and other human tissue. And it can be used to identify the constituent materials through their absorption, which forms a characteristic fingerprint."

The company will be using terahertz pulse imaging which measures time-of-flight and received spectrum to produce position and composition data.

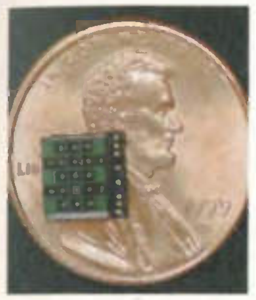
Toshiba is claiming several terahertz firsts. High-quality imaging of human tissue is one – which may led to a replacement for X-rays in certain circumstances. Three-dimensional images of objects constructed from scans and is another



A toothbrush handle grip up close – courtesy terahertz scanning.



Micromachined rocket technology



Micromachined thrusters may one day manoeuvre spacecraft and to prove it US firm TRW has fired one in a sub-orbital rocket test. "The test proves the technology behind this micro-thruster is well along in its development," said David Lewis, TRW's digital micro-propulsion project manager.

"We're very pleased with its performance and believe micro-thrusters have the potential to provide on-orbit propulsion for station keeping, orbital correction and attitude control for future very small satellites weighing from less than a pound to as much as 50 pounds [22kg]," said Lewis.

The thruster is an array of fire-once microthrusters, each the size of a poppy seed, made as a three-layer silicon-glass sandwich.

The lead-styphnate-containing propellant cells are in the middle layer. The front layer has a matching array of rupturable diaphragms and the back holds an igniter for each cell.

When ignited, each cell delivers a single impulse. The test array was fired more than 20 times at one-second intervals during the test. Each thruster delivered 10^{-4}Ns of impulse.

Mobile phone has video capability

Samsung has unveiled a mobile handset with video capability. The CDMA-standard phone transmits and receives at 144kbit/s and can decode MPEG4 video and MPEG2 audio. The two-inch TFT-LCD displays up to 200 000 colours.

DSP chip has multiple threading architecture

Imagination Technologies, the Hertfordshire-based design firm, has developed a multi-threaded digital signal processor architecture.

Dubbed Meta-1, the processor is the company's first foray into DSP and is believed to be the first such device to use simultaneous multi-threading (SMT).

"We always wanted to apply this technology to DSP," said Hossein Yassaie, Imagination's president and CEO.

SMT allows several tasks to run concurrently through one processor with no context switching overhead (see box).

"It's very much like having four virtual processors," said Yassaie of the Meta-1. The number of threads can vary between two and 16, he added.

In the Meta-1, a control system sets priorities for each of the threads to ensure they meet their real-time constraints.

As such it is ideal for real-time tasks such as audio and video processing.

A result of five year's work, Meta-1 can also mix signal processing tasks and general control functions, so avoiding the need for a separate microcontroller.

SMT also helps software development, as different tasks or threads can be written and tested by different teams and combined towards the end of the software design process.

Imagination, best known for its PowerVR graphics chip technology, has set up a division called Metagence Technologies to develop and market the Meta-1 processor. It is available for licensing now in applications such as MP3, digital hi-fi, speech recognition and networking.

The processor has already been designed into a digital audio chip set by Digital One, the UK's independent digital audio broadcaster. ■

Submarine mice

This is a computer mouse that works under water. The mouse was born out of the need to adapt software developed for traditional computing applications into the underwater environment. The marine mouse allows a diver to point and click on objects that are displayed on an underwater computer screen. It was developed at the Scripps Institution of Oceanography in the US and has already been used in a study to gauge the health of coral reefs.

The picture shows David Zawada, a graduate student at Scripps who is part of Dr Jules Jaffe's team that developed the new device. http://pandora.ucsd.edu

MOSFET power – pure and simple

David White's low-cost, high-quality, mosfet power amplifier delivers 100W or more depending on how you configure its output stage. An enhanced version of a popular topology, David's alternative offers 0.005% distortion at 1kHz and 1W.

Since the demise of the Maplin LP56 150W mosfet power amplifier kit I have had numerous requests to design a suitable replacement that retains the good performance and low cost of the original, but with none of its vices; poor power supply rejection and dependence on a fixed power-supply voltage.

Maplin's kit was based on an application note published by one of the major semiconductor manufacturers, as discussed below. My new power amplifier, described here, is an improved version of the venerable original design.

If you search the internet for audio power amplifier designs, variations of the circuit shown in **Fig. 1** crop up frequently. An unknown engineer working for Hitachi in the US[1] designed the original amplifier in 1977. Since then, it has spawned numerous commercial and DIY variants.

Remarkably, this amplifier topology still forms the basis of several commercial and DIY power amplifiers offered for sale today – among them products from BK Electronics, Marchand Electronics, and until very recently Maplin Electronics.

Tried and tested topology

Why has this design enjoyed such remarkable longevity? The distortion curves that appear in the Maplin literature, reproduced from Hitachi originals, are shown in **Fig. 2**. They promise outstanding performance from such a simple design.

Being something of a sceptic I initially dismissed these results as marketing puff. Eventually though, I got around to building a Maplin kit and hooking it up to my audio oscillator and spectrum analyser. The results I obtained showed that Hitachi's claims were fully justified; obviously this design merited further study.

There have been at least a couple of articles in *Wireless World* that have discussed the Hitachi design and/or improvements thereof[2,3]. One of the articles came close to including full constructional details, but some of the transistor type numbers were omitted[2].

Do transistor types matter in this design; can one not just use general purpose n-p-n and p-n-p types of sufficient voltage rating? The Hitachi design depends on high open-loop gain from the input and voltage amplifier stages so that global negative feedback can reduce the distortion produced by the mosfet output stage to insignificance. The transistors that Hitachi specify for the input and voltage amplifier long-tailed pairs are high voltage, high speed, high gain types with betas of around 400. They also have an enhanced TO92 package for increased power dissipation of 750mW in free air.

Transistors with these specifications are not common. Making substitutions invariably reduces the open-loop gain of the amplifier with a concomitant increase in distortion.

A further benefit of using high gain transistors for the input long-tailed pair is that the output offset voltage is reduced to a very low level; around 20mV in this design, with no trimming.

Can it be enhanced?

Maximising open-loop gain is obviously important then, but does the Hitachi design have any particular failings that are straightforward to remedy?

The biggest problem with the original design is that the input and voltage amplifier long-tailed pair collector currents depend on the supply voltage. This is inconvenient if you want to use power-supply voltages different from the original values. It also leads to poor power-supply rejection.

It is obvious from Fig. 1 that the collector currents of the input long-tailed pair are set by the positive supply voltage, +V, and the value of R_3. It is perhaps not quite so obvious that the collector currents of the voltage amplifier long-tailed pair are set by +V, –V and the value of R_9.

Although current mirror R_8, R_{11}, D_1 and Tr_5 ensures that the collector currents of Tr_3 and Tr_4 are equal to within 1-2%, their collector voltages are quite different. Negative feedback ensures that the collector voltage of Tr_4 is within $\frac{1}{2}V_{bias}$ of 0V. There are no constraints though on the collector voltage of Tr_3, which consequently sits at around 5V above the negative supply rail. Voltage across R_9 is therefore equal to the difference between +V and –V, less about 5V, so that the quiescent current in Tr_3 and Tr_4 can be set by choosing an appropriate value of R_9. Quiescent current is usually about 7mA.

A final point worth noticing from Fig. 1 is the use of a simple preset resistor, RV_1, for the bias generator; this is perfectly in order because the drain currents of the lateral mosfets used in the Hitachi design have a negative temperature coefficient above 100mA or so. As a result, they require no thermal compensation.

Lateral mosfets are specially designed for use in audio power amplifiers but they do have some disadvantages in the context of this design; the first of which is cost. Lateral mosfets with a power rating of around 100W cost in the region of £5 each when purchased in ones and twos. Vertical mosfets or hexfets with the same power rating, primarily intended for use in power switching circuits, cost around £1 each even in small quantities.

Further bonuses for vertical mosfets are that they usually have a higher transconductance than comparable lateral types. This leads to the possibility of improved linearity. They also have much lower values of $R_{DS(on)}$, which means a higher power output from any given design.

Performance of the 100W power amplifier

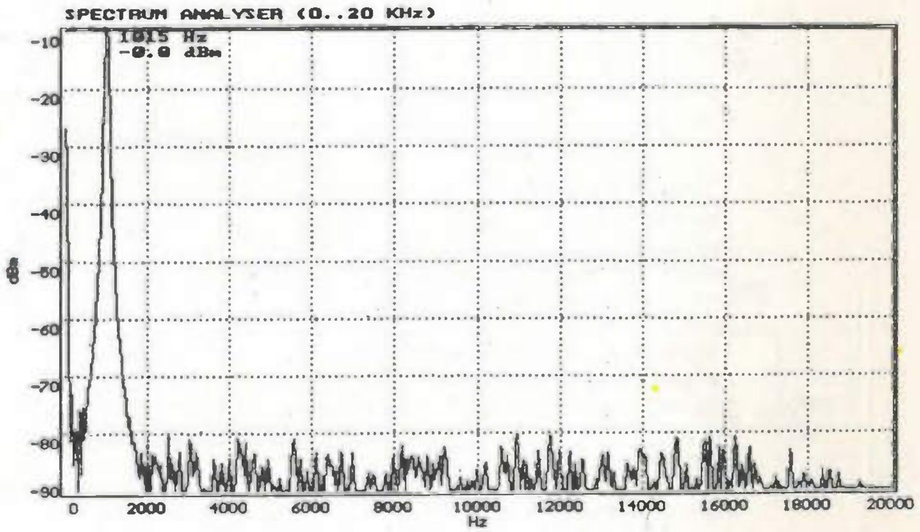
Performance figures for the power amplifier with ±50V supply rails are given below. If loads of less than the typical 6-8Ω are anticipated, it would be sensible to use at least two paralleled pairs of output devices.

Noise was measured with a floating input and unrestricted bandwidth, so the more usually quoted shorted input/bandwidth limited/weighted figure will be considerably lower than that given here.

Input sensitivity	1.28V rms – full output into 8Ω
Input impedance	47kΩ
Continuous output power	100W into 8Ω – 1kHz sinewave
Output offset voltage	20mV
Full power bandwidth	5Hz to 125kHz, –3dB
Slew rate	33V/µs
Noise	better than –80 dB unweighted
Total harmonic distortion	0.005% @ 1kHz, 1W/8Ω <0.01% @ 20Hz-20kHz, 1-100W/8Ω

At the expense of halving the input impedance, output offset voltage can be improved if R_2 is reduced to 22kΩ. The distortion spectrum for a 1W output into 8Ω at 1kHz (no bandwidth limiting) is shown below. It consists of the fundamental and noise, no harmonics of 1kHz are visible within the 16-bit resolution of the real-time spectrum analyser. The –85dB noise floor is due to the spectrum analyser; the power amplifier noise floor is much lower.

Unfortunately, the spectrum analyser doesn't have an averaging facility, but averaging by eye over a few seconds reveals a trace of second harmonic distortion consistent with the 0.003% indicated on the Hitachi/Maplin curves.



Distortion spectrum of the enhanced mosfet power amplifier for 1W output at 1kHz.

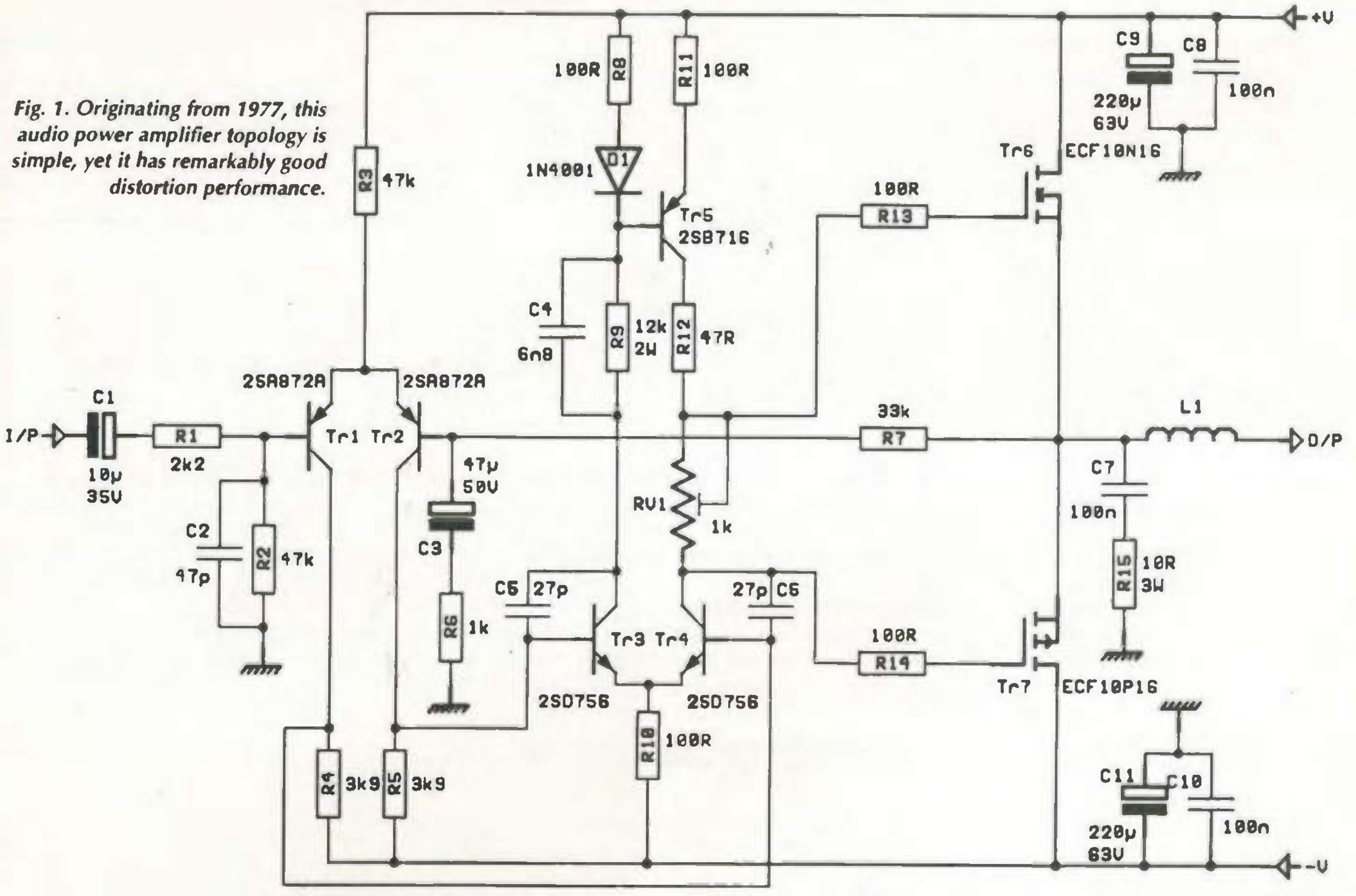


Fig. 1. Originating from 1977, this audio power amplifier topology is simple, yet it has remarkably good distortion performance.

Against this must be set typical gate thresholds of some 2-4V for vertical mosfets, compared to 0.5-1.5V for lateral types. This means a wider crossover region for vertical-fet-based class AB amplifiers.

Up to some 1-2A of drain current temperature coefficients are positive. This means that temperature compensation of the bias voltage is necessary.

My experiments have shown that vertical fets perform just as well as lateral fets in Hitachi-type designs, albeit at the small extra expense and complication of a V_{be} multiplier-based bias generator.

The new design

Figure 3 shows the new design, which is both lower in cost and higher in performance than the original. Tail current of the input pair is now set to approximately 1mA by means of the constant current source Tr_1. This transistor is fed from a 15V zener regulated power supply based around C_3, R_3, ZD_1.

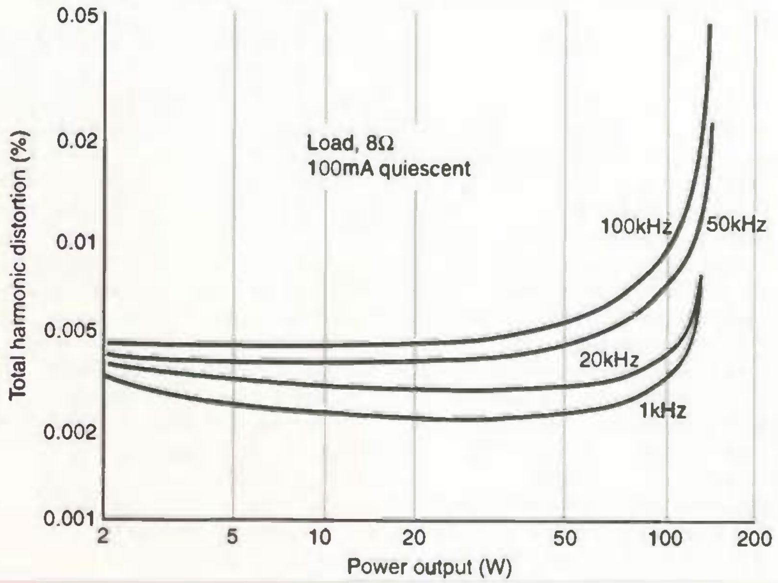


Fig. 2. Typical THD versus output characteristics for the amplifier topology of Fig. 1.

Using a constant-current source in this way makes the current of the input pair completely independent of the value of the positive supply voltage, +V. It leads to a considerable increase in power supply rejection ratio for the power amplifier as a whole.

A gate-source connected junction fet is not as good a constant current source as the conventional double transistor, or even the diode/transistor, circuit. It is adequate for the task at hand though, as well as having the merits of simplicity and low cost.

A 1mA constant-current diode – internally itself a selected gs-connected jfet – could be used in place of Tr_1 but these tend to be more expensive than straightforward jfets.

Collector currents of the voltage amplifier pair, $Tr_{9\&10}$, are no longer set by a simple resistor, as in the original design. Instead, constant current source R_{10}, R_{11}, LD_1, LD_2, Tr_6, Tr_7 set the collector currents.

These components comprise a complementary pair of diode/transistor constant current sources whereby each current source feeds the voltage reference (the led) of the other. There is approximately 1.6V across each led so that the constant current sources contribute

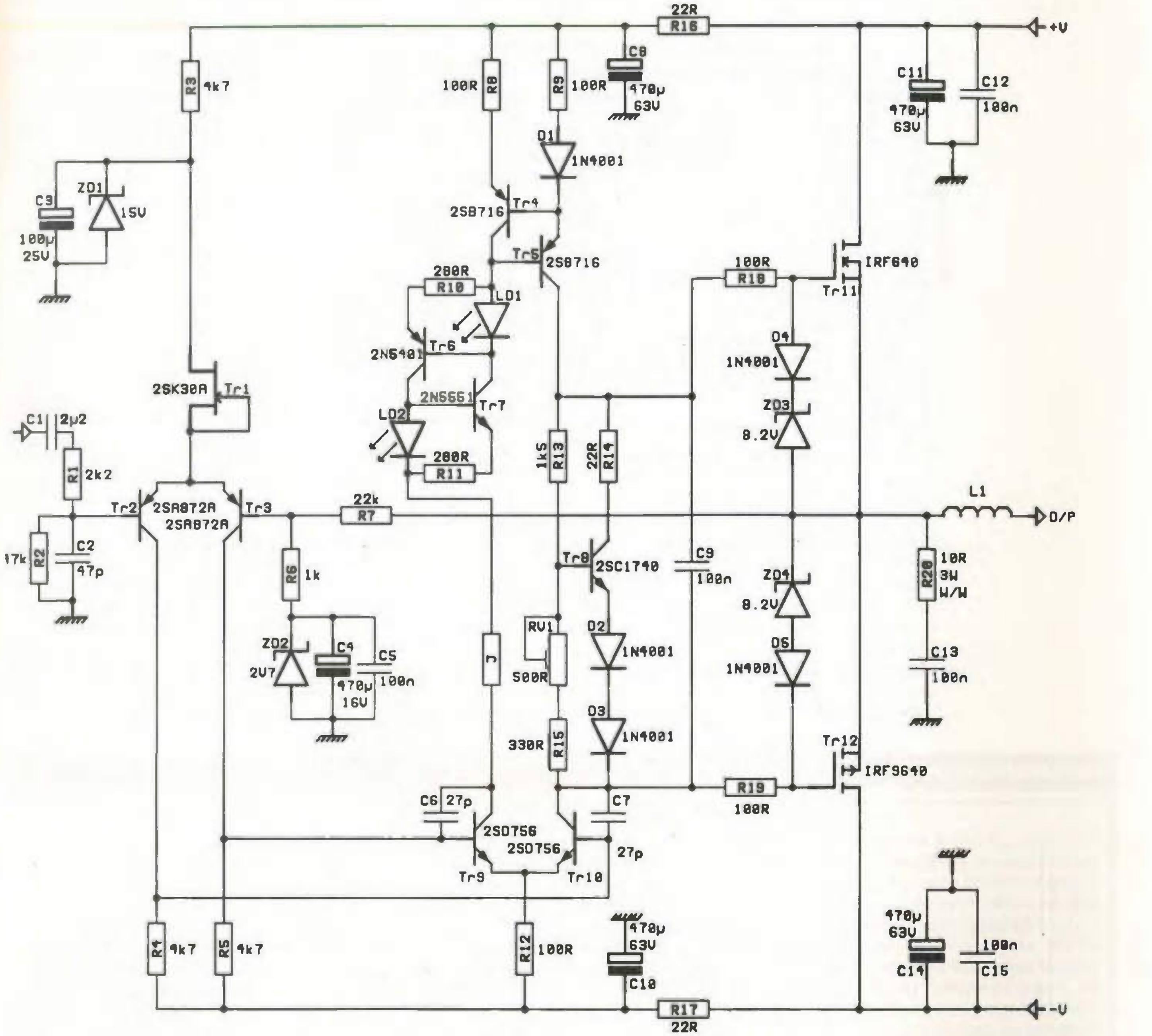


Fig. 3. An important enhancement of the 100W power amplifier is the use of constant-current loads. Among other things, these make the amplifier's performance less dependent on power supply voltage. Another important difference is the replacement of the lateral mosfets with vertical types – which are significantly cheaper.

$(1.6–0.6)/R_{10}$ and $(1.6–0.6)/R_{11}$ towards the total Tr_9 collector current of approximately 7mA.

It's not possible to use a simple gate-source-connected jfet as a constant-current source in this situation because low current devices with $V_{DS(max)}$ greater than 50V are almost impossible to obtain. Constant current diodes with 100V ratings are readily available but not at currents as high as 7mA.

The collector current of Tr_{10} is tied to the constant current source value of 7mA by the Wilson current mirror R_8, R_9, D_1, Tr_4, Tr_5. A Wilson current mirror provides a higher impedance collector load for the voltage-amplifier pair than the simpler current mirror in the original Hitachi design with a consequent increase in open loop gain.

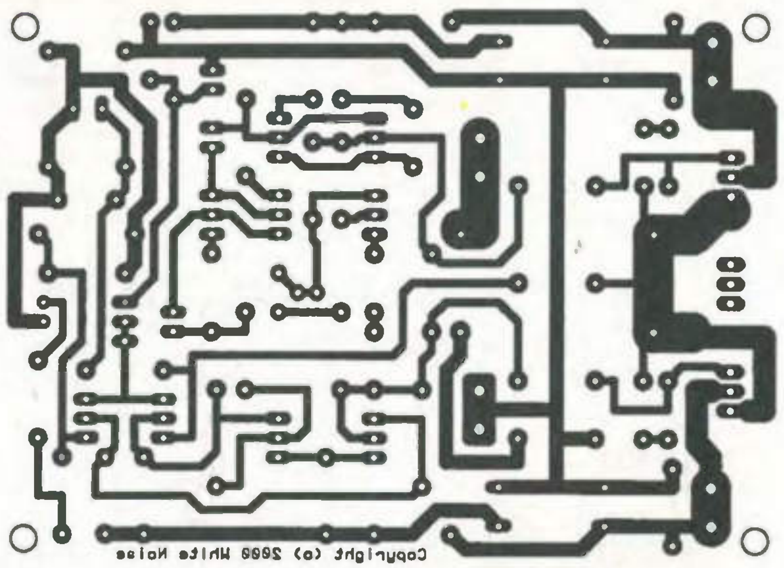
The high open loop gain, and associated low distortion when negative feedback is applied, will be reduced considerably if the voltage amplifier stage is loaded by the output stage to any great extent. That is why this circuit is much more effective with a simple, high input impedance, mosfet output stage than it would be when much more heavily loaded with a bjt output stage.

In order to provide temperature compensation for the output fets, a V_{be} multiplier, Tr_8, is used as the bias generator. This transistor must be fixed to the same heatsink as the output devices.

The positive temperature coefficient of drain current for vertical fets is smaller than that for bipolar transistors. Consequently, diodes D_2 and D_3 are included in the emitter of Tr_8. These diodes are located on the pcb rather than the output devices heatsink.

Because of these diodes, the transistor contributes only one third to the total temperature coefficient of the V_{be} multiplier, which is really a V_{be} plus two diode drop multiplier.

Alternatively, Tr_8 can be a small power vertical fet, such as an IRF630. In this case D_2 and D_3 are not necessary but the values of R_{13} and R_{15} will need to be adjusted appropriately.



Using a vertical fet for Tr_8 will give better thermal tracking of V_{bias}. Note that the bias voltage is now independent of supply voltage fluctuations to a much greater extent than in the original Hitachi design. This is because the collector load of Tr_{10} is effectively a constant current source.

The circuit of Fig. 3 will work equally well – albeit at somewhat higher cost – with lateral fets in place of vertical fets as Tr_{11} and Tr_{12}. Suitable changes will need to be made to the bias generator so that it produces 1-4V rather than 4-10V.

If lateral fets are used instead of vertical fets and the bias generator remains a V_{be} multiplier then Tr_8 must not be placed in thermal contact with the heatsink carrying the output devices.

Alternatively the whole bias generator circuit comprising R_{13}, R_{14}, R_{15}, RV_1, D_2, D_3, and Tr_8 may be replaced by a simple 500Ω or 1kΩ preset resistor. Suitable pairs of lateral fets are the Exicon ECX10N16/ECX10P16, Semelab/Magnatec BUZ900/BUZ905 and Hitachi 2SJ162/2SK1058.

A final pair of small embellishments involves R_{16}/C_8 and R_{17}/C_{10}. These provide a degree of decoupling for the input and voltage-amplifier stage power supplies from the output stage power supplies. Diodes D_4/ZD_3 with D_5/ZD_4 provide rudimentary – but nevertheless very effective – overcurrent protection for the output devices.

Modifications

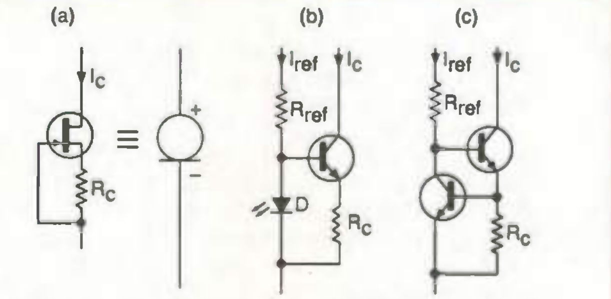
Power amplifier builders are inveterate tweakers. I have seen versions of the original Hitachi design with output powers up to 500W and beyond used for gigging and small-stadium rock. If you want to go this route with the present design here are a few guidelines.

Constant-current sources

The basic circuits of the three most commonly used constant-current sources are illustrated in the figure. All three have the constant-current programmed by a single resistor, R_C, and all may be embellished to some extent.

The simplest constant current source, a 'gate/source connected jfet' as shown in (a), uses only a junction fet and an optional resistor to provide a current anywhere up to I_{DSS} (R_C=0) for the fet. It has the merit of being a two-terminal device but the 'constant' current, I_C, will vary by around 5% between the extremes of applied voltage.

Junction fets are not readily available with $V_{DS(max)}$ greater than 50V, which limits the usefulness of this type of current source. However, integrated versions are available as constant-current diodes, in small signal diode or transistor-type packages, with ratings of up to 100V and a maximum current of 10-15mA.



Three way of implementing a constant-current source, (c) being the most constant over a given voltage range.

The diode/transistor type of constant current source, (b), is much more versatile, and the variation of current I_C with applied voltage is about five times smaller than that for the gate/source connected junction-fet type. The requirement to provide power for the voltage reference, D, means that this current source is a three terminal type. Diode D can be two or more small-signal diodes, a light-emitting diode (led), or a zener diode. Zener diodes are best avoided because they are noisy, and small signal diodes because the temperature coefficient of the current I_C will be greater than that for a current source constructed with a single led.

It is quite common to find R_{REF} in this circuit replaced by a constant-current diode; the constant current through D results in a more stable reference voltage. A red led typically has a forward drop of 1.6V when passing 3-5mA. This means that the voltage across R_C is typically 1V if V_{BE} for the transistor is assumed to be 0.6V. The programmed current is then simply $1/R_C$mA if R_C is expressed in kilo-ohms. Maximum voltage and current ratings are determined by the transistor chosen and the power dissipation capability of R_C.

The transistor/transistor constant current source in (c) shows current variations with applied voltage about five times smaller than those for the diode/transistor type. This makes it the best performer overall. It is also a three-terminal type, limited only by the type of transistors and resistor dissipation chosen.

The lower transistor robs the upper transistor of base current when the voltage across R_C exceeds 0.6V so that the programmed current is $0.6/R_C$ mA where R_C is in kilo-ohms. Unfortunately, complementary versions of this type of constant current source can't be connected back to back to produce a two terminal device in the same way as the diode/transistor type can.

It is possible to use paralleled output devices in order to obtain higher output power and/or reduced heat dissipation by individual mosfets. If you do this, source resistors of around 0.22Ω must be used with multiple vertical fets in order to ensure proper current sharing.

If at all possible vertical fets of the same polarity should also be matched for equal gate threshold voltages. Lateral fets are much more tolerant and may be paralleled without any special precautions.

When parallelling mosfets however, bear in mind that while mosfets have a DC input resistance of the order of teraohms, their input impedance at 10kHz is only around 20kΩ. Too many paralleled output devices will significantly load the voltage amplifier at high frequencies, leading to a reduction in gain and increased distortion.

If you want to use more than three or four pairs of output mosfets it's better to drive them from the voltage-amplifier stage via source followers running at a drain current of 15-20 mA.

For higher output powers, the supply voltages can be increased from the indicated ±40V up to ±60V. No changes to the power amplifier are needed, other than ensuring that the electrolytic capacitors have adequate voltage ratings, and that at least two pairs of output mosfets are used.

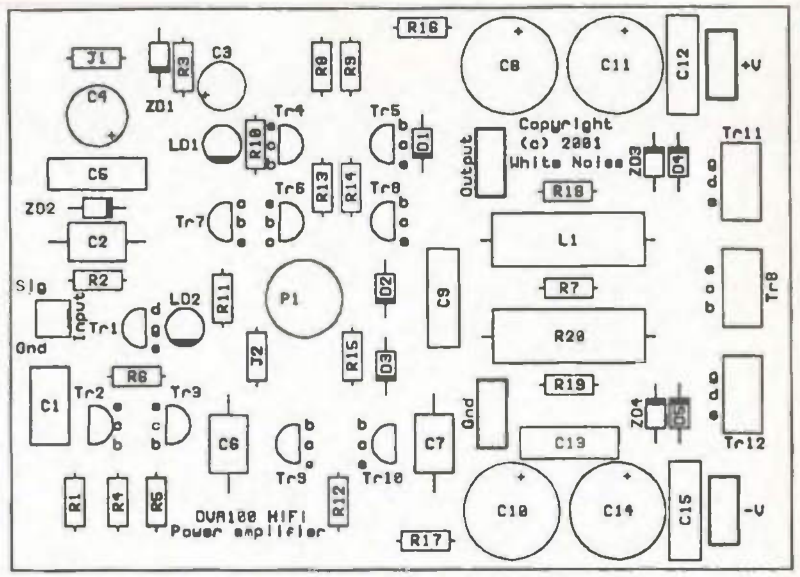
It might also be worthwhile to increase the constant current provided by Tr_6/Tr_7 from 7mA up to 10 or 12mA if more than two pairs of output devices are used without source followers. In this case, Tr_{4-7}, Tr_9, and Tr_{10} will require heatsinks or replacement by higher dissipation types.

Beyond ±60V Tr_{4-7}, Tr_9, and Tr_{10} will have to be substituted with higher voltage, higher dissipation, but invariably lower gain types; possibilities are TO126 devices such as the 2SA1209/2SC2911 or MJE340/MJE350 complementary pairs.

Its also possible to configure the power amplifier for lower power output by making Tr_{11} an IRF630, Tr_{12} an IRF9630, and reducing the power supply voltages to ±30V or so.

Implementing the design

Power amplifiers are best built on printed circuit boards. I have successfully built many a prototype on Veroboard though, before I taught myself how to expose and etch photosensitive pcb stock repeatedly



and reliably. This is much easier than most descriptions suggest, but you need to use the right materials.

I can provide professionally made, single sided, glass fibre pcbs. These are solder masked on both sides and silk screened with the component locations on the top side as detailed above.

Construction using a pcb is straightforward.

Setting up requirements

Before setup and testing, the output devices must be attached to, and insulated from, a heatsink of suitable thermal capacity.

Check for short circuits, and that RV_1 is rotated fully anticlockwise – i.e. at maximum resistance – before applying power to the circuit.

Adjust RV_1 until the quiescent current is approximately 100mA and leave the amplifier powered up for half an hour or so to attain thermal equilibrium; the quiescent current will usually fall a little during this period. If necessary increase the quiescent current back up to at least 80mA. Provided an adequately-rated heatsink is used, the quiescent current may be set to higher values than this if desired.

Upgrading the original design

If you already have a Maplin 150W power amplifier, it is possible to make the most worthwhile improvements detailed here on the original pcb. Resistor R_3, which is 47kΩ, can be replaced with a 1.0-1.5mA constant current diode or a gate-source connected jfet selected for 1.0-1.5mA I_{DSS}. A commonly-available 50V-rated constant-current diode, or a jfet rated at 50V $V_{DS(max)}$, will be adequate for use with the standard ±50V power supply, but a 100V constant-current diode will be required if your supply voltages are any higher.

If the constant-current source comprising R_{10}, R_{11}, LD_1, LD_2, Tr_6, Tr_7 in Fig. 3 is made up on a small piece of Veroboard, it can be soldered to the Maplin pcb in place of R_9. This constant-current source can be used with supply voltages of up to ±75V. Beyond this, the 2N5401 and 2N5551 will have to be replaced with higher V_{CE}-rated transistors.

Resistor R_3 can also be replaced with this type of constant current device, which is superior to a constant current diode or gate-source connected jfet as well as having a 150V rating. ■

References

1. Hitachi Mosfet Handbook, Section 6.1, "Typical Applications - Audio Power Amplifier Uses", pp143-151
2. Linsley Hood, J. "80-100W Mosfet Audio Amplifier", *Wireless World*, August 1982, pp 28-32.
3. Self, D. "Sound Mosfet Design", *Electronics World & Wireless World*, September 1990, pp760-764.

Ready-made PCBs

A stereo pair of printed circuit boards, as mentioned in the main article, may be obtained by sending a cheque for £35, drawn in favour of the author, to the following address, Dr David NJ White, 11 Station Road, Bearsden, Glasgow G61 4AW. The price includes shipping by recorded delivery.

Everything's going to Java

Embedded Java is beginning to take off as UK firms push to the front of development in that area. With the plethora of Java-hungry information devices on the market it looks a good bet, says STEVE BUSH

June 4 2001 was a good date to have put in your diary if you were thinking of using Java in an embedded application. That is the day this year when the JavaOne conference kicked off in San Francisco and Java processor companies were wanting to show something significant to gain credibility in the market.

Java processors are the key to embedded Java. Without them the associated Java virtual machine (JVM) is huge and clumsy, devouring more processor power and memory than an embedded system can hope to provide cost-effectively.

With them, Java starts to look quick on its feet and almost as svelte as assembler code.

The UK is in the forefront of embedded Java processor development with start-ups DCT and Vulcan Machines, as well as established ARM, all looking for a slice of the action.

Over in the US, Ajile Systems, inSilicon and Patriot Scientific are selling, or are about to sell Java processors or co-processors.

Obvious applications for Java processors are in mobile phones and information appliances, where they will handle Java Applets on the display and may go on to take on many behind-the-screen functions as the pool of Java intellectual property increases. Less obvious applications are simple devices that need to be connected in some way.

"Any networked device [could use Java], due to the networking classes in Java," said Larry Lorden, head of software tools at DCT. "Even if it only needs a very, very simple programme to send a packet." Such devices include, perhaps, a radio-connected remote rain gauge.

Vulcan's Rob MacAulay goes one further: Java should be used in "everything", he said. "Perhaps in the future washing machines and cookers – but I am a Java advocate."

Jini, the self-controlling Sun network, could be the key to widespread embedded Java.

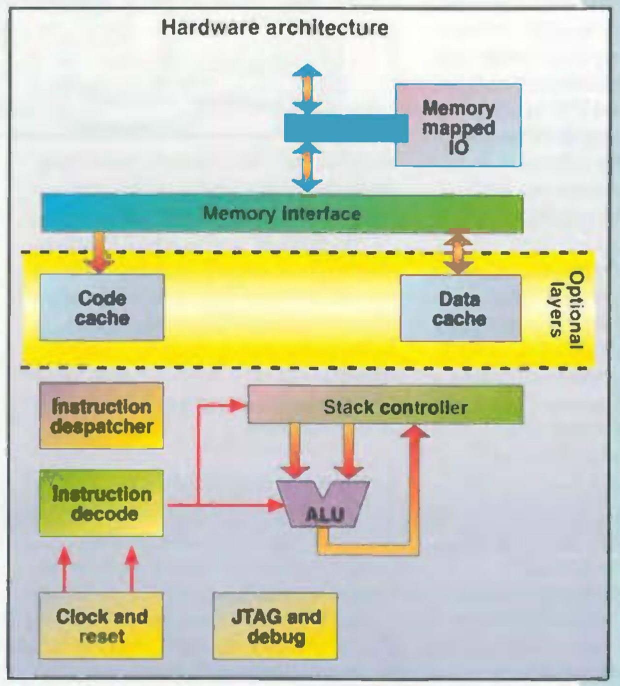
"One of the Holy Grails of Java is robust self-healing networks connecting devices, built on Jini," said DCT's Lorden.

Jini enables connected, say Bluetooth-connected, devices to network automatically when they come in range or are turned on.

Putting a Java processor inside a printer would be a cost-effective route to Jini printers without the need for proxy servers to intervene.

Unfortunately, according to Lorden, Jini is based on the Java RMI (remote methods invocator). Both Jini and the RMI were initially developed with desktop computer use in mind and are too big for embedded use.

Luckily, help is coming. "Sun is working on a lite version of Jini," said Lorden.



Royston-based Vulcan Machine's Java processor is called Moon.

Real-time Java is possible, but not yet an agreed reality. Virtual machines with proper real-time knowledge will come when the industry works out what is needed. "The hooks are already there," said Jon Howes of Java software company NeuW. "A concept of scheduling is built in." ■

Why Java?

Firstly, Java is relatively easy to write. True, there is a shortage of Java programmers, but: "Typically the shortage of Java programmers is for Web sites. We are talking about embedded developers," said Bob MacAulay, technical director at Vulcan Machines.

Also: "You can develop and test Java on a PC then download it onto your embedded device without having to do anything else," said Larry Lorden, at Java processor company DCT. "And Java is easy to maintain," he said. "A person who does Java can understand someone else's very easily. Sun developed Java so that it can only really be written in one style – unlike C which can be written in several and can be hard to follow if you are not familiar with all of them."

Java virtual machines also include proper exception handling, giving a system protection against stack over-runs and other 'features' that sometimes get written into code.

Designing radio receivers

In a set of four articles, Joe Carr looks at receiver design from the ground up. This first article outlines various radio receiver architectures.

The radio receiver is a device for receiving radio signals. How it does that job is different depending on the receiver architecture. That's what this article is all about. First, before covering receiver architectures I'll discuss a little about the basic problem of radio reception: separating signals from noise.

Signals, noise and reception

No matter how simple or fancy the system may be, the basic function of a radio receiver is the same: to distinguish signals from noise. The concept 'noise' covers both man-made and natural radio frequency signals. Man-made signals include all signals in the pass band other than the one being sought.

In communications systems, the signal is some form of modulated – AM, FM, PM, on-off telegraphy, etc. – periodic sine-wave propagating as an electromagnetic wave. The 'noise', on the other hand, may be a random signal that sounds like the hiss heard between stations on a radio.

The spectrum of such noise signals appears to be Gaussian, known as 'white noise', or pseudogaussian, in which case it is called 'pink noise' or 'bandwidth limited' noise. Alternatively, the noise may be man-made sparks or other electrical discharges, or even other signals that are not wanted.

In radio-astronomy systems the issue is complicated because the signals are also noise. The radio emissions of Jupiter and the Sun are very much like the signals that are, in other contexts, nothing but useless noise. In fact, in the early days of radar the galactic noise tended to mask returns from incoming enemy aircraft, so to the radar operators these signals were noise of the worst kind. Yet to a radio astronomer, those signals are the goal.

In satellite communications systems the 'signals' of the radio astronomer are limitations and annoyances at best and devastating at worst. The trick is to separate out the noise you want from the noise you don't.

Figure 1a) shows an amplitude-versus-time plot of a typical noise signal, while **Fig. 1b)** shows a type of regular radio signal that could be generated by a transmitter. Notice the difference between the two. The signal is regular and predictable. Once you know the frequency and period, you can predict the amplitude at other points along the time line.

The noise signal, on the other hand, is unpredictable. Knowing the cycle-to-cycle amplitude and duration – there is no true period – does not confer the ability to predict anything at all about the subsequent cycles.

In some receivers – especially those designed for pulse reception – the differences highlighted between Figs 1a) and 1b) are used to increase the performance of the receiver. An integrator circuit finds the time-average of the input signal.

True Gaussian noise integrated over a sufficiently long term will average to zero. This occurs because Gaussian noise contains all phases, amplitudes and polarities randomly distributed. Pseudogaussian noise is bandwidth limited, so may not integrate to zero, but very near it. The signal, on the other hand, will integrate to some non-zero value, so will stand out in the presence of integrated noise.

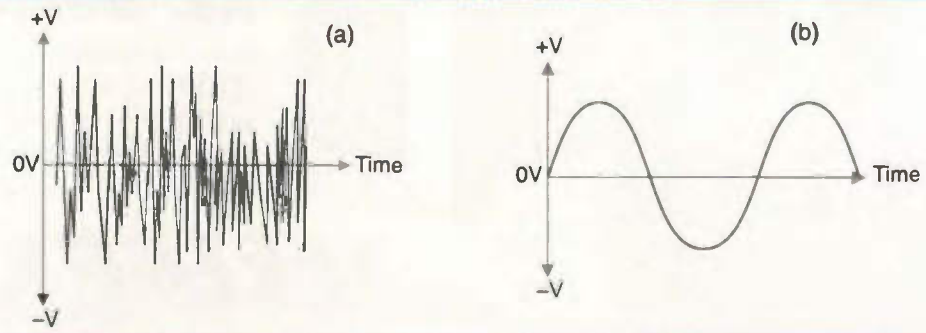


Fig. 1. In a) is an amplitude-versus-time plot of a typical noise signal while b) is a plot of a typical signal from a transmitter.

Thermal noise

Every electronic system – even a simple resistor – generates thermal noise, even if there is no current flowing through it. One of the goals of the system designer is to minimise the noise added by the system, so that weaker signals are not obscured.

One of the basic forms of noise seen in systems is thermal noise. Even if the amplifiers in the receiver add no additional noise – and they do – there will be thermal noise at the input due to the input resistance.

If you replace an antenna with a resistor that is totally shielded and matched to the system impedance, there will still be noise present. The noise is produced by the random motion of electrons inside the resistor. At all temperatures above absolute zero, which is about –273.16 °C, the electrons in the resistor material are in random motion. At any given instant there will be a huge number of electrons in motion in all directions.

The reason why there is no discernible current flow in one direction is that the electron motions cancel each other out even over short time periods. The noise power present in a resistor is:

$$P_N = KTBR \text{ watts} \qquad (1)$$

Here, P_N is the noise power in watts, T is the temperature in kelvins, K is Boltzmann's constant (1.38×10^{-23}joules/K) and R is the resistance in ohms. Note that by international agreement T is set to 290K.

Consider a receiver with a 1MHz bandwidth, and an input resistance of 50Ω. The noise power is,

$$(1.38\times10^{-23}\text{joules/K})\times290\text{K}\times1\,000\,000\text{Hz}\times50\Omega=2\times10^{-13}\text{W}$$

The reception problem

Figure 2 shows the basic problem of radio reception, especially in cases where the signal is very weak. The signal in Fig. **2a)** is embedded in noise that is relatively high amplitude. This signal is lower than the noise level, so is very difficult – perhaps impossible – to detect.

On the other hand, the signal in Fig. **2b)** is easily detectable because the signal amplitude is higher than the noise amplitude. It becomes difficult when the signal is only slightly stronger than the average noise power level.

The signal-to-noise ratio of a receiver system tells us something about the detectability of the signal. This ratio is normally quoted in decibels (dB), which is defined as:

$$SNR = 10\log\left[\frac{P_S}{P_N}\right]\text{dB} \qquad (2)$$

Here, SNR is the signal-to-noise ratio in decibels, P_S is the signal power level and P_N is the noise power level.

How high a signal-to-noise ratio is required? That depends on a lot of subjective factors when a human listener is present. Skilled radio operators can detect signals with a signal-to-noise ratio of less than a decibel, but the rest of us cannot even hear that signal. Most radio operators can detect 3dB signal-to-noise ratio signals, but for 'comfortable' listening a 10dB ratio is usually specified. For digital systems, noise performance is usually defined by the acceptable bit error rate, or BER.

Strategies

A number of strategies can be used to improve the signal-to-noise ratio of a system. First, of course, is to buy a receiver that has a low internal 'noise floor' and don't do anything to upset that figure. High-quality receivers have very low noise, but there is sometimes some creative specification writing in the advertisements. Different bandwidths are used for the measurements, and only the most favourable value – which may not be the bandwidth that matches your needs – is reported.

Common sense indicates that there are two approaches to signal-to-noise ratio improvement: either increase the signal amplitude or decrease the noise amplitude. Most successful systems do both, but the enhancements must be done carefully.

One approach to signal-to-noise ratio improvement is to use a preamplifier ahead of the receiver antenna terminals. This approach may or may not work, and under some situations may make the situation worse. The problem is that the preamplifier adds noise of its own, and will amplify noise from outside – received through the antenna – and the desired signal equally.

If you have an amplifier with a gain of, say, 20dB, then the external noise is increased by 20dB and the signal is increased by 20dB. The result is that the absolute numbers are bigger but the signal-to-noise ratio is the same. If the amplifier produces any significant noise of its own, then the signal-to-noise ratio will degrade. The key is to use a very low noise amplifier for the preamplifier. Using a low noise amplifier for the preamplifier may actually reduce the noise figure of the receiver system.

Another trick is to use a preselector ahead of the receiver. A preselector is either a tuned circuit or bandpass filter. It is placed in the antenna transmission line ahead of the receiver antenna terminals. A passive preselector has no amplification – it uses LC elements only – while an active preselector has a built-in amplifier.

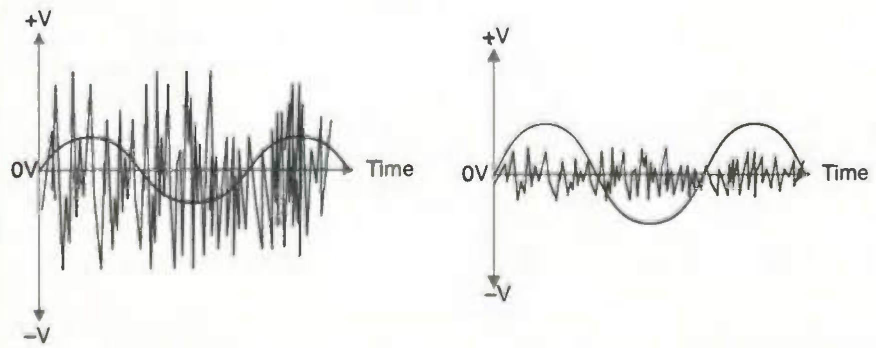


Fig. 2. In a), the wanted periodic antenna is buried in noise and may not be recoverable. Where the signal amplitude exceeds the noise amplitude, as in b), recovering the signal becomes a much easier job.

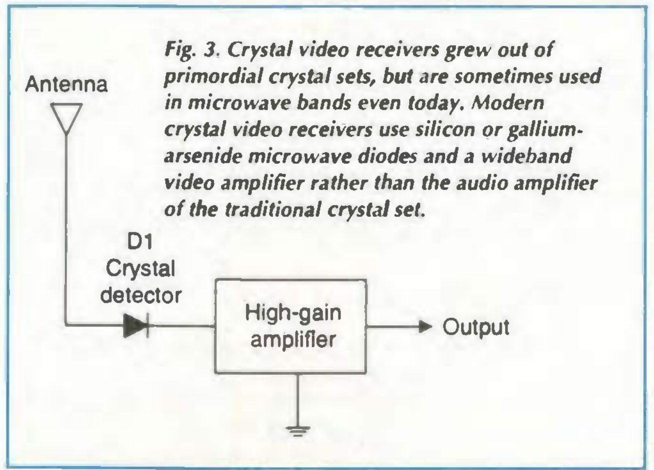


Fig. 3. Crystal video receivers grew out of primordial crystal sets, but are sometimes used in microwave bands even today. Modern crystal video receivers use silicon or gallium-arsenide microwave diodes and a wideband video amplifier rather than the audio amplifier of the traditional crystal set.

The amplifier should be a low-noise type. A preselector can improve the system because it amplifies the signal by a fixed amount, but only the noise within the pass band is amplified by the same amount as the signal. Improvement comes from bandwidth limiting the noise but not the signal.

Another practical approach is to use a directional antenna. This method works especially well when the unwanted noise is other man-made signal sources.

An omnidirectional antenna receives equally well in all directions. As a result, both natural and man-made external noise sources operating within the receiver's pass band will be picked up. But if the antenna is made highly directional, then all noise sources that are not in the direction of interest are suppressed.

Highly directional antennas have gain, so the signal levels in the direction of interest are increased. Although the noise also increases in that direction, the rest of the noise sources (in other directions) are suppressed. The result is that signal-to-noise ratio is increased by both methods.

When designing a communications system, the greatest attention should usually be paid to the antenna, then to an low-noise amplifier or preselector, and then to the receiver. Generally speaking, money spent on the antenna gives more signal-to-noise ratio for a given investment than the same money spent on amplifiers and other attachments.

Radio receiver architectures

Radio receivers are at the heart of nearly all communications activities. In this section I will outline the different types of radio receivers that are on the market. I will be showing how to interpret receiver specifications in a subsequent article.

Origins. The very earliest radio receivers were not receivers as we know them today.

Early experiments by Hertz, Marconi and others used spark gaps and regular telegraph instruments of the day. Range was severely limited because those devices have a terribly low sensitivity to radio waves.

Later, around the turn of the 20th century, a device called a Branly coherer was used for radio signal detection. This device consisted of a glass tube filled with iron filings placed in series between the antenna and ground. Although considerably better than earlier apparatus, the coherer was something of a dud for weak signal reception.

In the first decade of this century, however, Fleming invented the diode vacuum tube, and Lee DeForest invented the triode vacuum tube. The latter device made amplification possible and detection a lot more efficient.

A receiver must perform two basic functions. It must respond to, detect and demodulate desired signals; and secondly, it must not respond to, detect, or be adversely affected by undesired signals. If it fails in either of these two functions, then it is a poorly performing design.

Both functions are necessary. Weakness in either function makes a receiver a poor bargain unless there is some mitigating circumstance. The receiver's performance specifications tell us how well the manufacturer claims that their product does these two functions.

Crystal video receivers. Crystal video receivers, **Fig. 3**, grew out of primordial crystal sets, but are sometimes used in microwave bands even today. The original crystal sets – pre-1920 – used a naturally occurring PN junction 'diode' made from a natural lead compound called galena crystal. They also involved an inductor-capacitor tuned circuit.

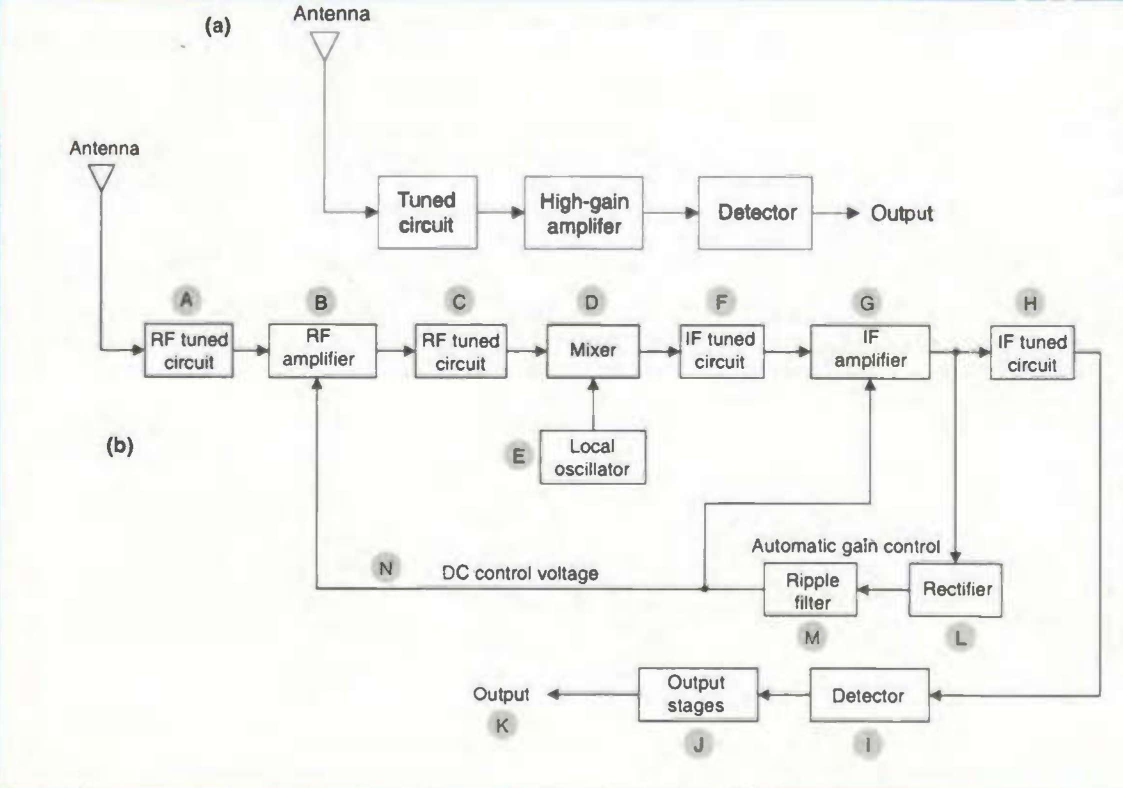


Fig. 4. Two types of tuned radio frequency receiver. The TRF receiver has an LC resonant circuit in its front-end, followed by one or more radio frequency amplifiers ahead of a detector stage. Fig. 4a) is a tuned gain-block receiver and is commonly used in monitoring very low frequency signals. Later versions use multiple tuned RF circuits between the amplifier stages and have mechanically linked tuned circuits, Fig. 4b).

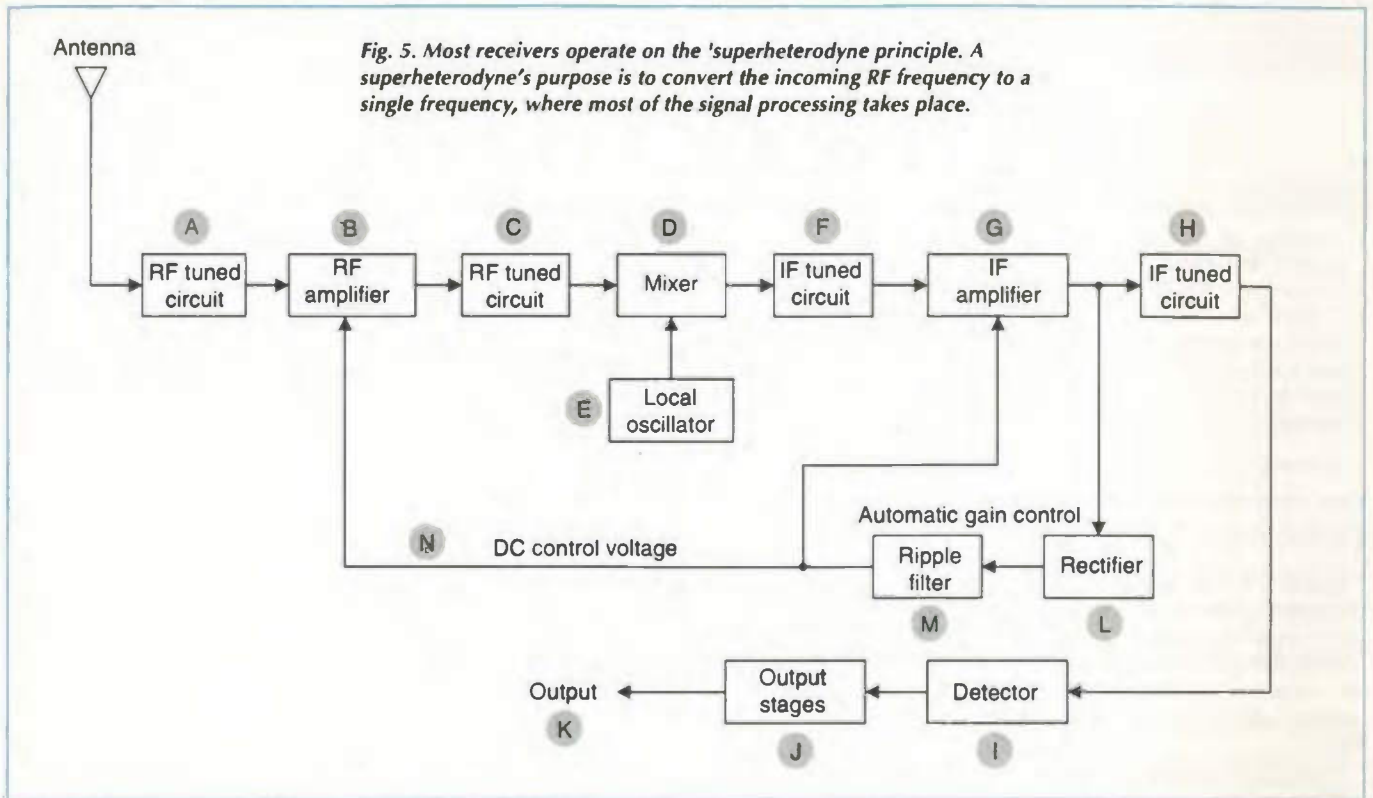


Fig. 5. Most receivers operate on the 'superheterodyne principle. A superheterodyne's purpose is to convert the incoming RF frequency to a single frequency, where most of the signal processing takes place.

Later, crystal sets were made using germanium or silicon diodes. When vacuum tubes became generally available, it was common to place an audio amplifier at the output of the crystal set.

Modern crystal video receivers use silicon or gallium-arsenide microwave diodes and a wideband video amplifier rather than the audio amplifier. Applications include some speed radar receivers, aircraft warning receivers, and some communications receivers – especially short-range.

Tuned radio frequency receivers. The tuned radio frequency, or TRF, radio receiver has an *LC* resonant circuit in its front-end, followed by one or more radio frequency amplifiers ahead of a detector stage. Two varieties are shown in **Fig. 4.**

The version in Fig. **4a)** is called a tuned gain-block receiver. It is commonly used in monitoring very low frequency VLF signals to detect solar flares and sudden ionospheric disturbances, or SIDs.

Later versions of the TRF concept use multiple tuned RF circuits between the amplifier stages. Early models used independently tuned *LC* circuits, but those proved to be very difficult to tune without creating an impromptu Miller oscillator circuit. Later versions mechanically linked – i.e. 'ganged' – the tuned circuits to operate from a single tuning knob.

Superheterodyne receivers. Figure 5 shows the block diagram of a superheterodyne receiver. I will use this hypothetical receiver as the basic generic framework for evaluating receiver performance.

The design in Fig. 5 is called a superheterodyne receiver. It represents the largest class of radio receivers; it covers the vast majority of receivers on the market.

The superheterodyne receiver block diagram of Fig. 5 is typical of many receivers. A superheterodyne's purpose is to convert the incoming RF frequency to a single frequency, where most of the signal processing takes place.

At the front-end section of the receiver are the radio frequency (RF) amplifier and any RF tuning circuits that may be used (A-B-C in Fig. 5). In some cases, the RF tuning is very narrow, and basically tunes one frequency. In other cases, the RF front-end tuning is broad banded. In those cases, bandpass filters are used.

Frequency-translator sections D and E in Fig. 5 are also considered part of the front-end in most textbooks, but here I have labelled them as a separate entity. The translator consists of a frequency mixer and a local oscillator. Output from the frequency translator is called the intermediate frequency, or IF.

The translator stage is followed by the intermediate frequency amplifier. The IF amplifier, sections F-G-H in Fig. 5, is basically a radio-frequency amplifier tuned to a single frequency. The IF can be higher or lower than the RF frequency, but it will always be a single frequency.

A sample of the IF amplifier output signal is applied to an automatic gain control, or AGC, section, blocks L to M in Fig. 5. The purpose of this section is to keep the signal level in the output more or less constant.

The AGC circuit consists of a rectifier and ripple filter that produces a DC control voltage. The DC control voltage is proportional to the input RF signal level, *N*. It is applied to the IF and RF amplifiers to raise or lower the gain according to signal level. If the signal is weak, then the gain is forced higher, and if the signal is strong the gain is lowered. The end result is to smooth out variations of the output signal level.

The detector stage, I in Fig. 5, is used to recover any modulation that is on the input RF signal. The type of detector depends on the type of modulation used for the incoming signal.

Amplitude-modulation, or AM, signals are generally handled in an envelope detector. In some cases a special variant of the envelope detector called a square-law detector is used. The difference is that the straight envelope detector is linear, while the square-law detector is nonlinear. Single sideband (SSB), double sideband suppressed carrier (DSBSC), and keyed CW signals will use a product detector, while FM and PM need a frequency or phase-sensitive detector.

The output stages, blocks J-K in Fig. 5, are used to amplify and deliver the recovered modulation to the user. If the receiver is for broadcast use, then the output stages are audio amplifiers and loudspeakers.

In some radio astronomy and instrumentation telemetry receivers the output stages consist of integrator circuits and DC amplifiers.

Heterodyning

The main attribute of the superheterodyne receiver is that it converts the radio signal's RF to a standard frequency for further processing.

Today, the new frequency, called the intermediate frequency or IF, may be either higher or lower than the RF. Early superheterodyne receivers always down-converted the RF signal to a lower intermediate frequency. The reason was purely practical. In those days, higher frequencies were more difficult to process than lower frequencies.

Even today, because variable tuned circuits still tend to offer different performance over the band being tuned, converting to a single IF frequency, and obtaining most of the gain and selectivity functions at the IF, allows more uniform overall performance over the entire range being tuned.

A superheterodyne receiver works by frequency converting – i.e. heterodyning – the RF signal. In the term 'superheterodyne', the prefix 'super' is the result of 1920s vintage advertising hype. Heterodyning occurs by nonlinearly mixing the incoming RF signal with a local oscillator, or LO, signal. When this process is done, disregarding noise, the output spectrum will contain a large variety of signals according to,

$$F_o = mF_{RF} \pm nF_{LO} \qquad (3)$$

Here, F_{RF} is the frequency of the RF signal, F_{LO} is the frequency of the local oscillator and m and n are either zero or integers (0, 1, 2, 3... n).

Equation 3 means that there will be a large number of signals at the output of the mixer. For the most part though, the only ones that are of immediate concern to understanding superheterodyne operation are those for which m and n are either 0 or 1. Thus, for our present purpose, the output of the mixer will be the fundamentals F_{RF} and F_{LO}, and the second-order products $F_{LO}-F_{RF}$ and $F_{LO}+F_{RF}$, Fig. 6.

Some mixers, notably those described as double-balanced mixers (DBM), suppress F_{RF} and F_{LO} in the mixer output, so only the second-order sum and difference frequencies exist with any appreciable amplitude. This case is simplistic, and is used only for this present discussion. Later on, we will look at what happens when third-order (2F1 ± F2 and 2F2 ± F1) and fifth-order (3F1 ± 2F2 and 3F2 ± 2F1) become large.

Note that the local-oscillator frequency can be either higher than the radio frequency – in which case it is termed high-side injection – or lower than the RF, which is low-side injection. There is ordinarily no practical reason to prefer one over the other, except that it will make a difference whether an analogue main tuning dial reads high-to-low or low-to-high – assuming one is used.

The candidates for IF are the sum LO+RF and difference LO–RF second-order products found at the output of the mixer. A high-Q tuned circuit following the mixer will select which of the two are used.

Consider an example. Suppose an AM broadcast band superheterodyne radio has an IF frequency of 455kHz, and the tuning range is 540 to 1700kHz. Because the IF is lower than any frequency within the tuning range, it will be the difference frequency that is selected for the IF. The local oscillator is set to be high-side injection, so will tune from (540+455)=995kHz, to (1700+455)=2155kHz. ■

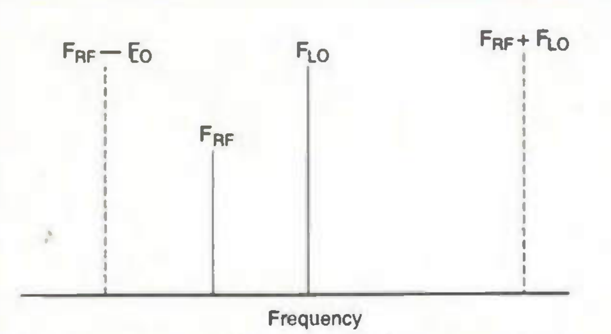


Fig. 6. Frequency F_{RF} is the incoming radio-frequency signal while F_{LO} is the local-oscillator frequency of a superheterodyne receiver. Second-order products F_{RF}–F_{LO} and F_{RF}+F_{LO} result when these two are mixed.

In the next article on receiver design, Joe deals with the various circuits inside a superheterodyne receiver.

Measuring RF millivolts

When measuring around 220mV, Cyril Bateman's simple RF millivoltmeter reads less than 7mV – or 0.25dB – down at 100MHz and remains useful at up to 500MHz. It works in conjunction with a standard DMM and involves just two ICs – or one if you're lucky.

Many oscilloscopes measure to at least 100MHz but most AC voltmeters work only at much lower frequencies. In addition, when measuring non-sinusoidal waveforms a method for measuring true RMS becomes essential.

Assuming a known waveform, an oscilloscope could be used to estimate RMS voltage, but for most waveforms an RMS responding meter must be used.

Examination of the various techniques used to measure AC voltages show that most methods have limited frequency range. Average responding millivoltmeters using rectifying techniques[1] can be used up to perhaps 10MHz. However, most RMS-responding meters, based on ICs using the 'implicit' solution method[2] are limited usually to 1 or 2MHz.

In the March 1999 issue 'Hands-on Internet' column[3] I discussed using the LT1088 to measure RMS voltages. As with the classic calorimetric method, this IC provides two identical resistive heating elements and temperature sensing diodes. By comparing the heating effects of the unknown voltage with that of a known adjustable DC voltage, the true RMS value of the unknown voltage is derived.

In the October 2000 issue 'Hands-on Internet' article, I discussed using two AD834 multiplying ICs to measure true RMS to 500MHz. This method used one multiplier as a mean squared detector of the input voltage, the second to produce the square root of the average of this mean squared voltage.

Other possible approaches – especially for low crest factor signals – can be based on using logarithmic detector ICs.

For some time I remained undecided as to which of these approaches to use in a new meter. The calorimetric method looked attractive and has been used to measure up to 100MHz, but it needs care to avoid damage to the sensors. The AD834 approach promised a higher operating frequency, but I doubted whether I had sufficient resources to attain this performance.

A new IC for instrumentation

In the event, an article in *Microwaves and RF*[4], about the AD8361 TruPwr, a new IC from Analog Devices, settled my debate. Designed by Barrie Gilbert this minute low-cost IC claims instrument-grade performance and accuracy up to 2.5GHz. Needing only a single rail 2.7 to 5.5V supply at 5mA, it is easily battery powered, **Fig. 1**.

The AD8361 follows the two AD834 multiplier approach to extract the RMS equivalent from any waveform. It comprises two identical squaring circuits, an error amplifier and output buffer all within its 8-lead micro_SOIC surface mount package, **Fig. 2**.

Intended for use in 50Ω measurements of voltage or power, it needs only a suitable input matching network, a filter and decoupling capacitors to convert its RMS input to the corresponding DC voltage. A small instrumentation amplifier can be used to adjust offset voltages and buffer the AD8361 output. **Fig. 3**.

These circuits all fit easily into a matchbox-sized printed board. **Fig. 4**.

With an input impedance ranging from 225Ω and 1.2pF at low frequency to some 80Ω and 1pF at 2.5GHz, relatively simple input matching circuits, described in the data sheet, can achieve an input VSWR of 1.5:1 in a 50Ω system.

While a VSWR of 1.5:1 is usual for high frequency voltage measuring instruments, any deviation from the ideal 1:1 implies a voltage or power measurement error.

Design target

My ideal objective was for a 0 to 1V capability and flat response to 100MHz, with –3dB at 1GHz and 1.1:1 or better matching up to 100MHz. This would allow my meter to accurately measure RMS volts, as well as power, in a 50Ω system.

To achieve these targets, the input matching and output scaling suggestions in the AD8361 data sheet would not suffice. Dedicated input and output networks were needed. While I could use the performance data for this IC in simulations, the capacitors, resistors and PCB strays posed considerable difficulty.

Above 100MHz, very small capacitors can be used. For input blocking, a 100pF COG will suffice, and for the averaging filter circuit, 1nF. The demonstration board provides 0402 and 0603 size components. Such near-ideal components simplify the design task,

At low frequency, much larger capacitance values are needed, so these small capacitors cannot be used. To measure down to 1kHz the AD8361 requires a DC blocking input coupling capacitor of 1μF to ensure –3dB, and 10μF for a flat response. Although a 4.7μF Y5V capacitor in 1206 size is available from some suppliers, even this would not guaran-

RMS measurements

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

Now Barrie 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 with up to 30dB dynamic range. This new technique can measure high crest-factor waveforms with an accuracy similar to a thermal bridge.

Accepting the thermal bridge technique as reference, the AD8361 provides a wider dynamic range, faster response and much smaller board volume. As for accuracy, measuring single channel CDMA it provides +0.2dB, for multi channel W-CDMA it reads +0.8dB, **Fig. A.**

This compares favourably against diode detectors at +2dB and more than +5dB, while the logarithmic amplifier approach provides +3.55dB and more than +5dB for the same waveforms.

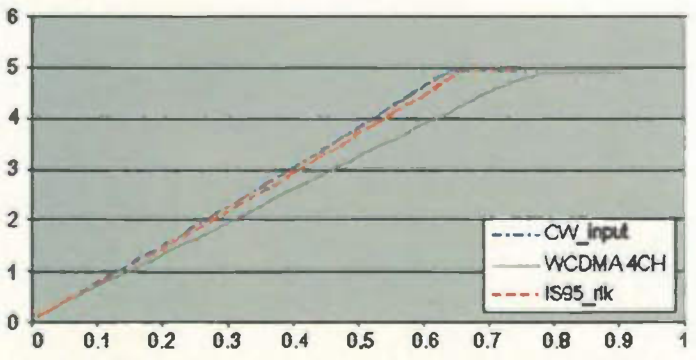


Fig. A. This plot, taken from an Analog Devices press release, shows measurements of three common signals. The vertical 'Y axis' represents the output from the AD8361. The horizontal 'X axis' plots the response from the thermal bridge, namely a Boonton model 51101.

RMS-to-DC converter chip

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 – and up to 2.5GHz.

This chip provides three operating modes – ground referenced, supply referenced or using its internal 350mV bandgap reference. It 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.

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, Fig. 2.

The second squaring cell closes 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 this 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. Take care though not to exceed the maximum permitted supply of 5.5V.

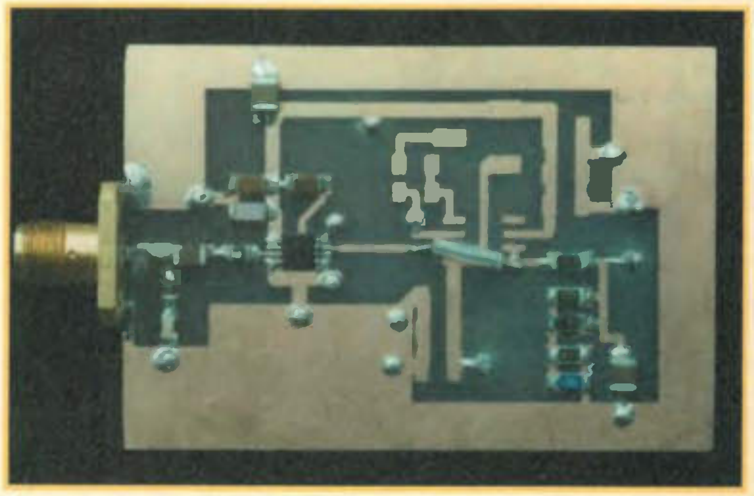


Fig. 1. Using the AD8361 to accurately measure true RMS to 1V, from 10kHz to 100MHz. The left-most edge of the PCB has been removed to accept an SMA stripline connector. This AD8361 had a small offset so IC_2 and associated components (Fig. 7) could be omitted.

Table 1. Performance figures, showing VSWR and 0dBm voltage error by frequency. Note that at 50Ω impedance, 0dBm is 223.6mV.

Frequency	Return loss	VSWR	Millivolts	Relative 0dBm
1kHz			195.8	–1.15dB
10kHz			222.8	–0.03dB
100kHz	–29.4 dB	1.07:1	223.2	–0.016dB
1MHz	–31.4 dB	1.055:1	223.2	–0.016dB
15MHz	–31.8 dB	1.053:1	222.1	–0.06dB
25MHz	–31.8 dB	1.053:1	221.8	–0.07dB
50MHz	–31.3 dB	1.056:1	218.7	–0.192dB
100MHz	–30.5 dB	1.061:1	217.3	–0.248dB
200MHz			211.8	–0.47dB
250MHz			203.2	–0.83dB
300MHz			195	–1.19dB
400MHz			175.1	–2.12dB
500MHz			157.7	–3.03dB

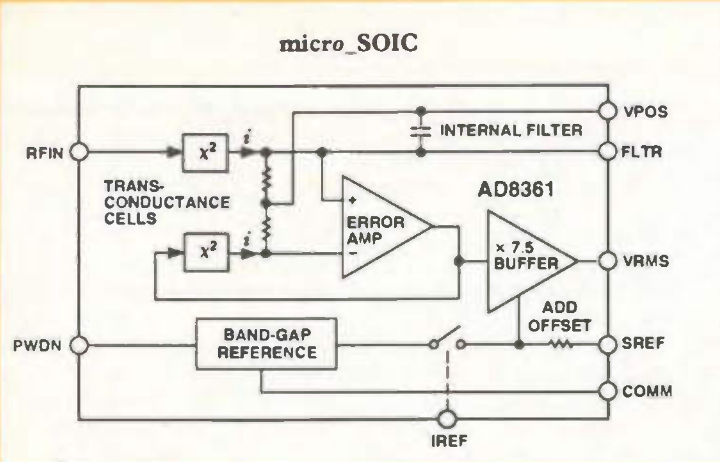


Fig. 2. Block diagram of the TruPwr AD8361 IC shows how the essential elements of two high-frequency multipliers, error and output amplifiers have been combined into one minute micro-SOIC package.

tee a flat response down to 1kHz.

Capacitor problems

Such large components naturally introduce considerable difficulty at high frequencies. Most capacitors exhibit multiple resonances. The most common, since it occurs at the lowest frequency, is a series self resonance. Depending on physical size and dielectric materials, parallel resonances then occur at higher frequencies.

Any parallel resonance of the DC blocking capacitor in the desired pass band would introduce major measurement errors. For these reasons I chose to use a 1μF capacitor in Z5U dielectric and case size 1206, restricting my flat response to 10kHz.

A 1μF 1206 capacitor in Z5U dielectric, exhibits four main problems at high frequencies:

- A series self-resonance frequency of around 3.5MHz.
- Considerable loss of capacitance with frequency.
- Significant frequency-dependent ESR, relative to the AD8361 input impedance.
- Notable self inductance.

These factors inhibit simple analysis of the input matching circuit using a Smith chart or conventional 'Spice' simulation, since both methods assume ideal capacitor behaviour.

For most of the frequency band, this 1μF capacitor would behave as a DC blocking inductor, so I first needed to establish its series self-resonant frequency. Mounting the 1μF Z5U 1206 size capacitor on my test jig[5] confirmed its self-resonant frequency as 3.647MHz and self inductance as 1.5nH. A few impedance calculations[6] also established its reduction of capacitance with frequency.

These values were converted into frequency-dependent equations and input as 'freq' or 'frequency' parameters, to better describe this capacitor for frequency domain simulations using the Microcap MC6 simulator.

Similar equations for the AD8361 were derived from Fig. 13 in the device's data sheet. These were also used as 'frequency' parameters to allow MC6 to better model the AD8361 actual input impedance by frequency, Fig. 5.

The maximum linear input voltage of the AD8361 using a 5V supply is 0.66 volts. An additional resistor was placed in series with the AD8361 input to ensure that its input voltage did not exceed this linear range when subjected to my desired 0-1V signal.

Following a number of simulation runs, I chose to place additional inductance in series with the shunt resistor R_1 to flatten the circuit's input impedance at higher frequencies. This added inductance was chosen to counterbalance the inductances of the circuit board tracks, the 1μF capacitor and R_2.

After many simulation attempts, I was left with an input resonance near 1GHz. This frequency being outside my facilities for practical impedance or VSWR measurement, I decided for safety's sake to reduce my upper frequency –3dB target to 500MHz. Input to the AD8361 was deliberately rolled off at higher frequencies, to avoid this resonance. Fig. 6.

Voltage standing-wave ratio

When a signal generator having an output impedance of precisely 50Ω is connected to an exact 50Ω load, all the signal from the generator is absorbed into the load. However if connected to a non 50Ω load, then less power is absorbed and some signal is returned to be dissipated in the output impedance of the generator.

Obviously for accurate measurements of voltage or power in a 50Ω system, the meter should replicate this ideal 50Ω load.

Measurement of the level of this 'returned' signal, can be used to ascertain the impedance of a load. While many techniques may be used, the most convenient method for frequencies up to 100MHz is to use a 'reflection' bridge.

A reflection bridge is similar to the familiar resistive Wheatstone bridge, but it uses a balun transformer in place of the detector. This balun allows the bridge's floating unbalance voltage to be measured simply, using unbalanced 50Ω meters and coaxial connectors.

When a non-50Ω load is connected to the load port, the bridge acts to separate the signal returned from the load so that it is directed to the return port and not returned to the signal generator.

This returned signal is measured as being smaller than that returned when applying either open circuit or short circuit loads to the load port. It can be measured as a proportion of the open/short signal, in which case it is called 'reflection coefficient', or more usually in decibels as (–20 log reflection coefficient), when it is called 'return loss'.

A good reflection bridge loaded with a precision 50Ω load can separate the forward and return signals by at least 40dB. In other words, with a precision load, the return signal will be less than 1% of the forward signal. The very best precision loads and bridges can produce a return loss around –50dB or 0.316%.

Return loss, reflection coefficient and VSWR are simply ways to indicate the same deviation of load impedance from the nominal 50Ω.

$$VSWR = \frac{(1 + \text{reflection coefficient})}{(1 - \text{reflection coefficient})}$$

Unfortunately the simple case of a load connected directly to a generator is rare. In practice various cables and connectors may be used. Each cable and connector introduces an impedance mismatch though. Although each mismatch may be small, together they can become significant.

Why should this matter?

When measuring RMS voltage or power, any deviation from the ideal 50Ω loading returns some signal to the source, so the full amplitude will not be measured.

For this reason a VSWR of 1.5:1 has been accepted as a practical limit for voltage measurements. When voltage measured in a 50Ω system is used to calculate power levels, a better match is preferred.

A return loss of –26dB, or a VSWR around 1.1:1, ensures the load impedance lies between 55.27 and 45.23Ω.

Simulation versus schematic drawings

I could now draft possible circuit layouts to identify stray capacitances which should be included in my final simulation runs. With due allowance for these, I could determine the attenuator values needed to output 1 volt for a 1V full scale input, Fig. 5.

As you can see, the final schematic input circuit appears simple compared to that needed for acceptable simulations, Fig. 7.

So, were all my component measurements and simulation runs justified when the prototype circuit board was completed, calibrated and its performance measured? In short yes. As you can see from **Table 1**, the prototypes performed superbly.

PCB assembly

The circuit board has been arranged to accept various 50Ω input connectors as well as RG174 coaxial cable.

For a coaxial cable input, the full 2.5 by 1.5in board should be used. This can also accept an SMA PCB right angle socket connector, fitted on the ground plane side.

For best performance using a connector, the board should be shortened and an in-line connector – preferably SMA sized – fitted. It should be soldered so its outer body connects both upper and lower

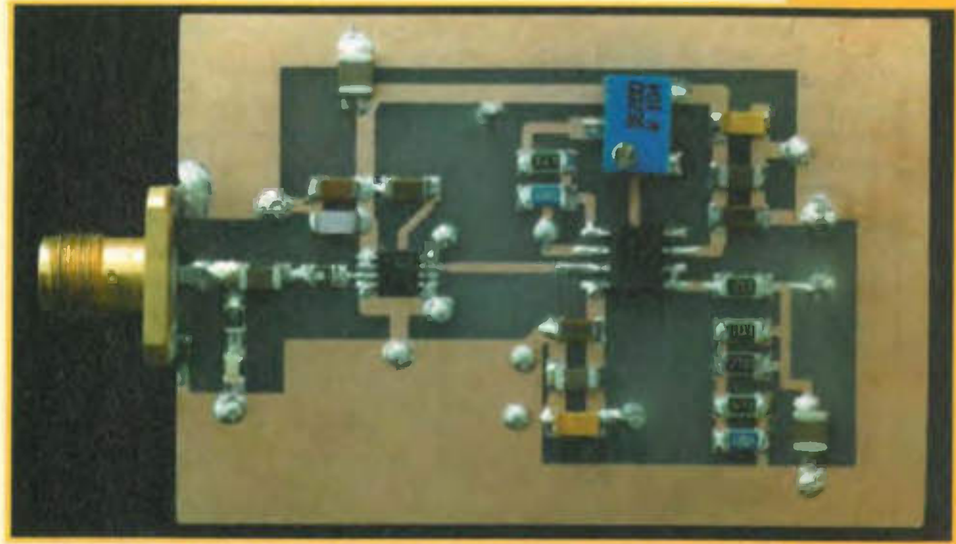


Fig. 3. Illustrating the complete 50Ω RF RMS measuring circuit capable of accurately measuring true RMS from 10kHz to 100MHz. The INA133 instrumentation amplifier can be used to correct no-signal offsets and provide increased output current. However this version needs a ±5 volt supply.

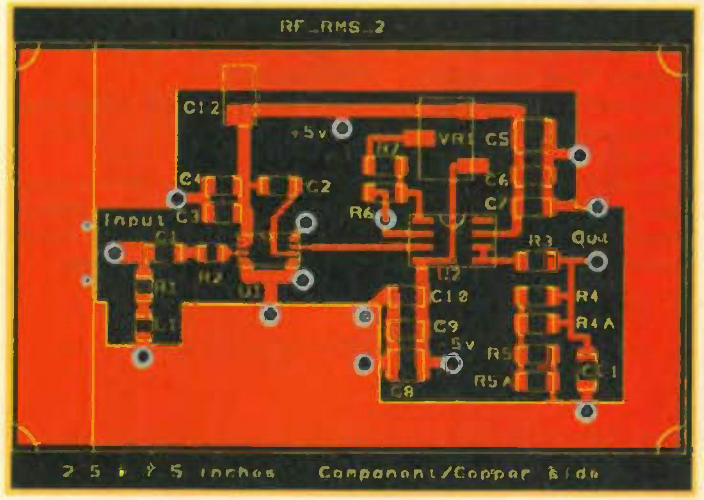


Fig. 4. Final PCB used for both versions and the performance measurements described in the main article. The 50Ω input can accept a PCB connector or coaxial cable. Removing the portion to the left of the vertical line enables various stripline-launching connectors to be used. These provide the best possible input VSWR.

Frequency-domain simulation

The capacitor, resistor and inductor models built into Spice-based simulators assume ideal loss-free components having a constant value regardless of frequency. For transient or time-domain simulation, Spice simulator models automatically provide a facility for amplitude-dependent changes for the semiconductor components. For passive components though, this requires user action.

Unfortunately, with real life passive components, most parameters are frequency dependent.

The latest simulators still assume ideal passive components in their libraries. Some though, among them MC6, provide the facility to override the internal models by using a frequency dependent expression. This is not usable for transient or time domain simulations. Regrettably, suitable model libraries are not provided as far as I am aware.

A restricted number of component models can be downloaded from Intusoft[7] and are supplied with their simulators. These offer a limited choice and rarely exactly fit one's component needs. The modelling approach they used was initiated in 1994 by John Prymak of Kemet.

Kemet[8] now offers a Spice based living data sheet for both their Ceramic and Tantalum capacitors. Available as a free download, this software provides on-screen plots of capacitor behaviour with frequency, including capacitance, ESR, tanδ, inductance, impedance and series and parallel resonances. For any one frequency of interest, the simulation circuit used and its component values can be displayed on screen for use in your transient analysis and narrow band frequency sweeps.

Unfortunately, these simulation component values cannot easily be extracted for use in your wide-band frequency-domain analysis.

The main problem is that this frequency dependent expression relates to an individual element. For example, a resistor that has resistance, capacitance and inductance requires three elements, each with its own frequency expression.

Capacitor models may be considerably more elaborate. It would be convenient if one could download manufacturers' macro models for these passive component parts – or better still S parameters, as has long been possible for ICs.

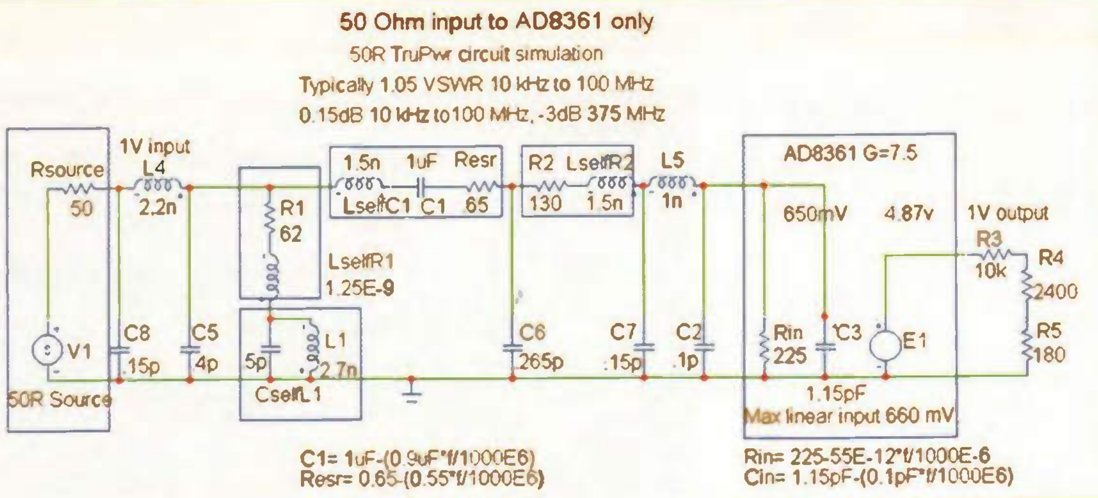


Fig. 5. Schematic used with the MC6 simulator. The 'frequency' expressions used appear below the circuit. Components surrounded by boxes represent the input capacitor, two resistors, the inductor and the AD8361 IC. Remaining elements represent circuit strays, the signal source and the output attenuator.

ground planes. Ideally, as in the photo, a dedicated stripline launching connector should be used.

The complete assembly can be fitted inside a thin-walled screening box, using mounting screws through the connector flange.

Although double-sided, the PCB is easily produced, since its ground plane is complete except for annular insulating bands around the supply and output Vero pins. It can be treated as though single-sided, the component side only being exposed, developed, etched then drilled.

The annular insulating bands in the ground plane are easily formed using a PCB counter-boring tool in the drill. Top and bottom ground planes were interconnected using eleven PCB-through pins, Fig. 4.

I mentioned earlier that the AD8361 in the micro-SOIC package is tiny – and it is. With a leg-to-leg spacing of 0.65mm, care is needed to correctly position all eight legs on the PCB prior to soldering. With each leg being less than 0.4 mm wide, a very fine soldering iron bit is essential.

Performance

The prototypes built and calibrated as shown easily attained my target specification of 1.1:1 VSWR. Measured input return loss was better than –29dB from 100kHz to 100MHz – a VSWR of less than 1.07:1. There's more on this in the panel entitled 'Calibration'.

From 10kHz to 25MHz, amplitude response measured flat. It was –0.25dB at 100MHz, –1.15dB at 1kHz and –3dB at 500MHz.

The above VSWR test results were obtained using an HP8721A reflection bridge to measure return loss and confirmed using an HP4815 vector impedance meter. For the amplitude results I used an HP331A voltmeter up to 1MHz and an HP8405A vector voltmeter for higher frequencies. ■

References

1. Measure AC millivolts to 5MHz, Cyril Bateman *Electronics World*, P.281 April 2000.

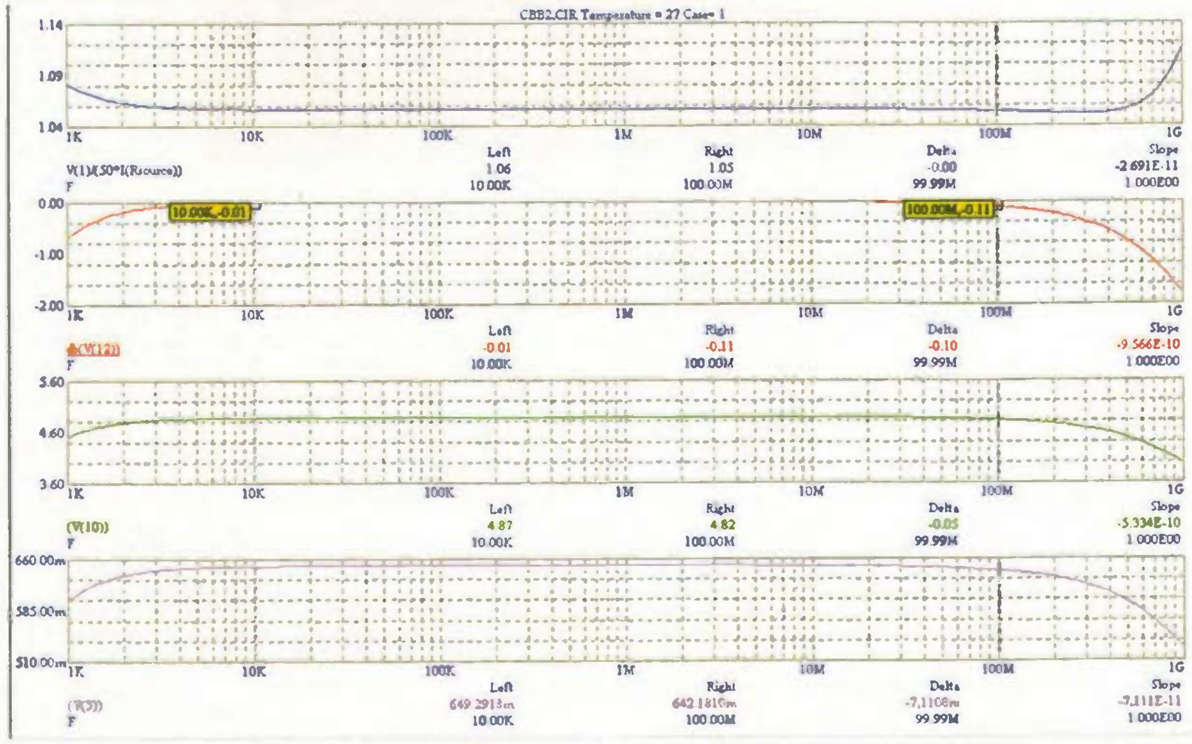


Fig. 6. Simulation results for the circuit of Fig. 5. Up to 100MHz, all simulation plots agree closely with the measured results. The top curve shows VSWR, the other three the attenuated output, AD8361 output and its input when stimulated by a constant voltage 50Ω source.

Measurement equipment used

The HP8721A reflection bridge is specified for use from 100kHz to 100MHz. Together with my precision SMA termination load, it has a measured directivity exceeding –40dB. This directivity permits accurate measurements of return loss to –30dB.

The HP4815 vector impedance meter is calibrated using its non-inductive 100Ω reference resistor. This meter is able to measure in-circuit impedances and phase angles from 1Ω to 100kΩ at frequencies from 500kHz to 108MHz.

The HP331A voltmeter and the HP8405A vector voltmeter have high input impedances. They were used with a MA-COM 50Ω through terminator and 10dB attenuator to establish the signal generator output at the attenuator at exactly 0dBm. The HP331A meter is specified to 3MHz, the HP8405A from 1MHz to 1GHz.

These voltmeters and the MA-COM through termination were replaced by the AD8361 PCB for the voltage measurements. The AD8361 PCB provides its own 50Ω termination so was connected directly to the 10dB attenuator.

Dynamic range was also checked at 100kHz on two prototypes. These were powered from 5.1V and a Hatfield 2100 switched attenuator was connected. The results were spot checked using HP8493A SMA 20 and 30dB certified precision attenuators.

The first AD8361 sample as Fig. 1 exhibited just 2.3mV output offset with no input signal, the second as Fig. 3 included the INA133, which was used to reduce the no-signal output offset from an initial 23mV to 5mV.

Sample one provided a 37dB dynamic range, being better than ±0.5dB from 10mV to 1V. This represents far better performance than Analog's claimed ±1dB dynamic range of 23dB.

Sample two had less range due to the AD8361's no-signal offset, which was close to the data-sheet maximum. The reference voltage of the INA133 instrumentation amplifier was adjusted using VR_1, to reduce no signal output to some 5mV. The output attenuator was then recalibrated to 0dBm.

Sample two provided a 22dB dynamic range at ±0.5dB from 50mV to 1V. It was –1dB at 35mV, to give a 29dB range. This sample also surpassed Analog's claimed ±1dB dynamic range of 23dB.

2. RMS to DC converters ease measurement tasks. http://www.analog.com
3. LT1088, Linear Technology Corporation. http://www.linear-tech.com
4. Revolutionary RF IC Performs RMS-DC Conversion. *Microwaves and RF*, September 1999.
5. Evaluate capacitors for SMPS designs, Cyril Bateman, *Electronics World*, P281 September 2000.
6. High frequency impedance meter, Cyril Bateman, *Electronics World*, P.24 January 2001.
7. Modelling a Capacitor, http://www.intusoft.com/caps.li
8. Kspice185_32.exe. J.Prymak, http://www.kemet.com

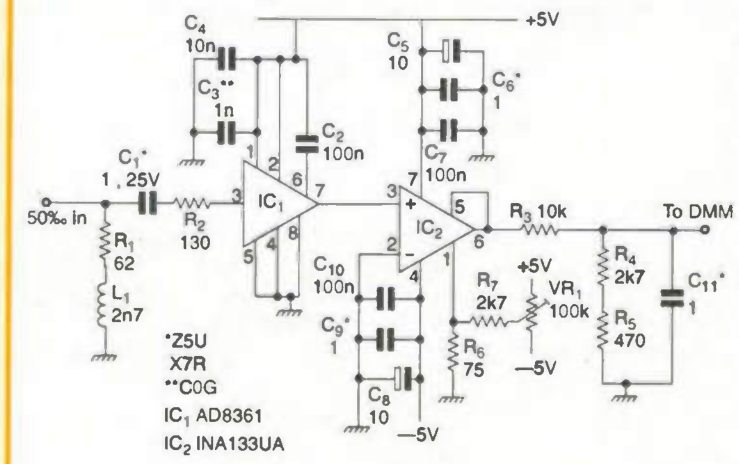


Fig. 7. Final schematic as built and photographed. Provided your AD8361 has a small no-signal offset, then as in Fig. 1, the INA133 and associated components may be omitted. A link is then needed to connect pins 3 and 6 of the INA mounting pads.

Calibrating the meter

The AD8361 data sheet provides detailed calibration instructions but these assume a perfect DC blocking capacitor only at the chip's input. To accommodate a 1V input, an additional attenuating resistor, R_2 as shown in my schematic, is essential.

The output attenuator resistors shown have been calculated to suit an AD8361 having the lowest specified gain. In practice, most AD8361 samples will exhibit higher gain, which can be compensated for by adding the appropriate value R_{4A} and R_{5A} resistors.

I used a simpler calibration sequence than that described in the data sheet.

Assemble the PCB using the nominal values shown but initially omit resistors R_{4A} and R_{5A}. Set VR_1 to mid travel then apply power, but not signal, to the board. Measure the DC level at the attenuator output. If it is greater than +5mV, adjust to obtain approximately +5mV.

Apply a known voltage of 0.25V AC at 10 to 100kHz. For accuracy, a 50Ω signal generator and 50Ω connecting cables should be loaded with a 10dB attenuator at the free end of the cable, then terminated with a 50Ω through load. The 10dB attenuator reduces any load reflections by 20dB as seen by the generator. These reflections are caused by a change in load impedance between the termination and the AD8361 assembly. Voltage at the output of the 50Ω through-terminating load is monitored using a known voltmeter and adjusted to the required level.

When correct, the 50Ω through-terminating load and monitoring meter should be removed and replaced by the AD8361 circuit board.

Measure the DC voltage V_1 at the junction of IC_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:

$$R_x = \frac{0.25\times 10000}{V_1 - 0.25} \text{ in ohms.}$$

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

AD8361 gain=7.5	$R_4 \| \| R_4$=2k4	$R_5 \| \| R_{5A}$=180
Nearest standard values	R_{4A}=21k5	R_{5A}=294
AD8361 gain=8.5	$R_4 \| \| R_{4A}$=1k8	$R_5 \| \| R_{5A}$=470
Nearest standard values	R_{4A}=5k36	R_{5A}=open

Apply signal voltages of 0.05, 0.25 and 0.9 and verify attenuator output accuracy at these voltages.

£500 voucher winner

Winners!

Two of the three winning entries from our ZXF36L01 design competition, announced in the March 2001 issue, are presented in the following articles. All three winners will be receiving a voucher, redeemable at Farnell Electronic Components. Winner David Flatt receives a voucher for £500, the runners up, Mike Button and S. Vijayan Pillai, will receive vouchers of £100 each.

Many thanks to all of you who entered. We hope to publish some of those entries in the near future. All entrants will receive a cheque covering the cost of the Zetex development system.

For more details on the ZXF36L01 filter/mixer, take a look at the March 2001 issue.

Audio spectrum display

Designed to check the spectrum of audio from a hi-fi at various points in the room, David Flatt's audio spectrum monitor wins our ZXT36L01 design competition. The monitor gives an indication of level at eight spot frequencies. Since full source code is presented, both the range and number of frequency steps can easily be altered.

The aim of this design was to examine the concept of making a portable instrument to do a frequency sweep through the audio spectrum. This allows a hi-fi set-up's performance to be assessed at different locations in a room.

Availability of the ZXT36L01 mixer and filter together with other parts already on hand allowed this to be done at relatively little cost. Since this was to be a concept proving exercise, there was no requirement to fully cover the audio spectrum. That could be done later if the circuit proved worthwhile.

The ZXF36L01 together with a PIC16C71 microcontroller and LCD display module form the core of the design. A push button – the only control – starts a sweep through eight spot frequencies. Each frequency is sent for a programmed number of cycles, during which time a capacitor is charged from the filter output. Before the next frequency is sent, the 16C71 performs an a-to-d conversion on the capacitor voltage. It then updates the display and resets the capacitor.

For the prototype, to simplify the programming effort only eight frequencies were used – 500, 1k, 2k, 3k, 4k, 5k, 6k and 7kHz. This choice of frequencies matched the frequency response of the microphone available for the design.

The complete circuit, Fig. 1, consists of seven sub-circuits, the ZXT36L01 mixer and filter, a voltage follower, a rectifier, an integrator, a buffer, a PIC16C71 microcontroller and an LCD module. All these sub-circuits operate from a +5V supply. The only other supply needed is a +2.5V reference voltage; this comes from a TLE2425 voltage splitter. One quad operational amplifier, TLC2274, is used for all signal conditioning.

Besides the push button that initiates a frequency sweep, one LED indicates that the processor is waiting for the button to be pressed, the other indicates when the local oscillator is active.

Details of the filter and mixer section

The ZXT36L01 is used in its mixer configuration with notch-pass filter. I used a centre frequency of 10kHz since this prototype only needed to sweep from 0.5 to 7kHz. Also, the 36L01 evaluation kit is set to operate at 10kHz by default.

All passive components connected to the ZXT36L01 are the same as on the evaluation kit, with the exception of R_2 of Fig. 1. Shown as R_{mix} in the application note, this component had to be increased to 100kΩ.

Keeping a central frequency, F_O, of 10kHz required local oscillator frequencies of 9.5, 9, 8, 7, 6, 5, 4 and 3kHz. Though no great accuracy was required for this application, some care is needed if bass frequencies are to be measured. Some means of trimming F_O may be necessary. Using software to compensate for F_O error would be one solution.

Signal conditioning

Signal conditioning for converting filter output to a voltage that is proportional to average signal level is performed in four stages.

Output from the filter of the ZXT36L01 was unable to drive the rectifier directly. As a result, one amplifier of the quad TLC2274 was used as a voltage follower.

A full-wave precision rectifier – the next stage – may seem over the top for this application but a full-wave rectifier halves the integration time needed.

The third stage is an integrator with a time constant given by,

$$V_{out} = \frac{1}{R_{16}} \times C_{int} \times V_{FO}$$

It is set to fully charge in around 40ms and discharge via RA4 in around 400µs. Resistor R_{20} determines the discharge rate. A value of 1kΩ limits the current through the open drain of RA4 to 5mA while giving enough time for CINT to fully discharge.

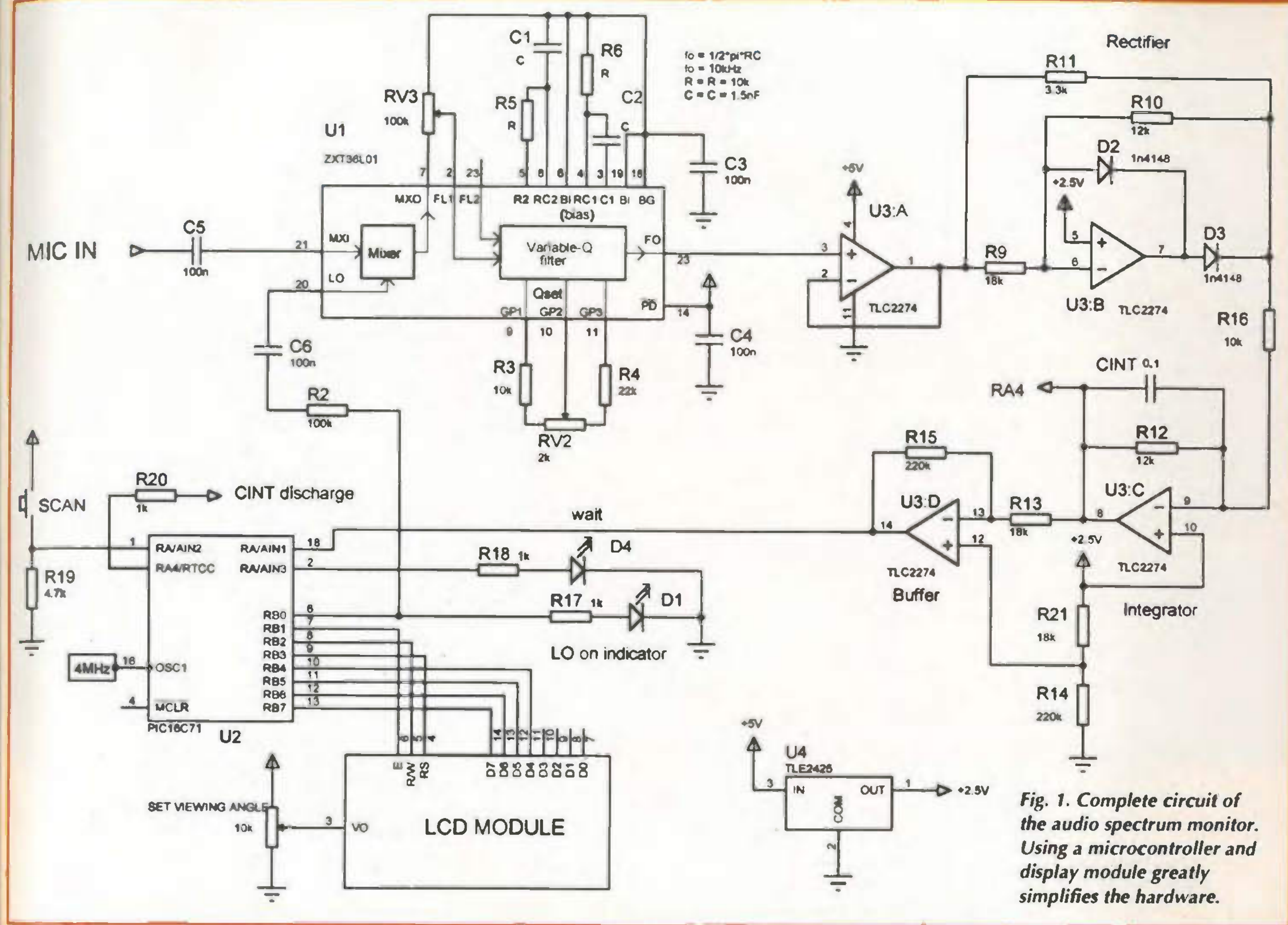


Fig. 1. Complete circuit of the audio spectrum monitor. Using a microcontroller and display module greatly simplifies the hardware.

A final buffer stage U3:D subtracts the 2.5V bias from the integrated voltage and provides gain to give a +4.65V output to the a-to-d converter input, when the filter output has been at its maximum during the integration period.

The signal-conditioning part of the circuit was simulated using ISIS, Proteus's simulation software. **Figures 2** and **3** show the results when driven by the maximum 1.6V pk-pk output swing of the filter. In practice the value of R_{21} had to be trimmed to get the desired output.

Charge time of 40ms determines the gain of the circuit after the filter. Increasing the number of cycles sent increases the gain. Each frequency can have a different number, thus allowing frequency compensation to be programmed into the instrument.

Why choose a PIC controller?

A PIC16C71 was chosen for the microcontroller because it had the necessary I/O lines and an a-to-d conversion channel. A 4MHz oscillator unit provides the system clock. Line AIN1 was used for the a-to-d channel, RA2 for the scan button, RA3 for the LCD indicator, RA4 for integrator control, RB0 for output to the local oscillator and RB1-7 for LCD module control and data lines. Only AIN0 remains unused. This line can not be used for digital I/O.

Data lines to the LCD module could be made multifunctional with appropriate software allowing extra

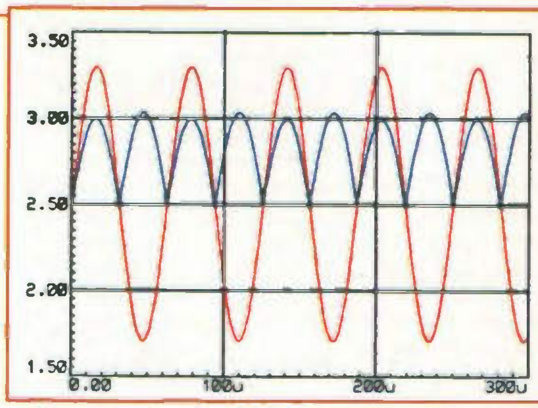


Fig. 2. Isis simulations of the rectifier's input and output.

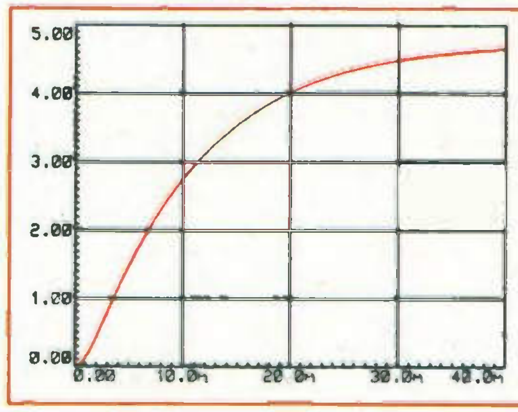


Fig. 3. Buffer output for maximum filter signal, again an Isis simulation.

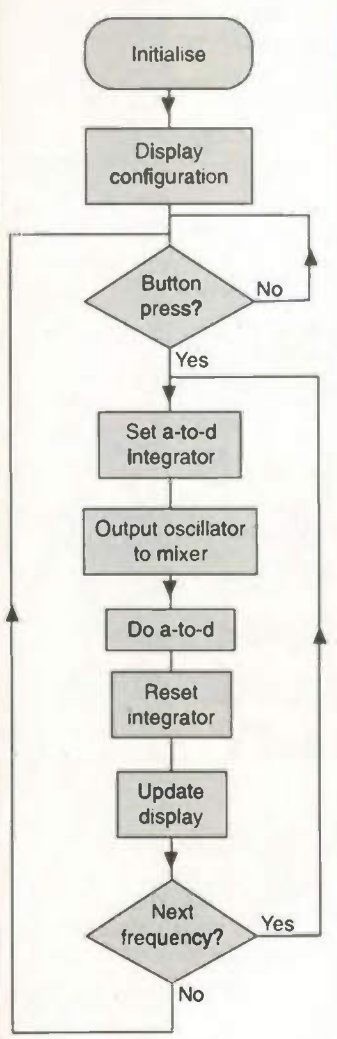


Fig. 4. Software flow

Fig. 5. Each press of the scan button updates the spectrum bar graph and displays the next frequency up the scale, represented by the left-most digit, together with its amplitude, which in this case is 178mV.

control. This extra capability could be used to change to different frequency sweeps, different scaling of the bar graph or incorporate some internal gain control, for example.

For clarity, reset circuitry on pin 4 is not shown. For the same reason, several power connections have been omitted from the circuit diagram.

Program details

Flow chart Fig. 4 shows the structure of the programme, Lists 1 and 2. Header file LCD_4.H, List 2, contains subroutines needed to use the LCD module. It also handles the graph display and character conversion. Some of these routines are adaptations from Bluebird Electronics Pic Millennium board used in the development. Besides initialisation, List 1 contains subroutines, the main program loop and the three look-up tables for controlling frequency and duration of local oscillator.

Critical timing loops rely on two tables, GET_INTH and GET_INTL. These hold coefficients that the subroutine DELAY uses to set the period of the frequency sent to RB0 by PULSE. Subroutine GET_INTH has a loop that is four cycles long and GET_INTL one of three cycles.

Using a combination of the two gives a frequency resolution of one cycle for each DELAY call. There are two calls per cycle giving a resolution of 2μs with a 4MHz oscillator.

Table 1 shows frequencies and error from the nominal intended value. These values seem very good until you consider the input values they relate to. Errors relating to input values are calculated and tabulated, also in table 1. These calculations are based on a nominal F_O of 10kHz

Table 1. Nominal and actual sweep frequencies using 4MHz oscillator.

Nominal LO	Actual Hz	% error	F_O–LO	% error
9.5kHz	9443	0.6	557	11.4
9kHz	8936	0.7	1074	7.4
8kHz	7949	0.7	2051	2.5
7kHz	7042	0.6	2958	1.4
6kHz	6028	0.5	3972	0.9
5kHz	4980	0.4	5020	0.4
4kHz	4019	0.5	5981	0.4
3kHz	2993	0.3	7007	0.1

and so by altering this in hardware there is scope to improve the bottom end of the frequency range. Versions of Microchip's PICs can run at 20MHz giving a resolution of 400ns. Using a higher frequency oscillator and trimming F_O give the lower usable frequency of the design.

The number of cycles sent is stored in the third look-up table, N_LOOP. For this prototype values selected gave a duration of 40ms for each frequency.

Before the integrator is reset, CALL A_D converts the voltage on CINT and stores the value in the ADRES register. After resetting the integrator CALL A_D_DIS updates the LCD module display.

First the hexadecimal value in ADRES is converted to ascii and then sent to the LCD-Module, it is then converted into a bar-graph value with a flag being set if the value of ADRES, 7 is set. The bar-graph value, register GS, is then used to load the relevant character generator address of the LCD module.

There are eight addresses per user character and one character for each frequency. After the frequency sweep GRAPH updates the bar graph by writing to the LCD module. The character sent is the user character, 0xfe or 0xff depending on the GSF register flag setting. The eight bits of GFS correspond to the eight frequencies.

A loop back to N_SCAN for a button press completes the programme.

LCD display module

The display is a 16-by-2-line display conforming to the industry standard HD44780. A bar-graph, formed by converting the eight analogue-to-digital conversion values into eight user symbols, is the main display. By using a combination of user, blank or clear characters, the bar-graph is extended to two lines.

A secondary display of the frequency number and its value is also available. Only one frequency can be displayed at a time and this is stepped on each button press detected. A typical display that can be expected is shown, Fig. 5, a particularly poor base response being represented.

In summary

Testing the circuit on bread board, using a signal generator, proved the circuit functioned as designed. However as expected bread board is not the environment to use if you want low noise and reliable circuit operation.

If time had allowed, I would have experimented with a microphone preamplifier with a fixed gain of 40dB. With the microphone on hand, this should provide around 300mV to the mixer at reasonable hi-fi volume levels. This voltage is the mixer's maximum rating. Since microphones are available with built in preamplifiers this stage may be redundant.

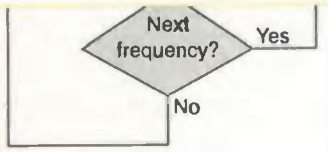
The design and development of the above has shown a low cost application for the ZXF36L01, with scope for development into other instrument applications in the audio field.

I have made no attempt to calibrate the prototype and so its readings are of a comparative nature. With a fully calibrated system the number of cycles per frequency can be altered to compensate for the frequency response of the microphone say.

The number of frequency steps that can be displayed is limited by the number of user symbols that the LCD module has. With more programming, the number of frequency steps can be increased to 40 by using five steps per symbol and limiting the display to one line. ■


```
        addwf   pcl,f
        retlw   .00             ;not used
        retlw   .3              ;9.5khz
        retlw   .3              ;9kHz
        retlw   .1              ;8kHz
        retlw   .3              ;7kHz
        retlw   .3              ;6kHz
        retlw   .5              ;5kHz
        retlw   .2              ;4kHz
        retlw   .3              ;3KHz
end_inth
        retlw   0
get_intl        ;interval coefficients low for delay
        addwf   pcl,f
        retlw   .00     ;not used
        retlw   .5      ;9.5kHz
        retlw   .6      ;9kHz
        retlw   .11     ;8kHz
        retlw   .11     ;7kHz
        retlw   .15     ;6kHz
        retlw   .18     ;5kHz
        retlw   .30     ;4kHz
        retlw   .43     ;3kHz
end_intl
        retlw   0
n_loop          ;number of pulses to do
        addwf   pcl,f
        retlw   .00     ;not used
        retlw   .190
        retlw   .180
        retlw   .160
        retlw   .140
        retlw   .120
        retlw   .100
        retlw   .80
        retlw   .60
end_n
        retlw   0
;delay 0.5ms * value in w register
delay1  movwf   count   ;use this register temporarily
dly1    movlw   0a8h    ;offset
```

```
        goto    pulse2
        return
;A/D routines
a_d                     ;gets a/d value
        movlw   0c9h    ;RC osc RA1,ADON
        movwf   adcon0
        movlw   .32
        movwf   count
s_wait  decfsz  count,f
        goto    s_wait
        bsf     adcon0,go
        nop
ad_go   btfsc   adcon0,go
        goto    ad_go
        return
a_d_dis                 ;displays a/d result
        movfw   adres   ;load a/d value and convert
        btfss   flags,0
        goto    up_dis
        call    h_to_d  ;it to decimal if flag set
        call    line2
        movfw   count4  ;write result
        addlw   030h
        call    wdata
        movlw   ' '
        call    wdata
        movfw   hunds
        call    wdata
        movfw   tens
        call    wdata
        movfw   units
        call    wdata
up_dis
        call    load_cg1 ;update cgram
        return
;Initalise ports and tmr0
init
        clrf    portb
        bsf     pa0
        movlw   0c7h    ;RA4 & RA3 output,others input
        movwf   trisa
```



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List 1. Software for the PIC controller in the audio spectrum display. E-mail j.lowe@cumulusmedia.co.uk for an electronic copy, quoting 'spectrum-monitor software' in the subject heading.

```
;***** equates *****
#define  ad0      porta,0
#define  ad1      porta,1
#define  con1     porta,2   ;start button
#define  con2     porta,3   ;led wait
#define  a_latch  porta,4   ;a/d latch
#define  osc_out  portb,0   ;osc output
#define  red1     portb,0   ;
#define  E        portb,3   ;lcd enable
#define  RW       portb,2   ;lcd read/write
#define  RS       portb,1   ;lcd register select
count    equ      0ch       ;general purpose counter
count1   equ      0dh
count2   equ      0eh
count3   equ      0fh
count4   equ      10h       ;frequ. count
count5   equ      11h       ;pulse count
temp1    equ      12h       ;use only in subroutines
del_reg  equ      13h
sdat     equ      14h
char     equ      15h
flags    equ      16h
gs       equ      22h       ;bar value
gsf      equ      23h       ;store flags
units    equ      24h
tens     equ      25h
hunds    equ      26h
         list p=16c71
         #include<p16c71c.inc>
         org      0
         goto init
;SUBROUTINES lcd subroutines using port b only
;includes routines for using character ram
         #include<lcd_4f.h>
;LOOK UP TABLES
;interval coefficients high for delay
get_inth
         addwf    pcl,f
         retlw    .00              ;not used
         retlw    .3               ;9.5khz
         retlw    .3               ;9kHz
         retlw    .1               ;8kHz
         retlw    .3               ;7kHz
         retlw    .3               ;6kHz
         retlw    .5               ;5kHz
         retlw    .2               ;4kHz
         retlw    .3               ;3KHz
end_inth
         retlw    0
get_intl          ;interval coefficients low for delay
         addwf    pcl,f
         retlw    .00       ;not used
         retlw    .5        ;9.5kHz
         retlw    .6        ;9kHz
         retlw    .11       ;8kHz
         retlw    .11       ;7kHz
         retlw    .15       ;6kHz
         retlw    .18       ;5kHz
         retlw    .30       ;4kHz
         retlw    .43       ;3kHz
end_intl
         retlw    0
n_loop            ;number of pulses to do
         addwf    pcl,f
         retlw    .00       ;not used
         retlw    .190
         retlw    .180
         retlw    .160
         retlw    .140
         retlw    .120
         retlw    .100
         retlw    .80
         retlw    .60
end_n
         retlw    0
;delay 0.5ms * value in w register
delay1  movwf     count     ;use this register temporarily
dly1     movlw    0a8h      ;offset
         movwf    count1
         decfsz   count1,f
         goto     $-1
         decfsz   count,f   ;decrement, until zero
         goto     dly1
         retlw    0
;Delay for frequency period
delay    movlw    HIGH get_inth   ;load pclath with hi
address
         movwf    pclath
         movfw    count4    ;get coeffs from look-up tables
         call     get_inth
         movwf    count1
         movlw    HIGH get_intl
         movwf    pclath
         movfw    count4
         call     get_intl
         movwf    count2
count_1  nop                ;high count
         decfsz   count1,f
         goto     count_1
count_2  decfsz   count2,f  ;low count
         goto     count_2
         return
;Pulse osc_out
pulse    bsf      osc_out
         nop
         nop
         nop
         call     delay
         bcf      osc_out
         call     delay
         decfsz   count3,f
         goto     pulse
pulse2   bsf      osc_out
         nop
         nop
         nop
         call     delay
         bcf      osc_out
         call     delay
         decfsz   count5,f
         goto     pulse2
         return
;A/D routines
a_d                         ;gets a/d value
         movlw    0c9h      ;RC osc RA1,ADON
         movwf    adcon0
         movlw    .32
         movwf    count
s_wait   decfsz   count,f
         goto     s_wait
         bsf      adcon0,go
         nop
ad_go    btfsc    adcon0,go
         goto     ad_go
         return
a_d_dis                     ;displays a/d result
         movfw    adres     ;load a/d value and convert
         btfss    flags,0
         goto     up_dis
         call     h_to_d    ;it to decimal if flag set
         call     line2
         movfw    count4    ;write result
         addlw    030h
         call     wdata
         movlw    ' '
         call     wdata
         movfw    hunds
         call     wdata
         movfw    tens
         call     wdata
         movfw    units
         call     wdata
up_dis
         call     load_cg1  ;update cgram
         return
;Initalise ports and tmr0
init
         clrf     portb
         bsf      pa0
         movlw    0c7h      ;RA4 & RA3 output,others input
         movwf    trisa
```

```
        movlw   00h         ;all output
        movwf   trisb
        movlw   b'10000000'     ;turn weak pullup off
        movwf   optreg
        movlw   2           ;RA0,RA1 analogue-RA2,RA3 digital
        movwf   adcon1
        bcf     pa0
        bcf     con2
;Program begins here
main
        movlw   014h        ;setup display
        call    delay1
        call    discfg
call    line1
        movlw   0           ;Table address of start of
message
        call    msg         ;prints 'Freq.'
        clrf    flags
        bsf     flags,0     ;display counter
        bcf     a_latch     ;external integrator off
n_scan  bsf     con2        ;wait for button press, light led
        btfss   con1
        goto    $-1
        movlw   07ah        ;wait for debounce
        call    delay1
        clrf    gsf
        bcf     _c
        btfsc   flags,7     ;update display counter
        bsf     _c
        rlf     flags,f
bcf     con2        ;led off
        movlw   8           ;number of frequency steps
        movwf   count4
n_freq  movlw   HIGH n_loop     ;Load pclath with hi
address
        movwf   pclath
        movfw   count4      ;start loop
        call    n_loop      ;load n pulses to do
        movwf   count3
        movwf   count5      ;n*2
        bsf     a_latch     ;extrnal integrator on
        call    pulse       ;oscillator on
        call    a_d         ;a/d
        bcf     a_latch     ;integrator off
        call    a_d_dis     ;update display
        bcf     _c
        btfsc   flags,7     ;update display counter
        bsf     _c
        rlf     flags,f
        decfsz  count4,f    ;? another frequency
        goto    n_freq
        call    graph       ;update bar graph
        goto    n_scan      ;? another scan, n_scan
        goto    $           ;stay here
        end
```

List 2. Subroutines for controlling the 16 by 4 LCD module. This file must be called 'lcd_4f.h' in order to be included in the main program.

```
;This is for 4-bit handshake mode data transfer for
;the 16C71,using port B. Timing for 4MHz oscillator.
;Includes graph routines used in ZETEX filter project.
load_cg1                    ;loads one cg character
        movfw   count4      ;address offset
        call    ch1a        ;set address of cgram
        movfw   adres       ;get data
        bcf     _c
        rlf     gsf,f       ;position flag register
        call    scale_bar1  ;turn data into bar
        call    cgd_big     ;load cgram
        return
;Display character array
graph
        clrf    count2
nxt_p   movfw   count2
        call    graph1      ;address of top line + count2
        movlw   0feh        ;default zero character
        btfsc   gsf,0       ;test high or low
        movfw   count2      ;address of character
        call    wdata
        movfw   count2
        call    graph2      ;bottom line
        movlw   0ffh        ;default 1's character
        btfss   gsf,0
        movfw   count2
        call    wdata
        rrf     gsf,f       ;ready for next plot
        incf    count2,f
        btfss   count2,3
        goto    nxt_p
        return
;Scale linearly
scale_bar1
        clrf    gs
        movwf   temp1
        movf    temp1,f     ;update zero flag
        bz      all_0
        bcf     count,0
        btfsc   temp1,7     ;test for msb
        bsf     count,0     ;set coz
        movlw   HIGH convert
        movwf   pclath
        swapf   temp1,w
        andlw   07h         ;strip, leaving 3 low bits
        call    convert
        btfsc   count,0
        bsf     gsf,0
        movwf   gs
        return
all_0   return
convert addwf   pcl,f
        retlw   1
        retlw   3
        retlw   7
        retlw   b'00001111'
        retlw   b'00011111'
        retlw   b'00111111'
        retlw   b'01111111'
        retlw   b'11111111'     ;bar pattern
;display using two vertical characters
cgd_big
        clrf    count
next_ch movlw   0
        btfsc   gs,7
        movlw   0ffh        ;if high send ones
        call    wdata
        rlf     gs,f        ;ready for next character
        incf    count,f
        btfss   count,3
        goto    next_ch
        return
;Setup and primatives
set_lcd bcf     E
        bcf     RS
        bcf     RW
        return
disclr  call    set_lcd
        movlw   20h
        movwf   portb
        call    pulsee
        movlw   80h
        movwf   portb
        call    pulsee
        return
short   movlw   .1          ;delay for 4MHz osc
        movwf   del_reg
        decfsz  del_reg,f
        goto    $-1
        return
send    call    set_lcd
        movwf   char        ;temp store
        movf    char,w
        andlw   0xf0        ;get upper nibble
        movwf   portb
        call    pulsee
        call    ckbusy
        swapf   char,w
        call    set_lcd
        andlw   0xf0        ;get lower nibble
        movwf   portb
        call    pulsee
        call    ckbusy
        return
wdata   call    set_lcd
        movwf   char        ;temp store
        movf    char,w
        andlw   0xf0        ;get upper nibble
        movwf   portb
        bsf     RS
        call    pulsee
        call    ckbusy
        swapf   char,w
        call    set_lcd
        andlw   0xf0        ;get lower nibble
        movwf   portb
        bsf     RS
        call    pulsee
        call    ckbusy
        return
ckbusy
        bsf     pa0         ;page1
        movlw   0x80        ;set bit 7 as input
        movwf   trisb
        bcf     pa0         ;page0
        bcf     RS
        bsf     RW
        bsf     E
        call    short
        movfw   portb       ;read high nibble
        bcf     E
        movwf   sdat
        call    short
        call    pulsee      ;skip low nibble
        btfsc   sdat,7      ;test high nibble
        goto    ckbusy
        bsf     pa0         ;page1
        clrf    trisb       ;reset as output
        bcf     pa0         ;page0
        return
discfg  call    disclr      ;setup for 4bit two lines
        movlw   0ch
        call    send
        movlw   06h
        call    send
        movlw   01h
        call    send
        call    short
        return
pulsee
        call    short
        bsf     E
        call    short
        bcf     E
        return
;Set data address subroutines
line1   movlw   80h
        call    send
        call    short
        return
line2   movlw   0c0h
        call    send
        call    short
        return
graph1  addlw   087h
        call    send
        call    short
        return
graph2  addlw   0c7h
        call    send
        call    short
        return
;start address of cgram
ch1
        movlw   40h
        call    send
        call    short
        return
;start address of cgram + offset
ch1a
        movwf   temp1
        decf    temp1,f
        rlf     temp1,f     ;multiply w by 8
        rlf     temp1,f
        rlf     temp1,f
        movfw   temp1
;& add to start of cgram address
        addlw   40h
        call    send
        call    short
        return
msg     movwf   temp1       ;display message
        call    Table
        andlw   0ffh        ;check if end of msg
        btfsc   _z
        return
        call    wdata
        movf    temp1,w
        addlw   1
        goto    msg
;subroutines for htu
ten     equ     b'00001010' ;use equate
h_to_d                      ;convert hex to decimal
;       movlw   .154        ;test number
        movwf   count
        clrf    units
        clrf    tens
        clrf    hunds
        movf    count,f     ;update zero flag
        bz      end_htu     ;zero do nothing
nxt_htu call    test_ten
        decfsz  count,f
        goto    nxt_htu
end_htu movlw   030h        ;convert to ascii
        addwf   units,f
        addwf   tens,f
        addwf   hunds,f
        return
test_ten
        incf    units,f
        movlw   ten
        xorwf   units,w
        btfsc   _z
        call    test_hun
        return
test_hun
        clrf    units
        incf    tens,f
        movlw   ten
        xorwf   tens,w
        btfsc   _z
        call    set_ht
        return
set_ht
        clrf    tens
        incf    hunds,f
        return
;Messages
Table   addwf   pcl,f
        retlw   'F'
        retlw   'r'
        retlw   'e'
        retlw   'q'
        retlw   '.'
Table_end
        retlw   0
;End
```

High-quality speech processor for comms

Mike Button has found that the ZXF36L01 filter makes implementing a high-quality speech processor for SSB a much easier job. His circuit takes the speech signal outside the audio band, where it can be clipped without introducing significant amounts of distortion.

The arrival of the Zetex ZXF36L01 variable Q filter with integral mixer has made the provision of a high quality speech processor for single side-band transmission much simpler.

It has been demonstrated that in ordinary speech, intelligence is conveyed mainly by the frequency content of the signal and much less than by amplitude variation. [Thomas & Niderjohn 'The intelligibility of filtered-clipped speech in noise', published in *Journal of the Audio Engineering Society*, Vol. 18 No 3., 1970]

Where the transmission of intelligence benefits from a signal having a low peak-to-trough power ratio, such as the audio modulation for single side band transmissions, a peak amplitude limiting circuit is often employed. By amplifying the audio signal and then 'clipping' the output amplitude to a defined maximum level, the lower level parts of the signal are raised above the transmitting media noise background.

The use of a simple amplitude clipper, while maintaining a relatively constant output for varying levels of input, causes signal distortion by producing

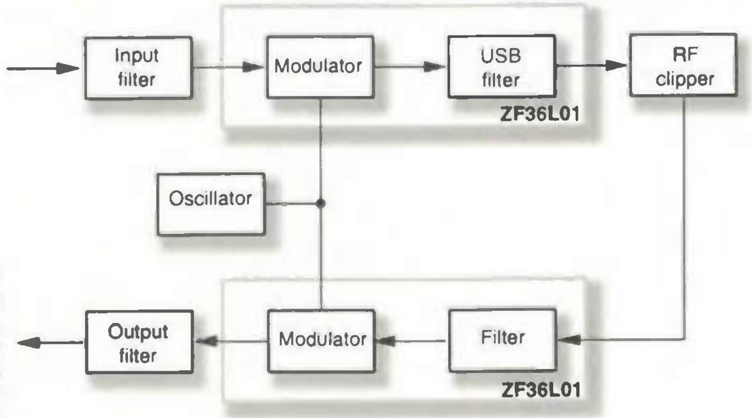


Fig. 1. Simple clippers enhance speech intelligibility, but they also distort the signal. By moving the signal outside the audio band and limiting it there, harmonic signal distortion is reduced.

Chart of filter centre frequency for given resistance and capacitance values. Resistance R is in kΩ and frequency in Hz.

R	1.0nF	1.2nF	1.5nF	1.8nF	2.2nF	2.7nF	3.3nF	3.9nF	4.7nF	5.6nF	6.8nF	8.2nF
1.0	159155	132629	106103	88419	72343	58946	48229	40809	33863	28421	23405	19409
1.1	144686	120572	96458	80381	65767	53588	43844	37099	30784	25837	21277	17645
1.2	132629	110524	88419	73683	60286	49122	40191	34007	28219	23684	19504	16174
1.3	122427	102022	81618	68015	55649	45343	37099	31392	26048	21862	18004	14930
1.5	106103	88419	70736	58946	48229	39298	32153	27206	22575	18947	15603	12939
1.6	99472	82893	66315	55262	45214	36841	30143	25506	21164	17763	14628	12131
1.8	88419	73683	58946	49122	40191	32748	26794	22672	18813	15789	13003	10783
2.0	79577	66315	53052	44210	36172	29473	24114	20404	16931	14210	11703	9705
2.2	72343	60286	48229	40191	32883	26794	21922	18550	15392	12918	10639	8822
2.4	66315	55262	44210	36841	30143	24561	20095	17004	14109	11842	9752	8087
2.7	58946	49122	39298	32748	26794	21832	17863	15114	12542	10526	8669	7189
3.0	53052	44210	35368	29473	24114	19649	16076	13603	11288	9474	7802	6470
3.3	48229	40191	32153	26794	21922	17863	14615	12366	10261	8612	7092	5882
3.6	44210	36841	29473	24561	20095	16374	13397	11336	9406	7895	6501	5391
3.9	40809	34007	27206	22672	18550	15114	12366	10464	8683	7287	6001	4977
4.3	37013	30844	24675	20563	16824	13708	11216	9490	7875	6609	5443	4514
4.7	33863	28219	22575	18813	15392	12542	10261	8683	7205	6047	4980	4130
5.1	31207	26006	20805	17337	14185	11558	9457	8002	6640	5573	4589	3806
5.6	28421	23684	18947	15789	12918	10526	8612	7287	6047	5075	4179	3466
6.2	25670	21392	17113	14261	11668	9507	7779	6582	5462	4584	3775	3131
6.8	23405	19504	15603	13003	10639	8669	7092	6001	4980	4179	3442	2854
7.5	21221	17684	14147	11789	9646	7860	6431	5441	4515	3789	3121	2588
8.2	19409	16174	12939	10783	8822	7189	5882	4977	4130	3466	2854	2367
9.1	17490	14575	11660	9716	7950	6478	5300	4485	3721	3123	2572	2133

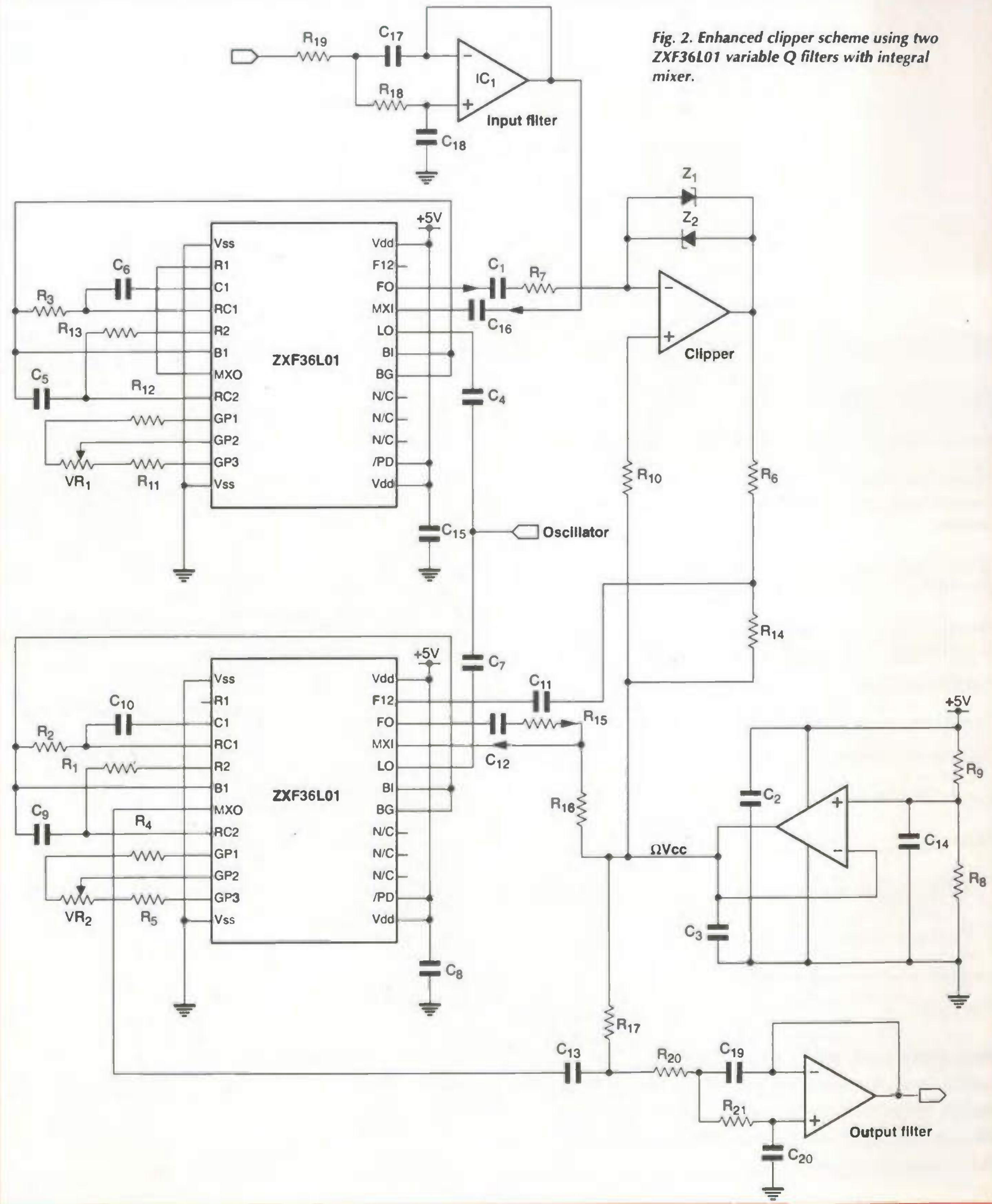


Fig. 2. Enhanced clipper scheme using two ZXF36L01 variable Q filters with integral mixer.

harmonics in the pass band. For example, an audio signal at 400Hz, when clipped, produces harmonics at 800, 1200, 1600... , 2800, 3200 etc., all of which are in band and contribute to the total distortion of the processed signal.

If amplitude limiting is arranged to take place outside the base (audio) band then harmonic signal distortion will be reduced.

The base band signal can be converted to a supersonic signal (VLF signal) using the modulator and filter such as those provided in the ZXF36L01.

Block diagram **Fig. 1** shows the principal functions of the speech processor.

After being filtered to remove all frequencies above the audio band, the audio signal is converted to a supersonic signal, using a modulator and a VLF oscillator. The resultant amplitude modulated signal is then filtered to obtain one sideband – normally the upper sideband – which is then fed into an amplitude limiter.

The resultant VLF 'squared' wave is then passed through a filter before being applied to a de-modulator fed with the same frequency as the modulator. The output of the demodulator is then passed through a low pass filter and buffer amplifier.

The final output is an audio signal with minimum amplitude variation with all the intelligence in the frequency of the signal.

Circuit example

The choice of the modulator frequency is a compromise between obtaining sufficient rejection of the input signal and the Q factor of the SSB filter. A suitable frequency is in the range of 10kHz to 20kHz. This frequency can be obtained by an oscillator and divide by two from a low cost 32.768kHz watch crystal.

Given the oscillator frequency of 16.384kHz and with an audio bandwidth of 3kHz the centre frequency of the filters should be 16384+1500=17884Hz. Using the table derived from a spread sheet program the values of the resistor and capacitor for the filter part of the ZXF36L01 is as follows: R=3.3kΩ, C=2.7nF.

Figure 2 shows the circuit arrangements. ■

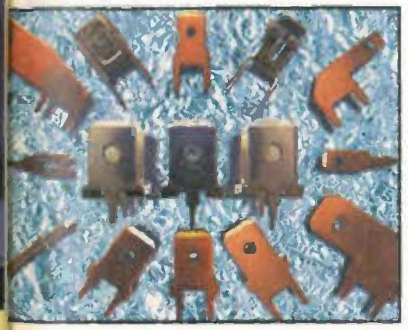
NEW PRODUCTS

Please quote *Electronics World* when seeking further information

PCB hardware connectivity products

Easby Electronics is offering Vogt cold-pressed metal parts for PCB connectivity. Products for direct PCB insertion include various sizes and styles of spade terminals, soldering lugs and bridge connectors, angle pins and straight contact pins. Metal contact pieces for PCBs are contained in a 13-section catalogue. Mounting spade terminals are available in 2.8, 4.8 and 6.3mm widths for single and multiple connections. The 2.8mm width single spade terminal, for example, comes in 86 styles for 90° connection and 24 for vertical connection. Soldering lugs come in more than 200 sizes, as well as specials that include bridge connectors. Other metal parts include angle pins, straight and wire-wrap angle pins, bus bars, push-fit contacts for board-edge connectivity and straight pins to receive standard cable sockets.

Easby Electronics
Tel: 01748 850555
www.easby.com



Intelligent PMC host I/O board for VME

Concurrent Technologies has announced a single-slot 6U VME PMC host board with a 566 or 733MHz Celeron processor. With two PMC sites, and the possibility to add a further two PMC sites in a second slot, the VP PMC/C1x is suitable for I/O intensive embedded applications. The board has two 32-bit PMC sites for user functions or for adding standard PMC modules. The PMC I/O connector signals are routed to P2, P0 or through the front panel of the board. An on-board connector allows for the addition of the optional PMC carrier board – the AD CR2/PMC – which allows the use of four PMC modules in two VME slots. The Celeron processor has 32kbyte of level one cache and a 128kbyte level two cache. The board uses the 82440BX chip set, which can support the 100MHz front side bus; this ensures that the board can also support higher performance Celeron and Pentium III devices. Features include support for up to 320Mbyte SDRAM, 10/100Mbit/s Ethernet interface via a front panel RJ45 connector, EIDE interface, 16Mbyte application flash EPROM and an optional 512kbyte battery-backed SRAM. Hardware byte-swapping and bus error interrupt hardware are provided. The board supports standard operating systems including Linux, QNX, Solaris, VxWorks and Windows NT.

Concurrent Technology
Tel: 01206 752626
www.cct.co.uk

Core module pair with Ethernet

Rabbit Semiconductor has announced two Ethernet core modules in its Rabbitcore line.

RF digitiser for 3G digital mobile test

IFR Systems is targeting the first in a series of new signal analysis instruments, the 2319E RF digitiser, at 2G, 2.5G and 3G digital mobile system testing. The instrument is designed to convert broadband radio frequency (RF) signals into digitised data for external processing in a PC. When combined with an optional high power digital signal processing (DSP) card and IFR software, the 2319E can be used as a substitute base station receiver during early research and development phases of new mobile phones. The 2319E is a benchtop or rack-mounted instrument for use in R&D. It provides a 20MHz digitisation bandwidth which the supplier claims is sufficient to capture four UMTS radio channels. The digitised output data is made available through a variety of interfaces for external processing in a PC using application software developed by IFR, the customer or by a third-party developer. The choice of output interface is determined by the bandwidth requirements of the application, selected from low, medium or high-speed options. Data may be transferred in real-time directly from the digitiser or captured into internal memory and output in bursts at a slower rate. In either case, data can be represented as a digital IF (intermediate frequency), or digital IQ (in phase & quadrature phase).

IFR Systems
Tel: 01438 772087

No bigger than a credit card, these microprocessor modules are for developing embedded systems with integrated Ethernet connectivity. The RCM2100 and RCM2200 provide control and communications capabilities with instant local or worldwide connectivity. They include a Rabbit microprocessor, up to 1Mbyte of flash and SRAM and an Ethernet interface and connector. Features include four serial ports, battery-backable clock, cold boot capability, slave mode operation and up to 40 I/O lines. The modules are supported by software including a TCP/IP stack, web server and C development system. The TCP/IP stack and related software are supplied in source form and there are no run-time royalties. They are also available without the integrated Ethernet port.
Rabbit Semiconductor
Tel: 00 1 530 757 8400
www.rabbitsemiconductor.com

Materials absorb microwaves

Two flexible, absorbent EMI materials have been introduced by Murata. The EA10 prevents abnormal oscillation in high frequency modules and suppresses interference between circuits. The EA20 has high μ, high loss characteristics for suppression over a wide band in digital products. Both cover 0.1 to 20GHz. They are flexible and can be held in place with adhesive tape. Thicknesses available are from 0.2 to 2.7mm. The thickness depends on the frequency to be suppressed.
Murata
Tel: 01252 811666
www.murata.co.uk

Current transformers get lift and drive

Reo has extended its range of current transformers to include versions for lifts, drives and medium power equipment using output bus bars. All provide a linear output signal proportional to the measured current. Compensated current transformers are used for the potential free measurement, or recording, of AC, DC or impulse current. The current transformers for lifts and drive-engineering applications are epoxy-sealed in housings with universal mounting flanges, allowing for use in any orientation. The current measuring range is 25 to 200A at mains frequency. The standard units have a quick release connector for the output cables. For the medium power range, the construction lets the transformer be slipped over a bus bar. Larger units can be sandwiched between two bus bars. These transformers are available for measuring up to 1200A. Other transformers provide a DC output signal that can be fed directly into a control system such as a PC or PLC. There is also a version that has built-in circuitry to control contacts, so they close when a preset current level, set by a trimmer, is reached.
Concurrent Technology
Tel: 01206 752626
www.cct.co.uk

Processor suits web communications

Sequoia is introducing a Crystal Communications processor for next generation Internet communications devices such as cable modems, ADSL modems, cable set-top boxes and VoIP telephones. The Cirrus Logic CS89712 is an integrated network enabled system-on-a-chip and combines a 74MHz ARM720TDMI CPU core with 10Mbit/s Ethernet connectivity. It incorporates system peripherals, such as SDRAM, flash, SRAM and ROM controllers, LCD controller, 10Mbit/s Ethernet Mac and Phy and a memory management unit. The integrated MMU, 8kbit cache and 48kbit SRAM make it suitable for designers wishing to incorporate open source operating systems, such as Linux.
Sequoia
Tel: 0118 976 9000
www.sequoia.co.uk

Multichip package for stepper motor driver

The TA84002F stepper motor driver from Toshiba Electronics uses a multichip module to improve thermal performance and reduce board space. For driving both windings of a two-phase stepper motor, the device incorporates a main bipolar chip and four p-channel Mosfets within a surface-mount HSOP20 package measuring 16.5 by 12.0 by 2.7mm. It is a chopper-type stepper motor driver with full-step or half-step variation by internal PWM current control. It has built-in thermal protection circuitry and requires no additional external diodes.
Toshiba Electronics
Tel: 01276 694730
www.toshiba-europe.com

1GHz connectors with integral ground plane

Samtec connectors with an integral ground plane to support applications running at up to 1GHz have been introduced by Deltron Roxburgh. The Flexspeed QTE interfaces include edge-mount and elevated designs with 16, 19

and 22mm board spacing. With versions from 40 to 200 I/Os, the interfaces terminate on 0.8mm pitch. The ground plane controls impedance at high frequencies. The connectors are tested for 50 and 70Ω systems and are supported by compliance data including VSWR, attenuation, crosstalk, propagation delay and rise time

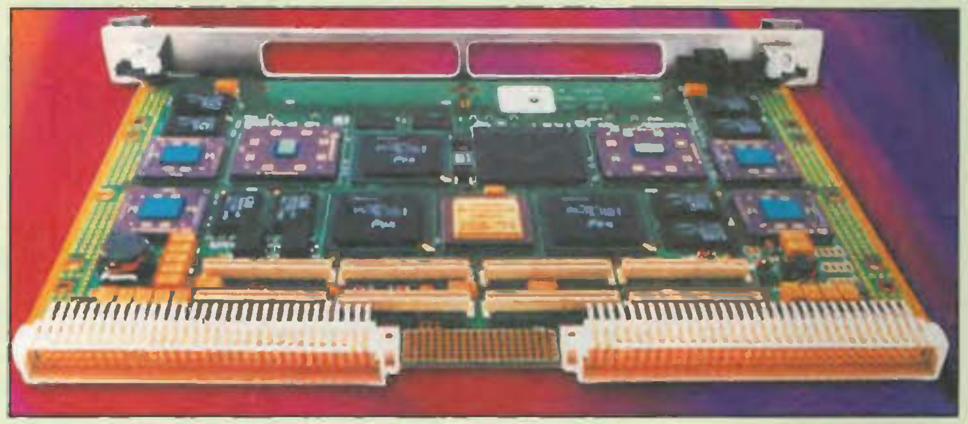
Surface-mount C and pi EMI filters

Surface mount EMI C and pi filters from Flint are available for networking, telecoms and military applications. The Syfer SBSG filters measure 5.25 by 3.2mm with current ratings of 5 and 10A for the pi and C sections respectively. Weighing 0.2g, they are available in 1 to 220nF capacitances with a tolerance of ±20 per cent. Capacitances up to 47nF are rated at 500V DC and operate from –55 to +125°C. They are for use with pick-and-place machines. A wet process multilayer technology is used.
Flint Distribution
Tel: 01530 5103333
www.flint.co.uk

at frequencies from 10MHz to 1GHz. An alternative board layout is provided for designers to improve electrical performance by using a ground plane embedded in the PCB. The interfaces can be arranged in up to five banks, with two rows of 20 pins per bank, providing up to 200 ways. The headers mate with Samtec's QSE sockets.
Deltron Roxburgh
Tel: 01724 408899
www.deltron.roxburgh.com

PowerPC vme dsp board for radar

Ixthos has introduced a VME DSP board combining Motorola's 7410 PowerPC Risc processors with Altivec and Ixthos' Champ common heterogeneous architecture for multi-processing. The board is suitable for processing real time radar, sonar, telecoms and video data. Running multiple, real-time operating systems such as Linux and VxWorks, the AV board can process more than 500 broadband-integrated video, voice and data channels in one 6U VMEbus slot. It can use up to four of the Risc microprocessors, each running at 500 to 600MHz. To facilitate high-speed data movement and predictable maths algorithm execution, it includes cache and local and global memories, seven on-board intelligent DMA controllers and a user-programmable interrupt multiplexer. There is support for two PMC I/O modules. Each PMC I/O expansion site is coupled to a dual processor cluster and provides a dedicated processing compute node to real world I/O. The architecture enables the development of systems based on two scalable, redundant signal processing elements where each contains two MPC 7410 PowerPC processors running at 450 to 600MHz. Each has access to over 2Mbyte of L2 cache, 128Mbytes of 64-bit-wide local SDRAM and a dedicated PMC site.
Ixthos
Tel: 001 703 779 7800
www.ixthos.com



Chip set for optical STM-64 networks

PMC-Sierra has introduced an OC-192 and STM-64 line rate optical networking architecture that lets ISPs and telecoms carriers deliver voice, video and data services across one optical transport backbone, reducing interoperability and scalability challenges. For broadband Internet infrastructure equipment for metropolitan area networks, the Chess-II channeliser engine for Sonet and SDH chip set enables optical transport functionality, sub-wavelength cross-connection and network service aggregation. The chip set contains the PM5317 Spectra-9953 STS-1 channelised framer and pointer processor, PM5307 TBS-9953 groomer and serialiser, PM5374 TSE-160 160Gbit/s STS-1 cross-connect and PM5395 CRSU-4x2488 four time OC-48 clock and data recovery chip. The architecture enables consolidation of traditional Sonet and SDH equipment, such as add-drop multiplexers, terminal multiplexers and broadband digital cross-connects, with IP routers and multi service switches. It helps remove MAN bottlenecks by aggregating services such as Gigabit Ethernet, IP, fibre channel and ATM into scalable Sonet and SDH platforms exhibiting sub-wavelength cross-connect capabilities. Services in the MAN are initially aggregated into pipes at STS-1 51.84Mbit/s level granularities. The chip set grooms these STS-1 pipes in such a way that multiple services can be transported over individual OC-192 and STM-64 wavelengths or multiple OC-48 and STM-16 wavelengths. This capability lets carriers manage and direct services through the MAN.
PMC-Sierra
Tel: 001 604 415 6065
www.pmc-sierra.com

Mobile-radio tester for Bluetooth

With its Bluetooth option for the CMU200, Rohde & Schwarz can provide a mobile-radio tester adding Bluetooth capability to standards such as GSM, Amps, TDMA and CDMAone. The multistandard platform can test several standards in parallel. Extra or future mobile-radio standards can be installed or added in parallel to Bluetooth. With two independent measuring and signalling units, it has a frequency range from 10MHz to 2.7GHz. Temperature controlled error correction increases absolute accuracy by a factor of three. Digital signal processors and parallel measurements reduce testing time by a factor of ten. Power consumption is 130W. Air flows only along those parts and components that produce heat.
Rohde & Schwarz
Tel: 01252 811377
www.pmc-sierra.com

DC/DC converter for industrial use

Insight Memec is offering 66 to 154V inputs on the CWPN20R DC/DC converter from C&D Technologies for industrial applications. The unit is available in two versions with outputs of 24 or 5V, providing power ratings of 15 and 20W respectively. Using forward converter topology with a 300kHz switching frequency, the converter achieves an efficiency of typically 82 per cent. It incorporates full regulation down to zero load, soft-start, undervoltage lockout, short-circuit protection, remote on-off function, internal temperature shutdown facility and overcurrent protection. Internal input filtering reduces conducted noise, while the unit's output filtering improves line and load regulation. Case temperature is specified as –40 to +100°C. Fabricated only with surface-mounted components, it has a standard pin-out and is housed in an aluminium shell measuring 50.6 by 40.4 by 10.0mm.
Insight Memec
Tel: 01296 330061
www.insight.uk.memec.com

Please quote *Electronics World* when seeking further information

Embedded computer with 800MHz Pentium

The Microspace MSMP3SEN and MSMP3SEV from Digital-Logic are PC/104 boards based on an 800MHz Pentium III processor. Features include 16 to 128Mbyte expandable SDRAM, 256kbyte L2 cache and a compact flash memory with up to 128Mbyte capacity. They have USB and 10/100baseT Ethernet interfaces. Applications are in navigation systems, telecoms, computer periphery, medical and measurement technologies, aviation, aerospace, automotive electronics and Internet terminals. There are an integrated RTC battery, support for EEPROM and external IrDA ports and system interfaces for PS/2 keyboard, PS/2 mouse, hard disk and floppy drive. The board contains a watchdog, RS232 ports for coms one and two and a printer interface. The MSMP3SEV has an SXGA-69030 video controller and 4Mbyte video RAM. The modules operate with a 5V, 2.5A power supply.

Digital Logic
Tel: 00 41 32 681 5800
www.digitallogic.com

Mosfet with built-in temperature sensors

The MG400A2YM60 silicon n-channel power Mosfet from Toshiba has a thermistor to detect case temperature and an integrated thermal sensor diode to monitor die temperature. For use in cars, the 80V, 400A module is primarily aimed at starter-motor applications within 42V systems. The device is fabricated using trench-gate technology in a half-bridge configuration, delivering an $R_{ds(on)}$ of typically 1mΩ at 25°C, with an inductance of less than 10nH between the p and n power terminals. the module will operate in ambient temperatures from –40 to +125°C. Equipped with Amp Cap automotive connectors, the device measures 106 by 48 by 34.2mm.

Toshiba Electronics
Tel: 01276 694730
www.toshiba-europe

Off-line switcher IC produces up to 23W

Power Integrations has announced an off-line switcher IC. The Tinyswitch-II outputs up to 23W, or 15W with a universal input, making it suitable for standby power requirements. It integrates a power Mosfet, oscillator and control circuitry on a CMOS chip. There are eight versions from 4 to 23W. Features include auto-restart for short circuit, open loop fault protection, frequency jittering for EMI filtering, current limit and thermal protection, programmable line undervoltage detection preventing power on-off glitches, and circuitry to cut audible transformer noise. Internal switching frequency is 132kHz, allowing the use of EF12.6 or EE13 cores. Ecosmart technology cuts energy consumption in standby and no-load conditions. It can cut the no-load power consumption in a 5W AC adapter to under 125mW with a 115V input.

Power Integration
Tel: 001 408 14 9200
www.powerint.com

Slider aids connector retention

Omron's XF2L FPC connectors occupy 18.9 by 3.5mm for 30 pins. The base sits flush with the board, so the connector footprint can be used to route tracks. Height is 1.2mm without the need for drilling or gouging the PCB. The slider construction has a guide to restrict up and down movements and a slider hold to fix the guide securely. The rear of the slider guide, which restricts play when opening and closing the slider, clicks when the slider is opened. They come in upper or lower contact types with four to 30 poles for 0.5A at 50V DC. They can withstand 250V AC for a minute and operate over the -30 to +85°C range. Solderable hold-downs improve board retention of the connector without drilling the PCB.

Omron
Tel: 020 8450 4646
www.omron.co.uk

Bus hold gives input option

Fairchild Semiconductor has announced additions to its LCX logic range for 3.3V applications. Made on CMOS, they provide overvoltage tolerance, 5V tolerant inputs and outputs, 24mA output drive, low static and dynamic

3G processor core supports CDMA2000

LSI Logic has announced its latest baseband processor core for 3rd generation (3G) mobile communications designs based on the CDMA digital radio technology.

The CBP4.0 is the latest member of the firm's SignalSphere wireless family. It is a single-chip CDMA baseband processor which supports the 3G standard, CDMA2000-A 1xRTT.

The processor is backward compatible to the IS-95B standard. Based on LSI Logic's G12 0.18-micron (0.13-micron L_{eff}) 1.8V CMOS technology, the CBP4.0 uses hardware accelerators to reduce the power consumption by using advanced clocking control, multiple sleep modes, and power management of on-chip analogue blocks, said the company.

The company claims that the CBP4.0 almost doubles the standby time of previous-generation products through the use of its Quick Paging Channel and ultra-low chip quiescent current.

The CBP4.0 has all of the baseband processing logic required to support CDMA2000-A cellular, PCS and AMPS mobile stations, said the firm.

It is available in a 280-pin chip scale package (CSP) measuring 16 mm by 16 mm.

LSI Logic
Tel: 01344 413209
www.lsilogic.com

Please quote *Electronics World* when seeking further information

power. The mixed-voltage capability lets designers use 5V components. The LCXH has bus hold circuitry to give users the option of holding the inputs at the last known state. This eliminates the need for pull-up resistors to prevent the device inputs from floating and causing potential system errors. Several have power up and down circuitry – the LCXZ line. These provide the option of having a buffer that will power up and down into a high impedance state. This adds an extra level of system integrity and predictability by defining a known state during the power sequencing. They are for telecoms and networking equipment. The LCXH and LCXZ have the option for 26Ω series resistors on the outputs to reduce noise.
Fairchild Semiconductor
Tel: 00 331 563 7210
www.fairchildsemi.com

Card PC system starter kit

For embedded applications, Gothic Components is offering a starter kit that comes programmed and configured for immediate use. Based on Epson's Cardsystem-586, the kit is a modular system consisting of a single-board computer with a Card-PC slot and built-in 16Mbyte Card-586 running at 133MHz. The Card-586 provides a complete PC system, requiring the addition of a monitor, keyboard and mouse. The kit incorporates two PCMCIA slots and MS-Romdos 5.0, which is preinstalled into the bios flash of the Card-586 to start the PC. Also available is a 68-pin PCMCIA Seata ADB card, with software preinstalled, giving a link for data exchange through the serial cable with another PC. Cables and manuals are included.
Gothic Components
Tel: 0118 978 7848
www.gothiccomponents

In-system PLD programming via web

The release 3.0 of Cypress Semiconductor in-system reprogrammable software is available on its web site. ISR release 3.0 puts in place all of the software building blocks needed to support the company's Delta39K family of CPLDs. In addition to support for Delta39K CPLDs, ISR release 3.0 adds a layer of capability by supporting the JEDEC standard test and programming language (STAPL) for programmable logic. STAPL is a vendor- and platform-independent language for programming devices using the IEEE-standard 1149.1 TAP controller, (JTAG), which allows the specification of both the programming data and programming algorithm in a single file.
Cypress Semiconductor
Tel: 0032 2357 0220
www.cypress.com

LCD driver for mobile phones

STMicroelectronics has introduced an LCD driver for mobile phones. Called STE2000 and STE2001, the chips integrate a complete controller/driver for 65x128 column black and white graphic liquid crystal displays (LCDs). Low power consumption means the chips are suitable for compact battery powered products including all types of mobile, music players and portable games, said the firm. The devices are also suitable for chip-on-glass and chip-on-flex assembly techniques, which are now the mainstream in these applications.

They contain all of the circuits needed to control and drive displays up to 65 by 128, including graphics functions and a selectable interface for connection to I2C, serial or parallel ports. No external components are required. The two types differ in the mechanical layout of the bonding pads, allowing flexibility in assembly of the display.
STMicroelectronics
Tel: 01628 890800
www.st.com

Sunlight-visible dot-matrix displays

Forge Europa has announced its sunlight visible LED dot matrix displays. The displays are encapsulated with water clear epoxy for the outdoor display market. Package styles available are 18 to 60mm character height in 5x7 and 8x8 dot configurations. Wavelengths include 632nm red, 591nm yellow and 502nm traffic light green. For true sunlight visibility there is the introduction of a 60mm 8 by 8 display having two chips per dot, with a 120° viewing angle. Most modules have industry standard pin configurations, but custom pin-outs can be made.
Forge Europe
Tel: 01229 580000
www.forge-europe.co.uk

Microcontroller with 116kbyte ROM

Epson has developed an 8-bit microcontroller with an integrated 116kbyte of ROM. This makes the E0C88816 suitable for large programmable and multilingual applications, such as battery driven metering and time management devices with country specific parameter settings. A 4224-bit display memory and a 512byte melody RAM lets the remaining 8kbyte be available for the program-dedicated variables and stack. Melody functionality is for indicating events and status visually and audibly. This microcontroller includes an LCD controller and driver for segment and dot matrix displays. Integrated booster and regulator circuits improve contrast for low-end and high-end LCD panels. A software-controlled contrast improves the information visibility on the display. There is a choice between die form and QFP.
Epson Electronics
Tel: 00 49 89 14005277
www.epson-electronics.de

CPU suits electrically noisy uses

Microchip has introduced a Picmicro microcontroller that provides an 18V 140mA communication interface and a high voltage interface for electrically noisy environments. The PIC16C433 is for consumer and industrial connectivity applications and suits domestic appliances, industrial controls and security and sensor applications. It includes a four-channel 8-bit A/D converter and an 8-bit real time clock and counter with an 8-bit programmable prescaler. It provides up to 2k words of OTP memory, 128byte of data RAM and 35 single-cycle instructions. The device has built-in Lin bus protocol capabilities and comes

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in an 18-pin package. Features include in-circuit serial programming, power-on reset, power-up timer, oscillator start-up timer, sleep mode, selectable oscillator options, watchdog timer with its own on-chip RC oscillator and transceiver thermal shutdown for device protection.
Microchip
Tel: 0118 921 5858
www.microchip.com

PCI Interface, UARTs and parallel port

For PCI bus interfacing, the OX16PC1952 from Oxford Semiconductor brings together a PCI interface, two serial channels and a parallel port in a 128-pin TQFP. With dual 15Mbit/s 16C950 Uarts and an IEEE1284 port, this bridge device provides an upgrade path to multi-function add-in expansion cards. In asynchronous mode with 128byte deep transmit and receive Fifos, higher data rates are possible. The Fifos help reduce CPU overhead and increase driver efficiency, such that a baud rate up to 60Mbit/s is achieved in external 1x clock mode. Each channel is software compatible with 16C550 type devices and the firm's OX16C950 Uarts. Features include automated in-band flow control and readable Fifo levels. The parallel port supports the Centronics interface standard. This 32-bit, 33MHz PCI interface chip is compliant with version 2.2 of the PCI bus specification and version 1.0 of the PCI power management specification. All default register values can be overwritten using a serial EPROM. The internal Uarts have multi-port features including shadowed Fifo fill levels and global interrupt source register.
Oxford Semiconductor
Tel: 01235 824900
www.oxsemi.com

Chip resistors get hot and cold

Rohm Electronics has launched miniature chip resistors for operation from –55 to +125°C that deliver fully rated performance to +70°C. The MCR03 precision thick film rectangular chip resistors are supplied in 1608 (0603) packages with dimensions of 1.6 by 0.8 by 0.45mm. Resistance values are between 100Ω and 1MΩ. Resistance tolerances are either ±5 or ±1 per cent. The devices have a power rating of 0.063W and a temperature coefficient of ±50ppm. Resistors are rated for a maximum continuous working voltage of 50V and a maximum overload voltage of 100V. The devices are supplied on tape-and-reel.
Rohm Electronics
Tel: 01908 282666
www.rohm.co.uk

Second generation 3.3V PLDs

Lattice Semiconductor has completed the production release of its second-generation Superfast BFW ISPLSI 2000VE family. The 3.3V in-system programmable Complex PLD family, comprised of five devices, provides logic densities from 32 to 192 macrocells. The fastest device, the ISPLSI 2032VE, raises the PLD speed-bar to 300MHz with 3ns pin-to-pin speeds. Input-to-clock setup times and clock-to-output delays are both 2.0ns. They use submicron E^2CMOS technology. Packages range from PLCC, PQFP and TQFP to 1mm ball pitch BGAs. They are boundary scan testable and comply with IEEE1149.1.
Lattice Semiconductor
Tel: 01932 582941
www.latticesemi.com

LCD controller for hand-held computer

Handspring has selected the Epson SED1376 LCD controller for its Visor Prism colour hand-held computer. It lets the computer display 65 536 colours and can interface directly with the LCD panel technology.
Features include programmable colour-depth and display resolution. It has 90, 180 and 270° HW-rotation and picture-in-picture-plus.
Epson Electronics
Tel: 0049 89 140050
www.epson-electronics.de

Compact PCI platforms for 4U

The G0405 and G0406 are compact PCI platforms from IBus Phoenix for use in 4U enclosures for telecoms, broadcast, telephony, industrial and medical OEMs. They suit applications with multi-system clustering capabilities. Enclosures are supplied with an H.110 passive backplane that supports hot-swap I/O cards.
Ibus Phoenix
Tel: 0239 242 4800
www.ibus.co.uk

Codec for appliance terminal adapters

From Dip is the AK2306 PCM codec in 5.0 or 3.3V variants. This 24-pin device combines built in +6 to –18dB transmit and receive volume controls, a 16 to 20Hz ring tone generator and a 9.8 by 7.6mm pin-to-pin VSOP24 package. It achieves idle channel noise levels of less than 0dBrnc0 with the 5V version and less than 5dBrnc0 with the 3.3V device. Features include a power down, mute and time slot swap on the PCM I/F. Applications include ISDN, TA and other home

31cm LCD with four backlights

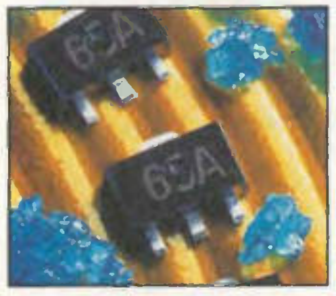
A 31cm SVGA colour TFT LCD module from Toshiba Electronics, the LTM12C285, is an 800 by 600 full-colour display with a typical luminance of 500cd/m^2. It incorporates four replaceable CCFL backlights, each with a specified lifetime of 50 000hr to half brightness. Applications include ATMs, information kiosks, POS terminals, industrial control panels and brightness-diminishing touch panels. The display can be used with 6 or 12 o'clock optimum viewing directions due to a built-in reverse-scanning function. It displays 262 144 colours and has vertical and horizontal viewing angles of 100 and 120° respectively. Contrast ratio is 250:1. Measuring 278.3 by 216.0 by 13.5mm, it has an active display area of 246.0 by 184.5mm and weighs about 900g.
Toshiba Electronics
Tel: 01276 694730
www.toshiba-europe.com

Please quote *Electronics World* when seeking further information

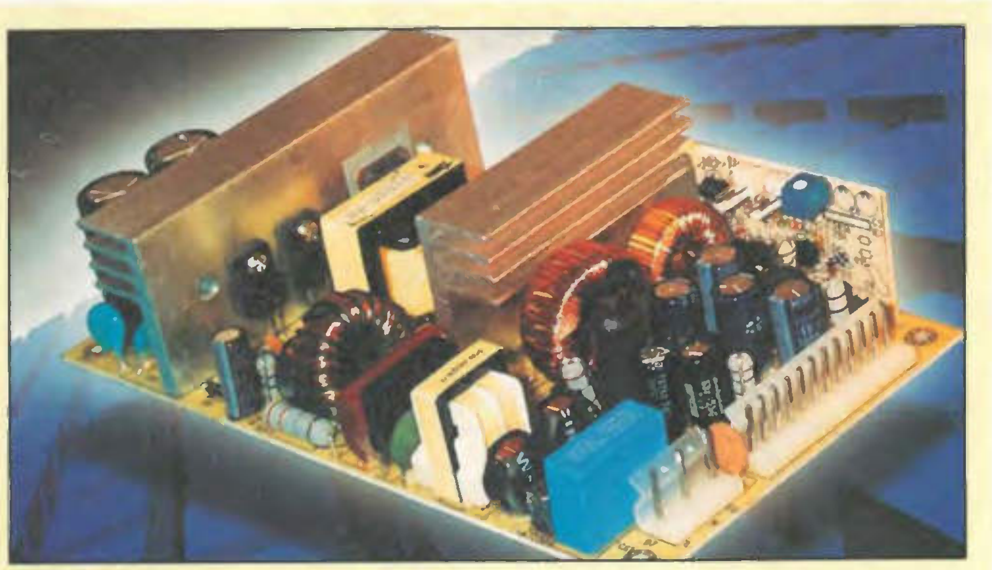
telecoms equipment such as home gateway units.
Dip
Tel: 01223 462244
www.dipInternational

Medium power n-p-n transistor for telecoms

With a V_{ceo} of 400V and a power dissipation of 1W, an n-p-n transistor from Zetex has useful gain up to 200mA. The FCX658A transistor comes in an SOT89 outline and provides



a switching capability for telecoms applications such as dialler circuits and hook switches. It can protect line switching in subscriber line interface circuits (SLICs) and modem circuits. Other applications are predrivers within HID lamp ballasts, high voltage motor drives and circuit breakers.
Zetex
Tel: 0161 622 4444
www.zetex.com



120W switch-mode power supply

Lambda has introduced a switched-mode power supply, the SC120, that satisfies global safety regulations and the latest EC directive on harmonic distortion. Various AC output options make it suitable for computer peripherals, point of sales and telecoms equipment. There are 11 standard models with either single, dual, triple or quad outputs including the 3.3 and 5V combinations. The open-frame unit measures 175 by 100 by 45 mm. Input voltage is 90 to 264V at 47 to 63Hz. It can deliver 110W with forced air cooling and a peak power output of 120W in all modes. Its line and load regulation is ±2 per cent on single output and from ±3 to ±5 per cent on multiple output versions. Operational range is 0 to +50°C. It carries type approval for UL1950, SA950, EN60950 and EN6100-3-2.
Lambda
Tel: 01271 856666
www.lambda-gb.com

Package for wireless handsets

Conexant Systems has introduced a multislot global package that combines all hardware and software functions required to build a GSM and GPRS handset. The hardware portion implements digital and analogue baseband processing, multiband and multislot power amplification, power management and transceiver functions. It uses a direct-conversion architecture, which eliminates intermediate-frequency conversion steps. It includes a worldwide, operator-qualified, network-approved protocol stack, and all other software for a complete Internet-enabled handset. Design support services and tools are provided. Integrated web-browsing software uses the Pixo wireless-phone operating system, microbrowser, application software and developer tools.
Conexant Systems
Tel: 01344 486444
www.conexant.com

Dual phase-locked loop frequency synthesisers

Three dual phase-locked loop frequency synthesisers from Fujitsu are for mobile phones, PDAs and other wireless communications products. The MB15F78UL PLLs support operating frequencies from 50MHz to 2.6GHz. Made on 0.35μm BiCMOS, the MB15F72UL has maximum operating RF and IF of 1300 and 350MHz respectively, while consuming 2.5mA at 2.7V. The MB15F73UL consumes 3.2mA at 2.7V and can operate at 2250 and 600MHz maximum RF and IF. The MB15F78UL has operating frequencies up to 2600 and 1200MHz and consumes 4.5mA at 2.7V. V_{cc} is 2.4 to 3.6V. The stable charge pump circuit is capable of dual current switching through serial data transfer. There are two packages. The thin bump chip carrier package BCC20 has an area of 12.24mm² and is 0.6mm high. The other is a 20-pin TSSOP L-leaded package.
Fujitsu
Tel: 01628 504600
www.fujitsu.fme.com

DC-to-DC converter for the rails

With an input range of 66 to 154V, the CWPN20R DC-to-DC converter from C&D Technologies is for applications in the rail industry. The unit comes in two versions with outputs of 24 or 5V, providing power ratings of 15 and 20W respectively. Using forward converter topology with a 300kHz switching frequency, it achieves an efficiency of typically 82 per cent. The device incorporates full regulation down to zero load, soft-start, undervoltage lockout, short-circuit protection, remote on-off function, internal temperature shutdown facility and overcurrent protection. Internal input filtering reduces conducted noise, while the unit's output filtering improves line and load regulation. Case temperature range is –40 to +100°C. Fabricated with surface-mounted components, it has a standard pin-out and is housed in an aluminium shell measuring 50.6 by 40.4 by 10.0mm.
C&D Technologies
Tel: 00353 474133
www.cdpowerelectronics.com

Video test generator

Based on a fast microcontroller, Roy Harding's video generator produces ten different test patterns. It is switchable for either PAL or NTSC and generates both S and composite video.

In the October 1999 issue of *Electronics World* I described how to construct an SVGA generator for computer monitor testing using a single fast micro-controller. The design proved to be popular with engineers and enthusiasts, so I decided to do another design.

The new design is based on the same micro that provides standard TV test signals. The Scenix device can operate at 50MHz but this design only requires it to run at a more modest 20MHz in turbo mode.

Only two ICs, the Scenix microcontroller and an Analog Devices RGB-to-video converter are used. The unit gives S Video and Composite outputs and can switch between PAL and NTSC. There are ten different test patterns available, among them a test card, colour-bars, grey scale, crosshatch, centring and colour and flashing screens. All waveforms are fully interlaced.

Signal formats

All of the timing signals originate in the microcontroller.

A 20MHz crystal provides the frequency reference for the software routines. These routines create waveforms as close as possible to the ITU-R BT.470-5 Conventional Television recommendations.

An example of the timings of a PAL video signal is shown in **Fig. 1**. I have implemented the NTSC signals at 525 lines, 60Hz for USA standards.

The microcontroller has just two buttons controlling its functions. One is for stepping through the waveforms, the other for switching NTSC/PAL output. Two LEDs indicate what type of output is active, these being linked to the mode pin of the converter chip.

An Analogue Devices AD772, **Fig. 2**, makes short work of all the digital-to-analogue conversion. Using either the 3.579 or 4.43MHz crystal for NTSC/PAL, it needs no external filters or delay lines like some older devices. It also uses few external components.

The generator gives a 2V signal into a unterminated load or a 1V output into a 75Ω load.

Circuit details

Figure 3 shows the complete circuit. Power is regulated at 5V with a standard 7805 positive regulator. A switch-type input connector allows for alternative battery power. I have found the generator to consume 90mA, giving several hours use from a set of five 800mAh AA-sized rechargeable cells.

The circuit is made up in two sections – a digital section provided by the micro-controller and an analogue section around the RGB converter.

On the AD722 chip, there are separate power and ground pins between its digital and analogue sections. Careful layout design is needed to get the best out of this device. Place separate decoupling capacitors directly by chip power pins.

You need to place a 0.1μF decoupling capacitor next to the regulator and another decoupling capacitor and electrolytic as close as possible to the Scenix chip power and ground connections.

The microcontroller's oscillator uses a 20MHz crystal, which must be a fundamental frequency type. The RGB output from the Scenix micro is full CMOS level and must be attenuated to give a 0.7V signal into the AD722.

Two crystals on the converter are controlled by switching the Scenix RB5 port and the 4066 analogue switch. The mode pin for NTSC/PAL is controlled by RB2, which also operates one of the LEDs.

Grey scale output is achieved by shorting the red, green and blue signals via the 4066. Then by switching combinations of outputs plus the extra load resistor on the green divider, you can build up a useful grey scale.

Software

The whole design revolves around software running on the Scenix microprocessor. Timings for the functions are critical. Each scan line must be balanced to within a couple of instructions. If it isn't, the image will be distorted.

Button presses are checked on each frame. A software trap waits for the button to be released for the changes to take place. The first button toggles the displayed output from PAL to NTSC and the second changes the display patterns by subsequent presses of the button.

I have commented the software listing to show the relevant sections of code.

The main software loop is quite small consisting mainly of the interlaced frame building blocks for the NTSC and PAL formats. All of the display patterns are subroutines called twice from within the main loop. As both fields are fed with the same data, flicker of the display is illuminated.

If you don't need the NTSC mode, you can leave off the 3.579MHz crystal, D_2, R_{15} and the B_1 button as the software defaults to PAL mode on power-up.

Components list

Resistors 1/8 watt 5% carbon film

R_1, R_4, R_6	10k
R_7, R_8, R_9	1kΩ
R_{10}, R_{11}, R_{12}	150Ω
R_{14}, R_{15}	220Ω
R_3, R_5, R_{13}	75Ω
R_2	10Ω
R_{16}	270Ω

Capacitors 50V polyester or ceramic or 10-16V electrolytics.

C_1, C_3, C_6, C_9, C_5	0.1μF
C_2, C_{10}	10nF
C_7, C_8, C_{12}	15pF
C_4, C_{11}	100μF

B_1, B_2	push buttons
Y_1	4.433MHz crystal for PAL
Y_2	20MHz fundamental crystal
Y_3	3.579MHz crystal for NTSC
P_7	3.5mm power connector
IC_1	Scenix SX18/AC/DP 18pin 50MIPS processor,
IC_2	4066
IC_3	AD722 Analogue Devices (only available in SMT)
IC_4	7805 5V positive regulator
D_1, D_2	3mm red LEDs
S_4	Phono socket
S_1	4 pin mini din

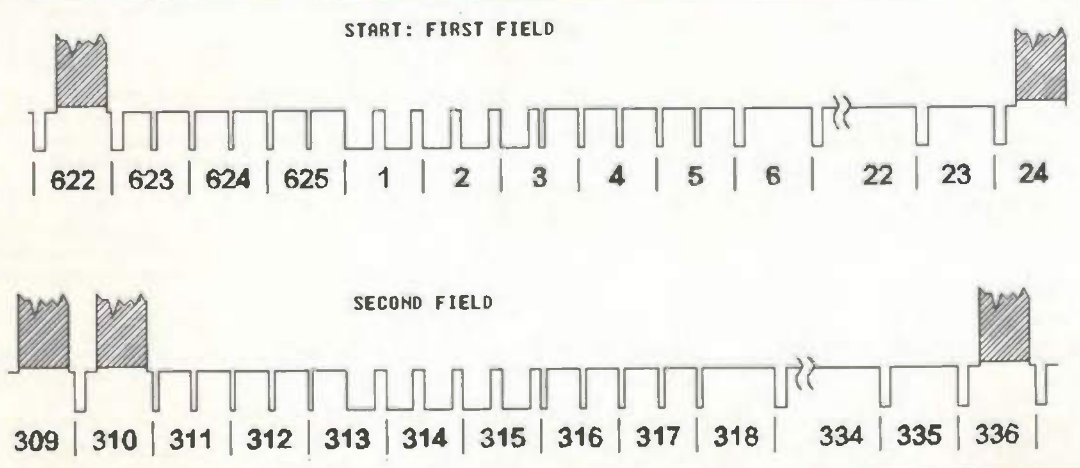


Fig. 1. Timings for a PAL video signal highlighting differences between the first and second fields.

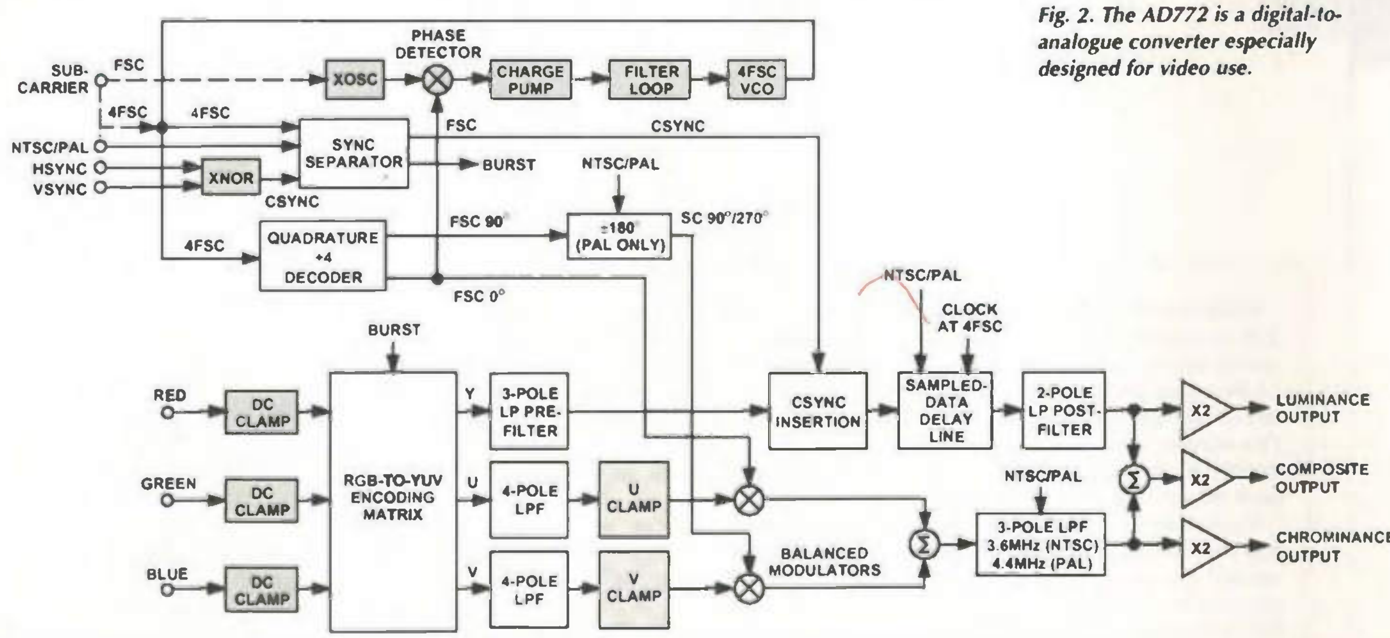


Fig. 2. The AD772 is a digital-to-analogue converter especially designed for video use.

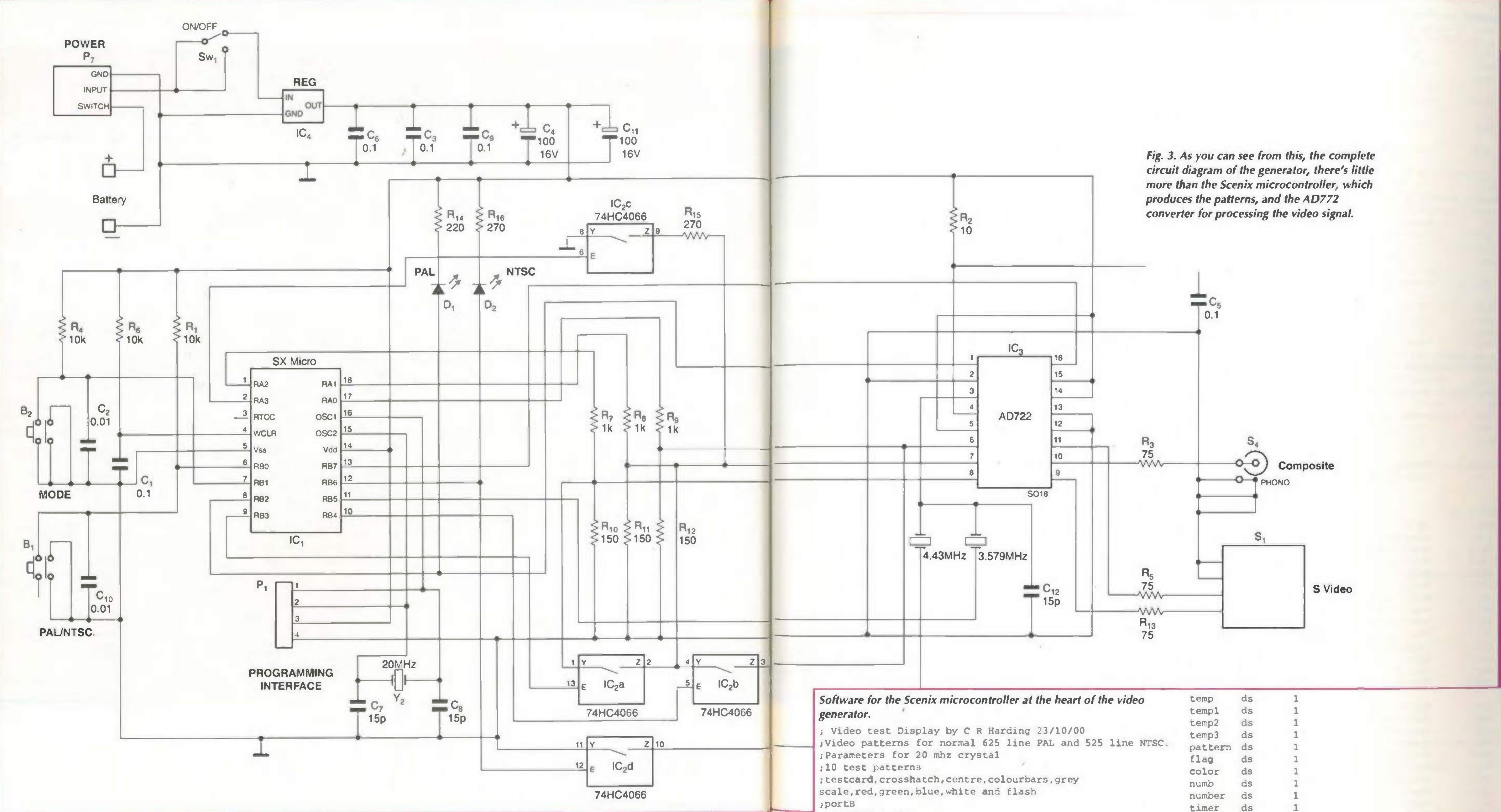


Fig. 3. As you can see from this, the complete circuit diagram of the generator, there's little more than the Scenix microcontroller, which produces the patterns, and the AD772 converter for processing the video signal.

Button checks are enclosed in the main loop as any interrupt routines badly distort the display image.

I have used port A for the RGB drive signals and the sync signal on port B. This enables you to send colour values directly to the port without having to mask the sync bits out of the equation.

If you want to add or change any display patterns to the generator, it is important that you match your line length timings with the line length already used by the other routines.

Note that if you want to re-program the Scenix micro, a special Parallax SX-KEY serial-programming adaptor is required. This programming device plugs onto P_1, the 4-way connector next to the micro. You will also need to remove the crystal if you want to run in debug mode. ■

Technical support

Parts, for this design including pre-programmed chips and PCBs can be obtained from the author. Please send an SAE marked clearly with the words 'Test Generator' to Electronics World, Anne Boleyn House, 9-13 Ewell Road, Cheam, Surrey SM3 8BZ to obtain a list.

Software for the Scenix microcontroller at the heart of the video generator.

```
; Video test Display by C R Harding 23/10/00
;Video patterns for normal 625 line PAL and 525 line NTSC.
;Parameters for 20 mhz crystal
;10 test patterns
;testcard,crosshatch,centre,colourbars,grey
scale,red,green,blue,white and flash
;portB
;port bit 7 sync
;port bit 6 ntsc/pal led
;port bit 5 ntsc crystal control
;port bit 4 GREEN/RED SHORT ;set bits 3&4 for grey scale
;port bit 3 BLUE/GREEN SHORT
;port bit 2 led/mode ntsc/pal
;port bit 1 Button1 pattern
;port bit 0 Button2 ntsc/pal
;portA
;port bit 3 GREEN LEVEL
;port bit 2 BLUE
;port bit 1 GREEN
;port bit 0 RED
          device   SX18L,oscxt3,turbo
          device   stackx_optionx,protect
          freq     20_000_000
          id       'CRHVideo'
          reset    start
          org 8
count     ds       1
port      ds       1
temp      ds       1
temp1     ds       1
temp2     ds       1
temp3     ds       1
pattern   ds       1
flag      ds       1
color     ds       1
numb      ds       1
number    ds       1
timer     ds       1
standard  ds       1
T1        ds       1
T2        ds       1
T3        ds       1
T4        ds       1
T5        ds       1
T6        ds       1
T7        ds       1
T8        ds       1
T9        ds       1
T10       ds       1
T11       ds       1
          org 100h
start
;setup ports
          mov      !rb,#%00000011  ;port B 6 out 2 in
          mov      !ra,#%00000000  ;port A 4 out
          mov      rb,#0
          mov      ra,#0           ;black
          mov      pattern,#0      ;colour bars
          mov      standard,#0     ;set PAL
```

```
mainloop
        mov     flag,#0         ;reset keypress flag
        mov     T1,#97          ;halfline delay low 25µs
        mov     T3,#28          ;halfline delay high 7µs
        mov     T5,#35          :Colour bar width delay
        mov     T6,#10          ;halfline2 delay low 4µs
        mov     T7,#115         ;halfline delay high
28µs
        mov     T8,#18          ;fullline delay low
4.7µs
        mov     T9,#233         ;fullline delay high
57.3µs
        cje     standard,#1,ntsc
pal
        mov     T2,#56          ;testcard
        mov     T11,#12         ;crosshatch
        mov     T10,#127        ;centre lines
        mov     T4,#144         ;colour bar frame lines
        or      port,#%01000000
        and     port,#%11111011 ;mode leds
        mov     rb,port
        mov     !rb,#%00100011  ;set input bit 5 ntsc
out
        jmp     frameloop
ntsc
        mov     T2,#33          ;testcard
        mov     T11,#10
        mov     T10,#104        ;centre lines
        mov     T4,#121         ;colour bar frame lines
        mov     !rb,#%00000011  ;set out low xtal ntsc
        or      port,#%00000100
        and     port,#%10011111 ;mode leds
        mov     rb,port
frameloop
        mov     temp,rb         ;buttons check
        and     temp,#3
        cje     temp,#3,exit    ;exit if button not
                                ;pressed
        call    @button
        cje     flag,#1,mainloop  ;jump to mainloop if
                                  ;set
exit
;First field sync
        mov     ra,#8           ;test sync signal
        cje     standard,#1,ntscloop
        call    @halfline1      ;line1
        call    @halfline1
        call    @halfline1      ;line2
        call    @halfline1
        call    @halfline1      ;line3
        call    @halfline2
        call    @halfline2      ;line4
        call    @halfline2
        call    @halfline2      ;line5
        call    @halfline2
        mov     temp,#16        ;17 lines
        call    @routines       ;display on screen
line311
        call    @halfline2      ;line311
        call    @halfline2
        call    @halfline2      ;line312
        call    @halfline2
        call    @halfline2      ;line313
        call    @halfline1
        call    @halfline1      ;line314
        call    @halfline1
        call    @halfline1      ;line315
        call    @halfline1
        call    @halfline2      ;line316
        call    @halfline2
        call    @halfline2      ;line317
        call    @halfline2
        mov     temp,#16        ;16 lines add 1 in
        ;display except crosshatch
        call    @routines
line623
        call    @halfline2      ;line623
        call    @halfline2
        call    @halfline2      ;line624
        call    @halfline2
        call    @halfline2      ;line625
        call    @halfline2
        jmp     frameloop
ntscloop
        call    @halfline2      ;line1
        call    @halfline2
        call    @halfline2      ;line2
        call    @halfline2
        call    @halfline2      ;line3
        call    @halfline2
        call    @halfline1      ;line4
        call    @halfline1
        call    @halfline1      ;line5
        call    @halfline1
        call    @halfline1      ;line6
        call    @halfline1
        call    @halfline2      ;line7
        call    @halfline2
        call    @halfline2      ;line8
        call    @halfline2
        call    @halfline2      ;line9
        call    @halfline2
        mov     temp,#10        ;10 lines
        call    @routines
        call    @halfline2      ;line263
        call    @halfline2
        call    @halfline2      ;line264
        call    @halfline2
        call    @halfline2      ;line365
        call    @halfline2
        call    @halfline2      ;line366
        call    @halfline1
        call    @halfline1      ;line367
        call    @halfline1
        call    @halfline1      ;line368
        call    @halfline1
        call    @halfline1      ;line369
        call    @halfline2
        call    @halfline2      ;line370
        call    @halfline2
        call    @halfline2      ;line371
        call    @halfline2
        mov     temp,#11        ;11 lines
        call    @routines
        jmp     frameloop
;subroutines lower half of page
        org 0h
halfline1
        and     port,#%01111111 ;bit 7 low vertical sync
        mov     rb,port         ;write value to port
        mov     temp,T1
floop1  nop
        djnz    temp,floop1
        or      port,#%10000000 ;bit 7 high
        mov     rb,port         ;write value to port
        mov     temp,T3
floop2  nop
        djnz    temp,floop2
        retp
halfline2
        and     port,#%01111111 ;bit 7 low vertical sync
        mov     rb,port         ;write value to port
        mov     temp,T6
floop3  nop
        djnz    temp,floop3
        or      port,#%10000000 ;bit 7 high
        mov     rb,port         ;write value to port
        mov     temp,T7
floop4  nop
        djnz    temp,floop4
        retp
fullline
        call    linesync
        mov     temp2,T9
floop6  nop
        djnz    temp2,floop6
        retp
;4.7us
linesync
        and     port,#%01111111
        mov     rb,port
        mov     count,#18       ;delay loop
        call    @dly
        or      port,#%10000000
        mov     rb,port
        retp
;7 bars red to white
bars
        call    linesync
        mov     count,#17       ;line blanking 5.8µs
        call    @dly
bv4
        mov     ra,#1           ;red
        call    @bardly
```



```
        mov     ra,#2           ;green
        call    @bardly
        mov     ra,#3           ;yellow
        call    @bardly
        mov     ra,#4           ;blue
        call    @bardly
        mov     ra,#5           ;magenta
        call    @bardly
        mov     ra,#6           ;cyan
        call    @bardly
        mov     ra,#7           ;white
        call    @bardly
        mov     ra,#0
        mov     count,#4        ;1.5µs
        call    @dly
        retp
line
        call    linesync
        mov     count,#18       ;line blanking 5.8µs
        call    @dly
        mov     ra,color
        mov     count,#206
        call    @dly
        nop
        mov     ra,#0           ;black
        mov     count,#4        ;1.5µs
        call    @dly
        retp
hatch
        call    @linesync
        mov     count,#18       ;line blanking 5.8uS
        call    @dly
        mov     temp,#16
hl
        mov     ra,#7
        mov     ra,#0
        mov     count,#10       ;delay loop
        call    @dly
        djnz    temp,hl
        mov     ra,#7
        mov     ra,#0
        mov     count,#3        ;blank
        call    @dly
        nop
        nop
        retp
bardly
        mov     count,T5        ;width delay
boop
        djnz    count,boop
        retp
dly
loop2   nop                     ;delay set in count
        djnz    count,loop2
        retp
;page
        org     200h
routines
        cje     pattern,#0,colourbarsA
        cje     pattern,#1,centre
        cje     pattern,#2,crosshatch
        cje     pattern,#3,colourbars
        cje     pattern,#4,grey scale
        cje     pattern,#5,red
        cje     pattern,#6,green
        cje     pattern,#7,blue
        cje     pattern,#8,white
        cje     pattern,#9,flash
crosshatch
blankloopc
        call    @fullline       ;fame blanking
        djnz    temp,blankloopc
        mov     temp2,T11       ;12 squares PAL / 10
                                ;squares NTSC
cra
        mov     color,#7        ;white
        call    @line
        mov     temp1,#23       ;gaps
crb
        call    @hatch          ;hatch
        djnz    temp1,crb
        djnz    temp2,cra
        nop
        call    @line           ;end line
        retp
grey scale
blankloopg
        call    @fullline       ;fame blanking
        djnz    temp,blankloopg
        call    @fullline       ;add extra line except
                                ;crosshatch
        mov     temp2,T4        ;122 ntsc
glpa
        call    @gbars
        djnz    temp2,glpa
        mov     temp2,T4        ;144 pal
glpb
        call    @gbars
        djnz    temp2,glpb
        retp
colourbars
blankloopb
        call    @fullline       ;fame blanking
        djnz    temp,blankloopb
        call    @fullline       ;add extra line except
                                ;crosshatch
        mov     temp2,T4        ;121 ntsc
lpa
        call    @bars
        djnz    temp2,lpa
        mov     temp2,T4        ;144 pal
lpb
        call    @bars
        djnz    temp2,lpb
        retp
colourbarsA
blankloopt
        call    @fullline       ;fame blanking
        djnz    temp,blankloopt
        call    @fullline
                ;add extra line except crosshatch
        call    @cass           ;8 lines
        mov     temp3,T4        ;144 PAL
clpa
        call    @barsA
        djnz    temp3,clpa
        mov     temp3,T2        ;pal 56 / ntsc 33
clpal
        call    @barsA
        djnz    temp3,clpal
        mov     temp3,#4
clpb
        call    @fullline
        djnz    temp3,clpb
        mov     color,#7
        call    @fullline       ;alternate lines
        call    @line
        call    @fullline
        call    @line
        call    @fullline
        call    @line
        call    @fullline
        call    @line
        mov     temp3,#4
clpc
        call    @fullline
        djnz    temp3,clpc
        mov     temp3,#56
clpd
        call    @gbars          ;grey scale
        djnz    temp3,clpd
        call    @cass           ;8 lines
        retp
centre
blankloopct
        call    @fullline       ;fame blanking
        djnz    temp,blankloopct
        mov     temp,#18
ct1
        call    @endsline       ;3 vertical lines
        djnz    temp,ct1
        mov     color,#7
        call    @line           ;top line
        mov     temp,T10
ct2
        call    @endsline
        djnz    temp,ct2
        call    @line           ;centre line
        mov     temp,T10
ct3
        call    @endsline
        djnz    temp,ct3
        call    @line           ;bottom line
        mov     temp,#18
ct4
        call    @endsline
```

```
        djnz    temp,ct4
        retp
red
        mov     color,#1        ;bit0
        jmp     full
white
        mov     color,#7        ;all on white
        jmp     full
green
        mov     color,#2        ;bit1
        jmp     full
blue
        mov     color,#4        ;bit2
full
blankloopf
        call    @fullline       ;fame blanking
        djnz    temp,blankloopf
        call    @fullline       ;add extra line except
                                ;crosshatch
        mov     temp2,T4        ;144 lines pal
wlpa
        call    @line
        djnz    temp2,wlpa
        mov     temp2,T4
wlpb
        call    @line
        djnz    temp2,wlpb
        retp
flash
blankloopfl
        call    @fullline       ;fame blanking
        djnz    temp,blankloopfl
        call    @fullline       ;add extra line except
                                ;crosshatch
        add     timer,#1
        cjae    timer,#50,black
        mov     color,#7
        mov     temp2,T4
flpa
        call    @line           ;white screen
        djnz    temp2,flpa
        mov     temp2,T4
flpb
        call    @line
        djnz    temp2,flpb
        retp
black
        mov     temp3,T4        ;black screen
blpa
        call    @fullline
        djnz    temp3,blpa
        mov     temp3,T4
blpb
        call    @fullline
        djnz    temp3,blpb
        cjbe    timer,#100,no
        mov     timer,#0
no
        retp
;page
        org     400h
endsline
        call    @linesync
        mov     count,#18       ;line blanking 5.8uS
        call    @dly
        mov     count,#14
        call    @dly
        mov     ra,#7           ;white
        mov     ra,#0           ;black
        mov     count,#86
        call    @dly
        mov     ra,#7           ;white
        mov     ra,#0           ;black
        mov     count,#86
        call    @dly
        mov     ra,#7           ;white
        mov     ra,#0           ;black
        mov     count,#14
        call    @dly
        mov     count,#4        ;1.5uS
        call    @dly
        retp
button
        mov     temp,rb         ;press check
        and     temp,#3
        cje     temp,#2,ntscpal
patchange
        add     pattern,#1      ;next pattern
        cjbe    pattern,#9,bexit
        mov     pattern,#0      ;back to start
bexit
        mov     temp,rb         ;press check
        and     temp,#3
        cjne    temp,#3,bexit   ;wait release
        mov     flag,#1         ;set flag
        retp
ntscpal
        xor     standard,#1     ;toggle ntsc/pal
        jmp     bexit
cass                            ;10 line loop
        mov     temp1,#8
casloop
        call    @linesync
        mov     count,#18
        call    @dly
        mov     temp2,#12
casl
        mov     ra,#7           ;white
        mov     count,#6
        call    @dly
        mov     ra,#0           ;black
        mov     count,#7
        call    @dly
        djnz    temp2,casl
        mov     count,#4        ;1.5µs
        call    @dly
        nop
        nop
        nop
        djnz    temp1,casloop
        retp
;7 bars
barsA
        call    @linesync
        mov     count,#18       ;line blanking 5.8uS
        call    @dly
        mov     ra,#7           ;white
        call    @bardly
        mov     ra,#3           ;yellow
        call    @bardly
        mov     ra,#6           ;cyan
        call    @bardly
        mov     ra,#2           ;green
        call    @bardly
        mov     ra,#5           ;magenta
        call    @bardly
        mov     ra,#1           ;red
        call    @bardly
        mov     ra,#4           ;blue
        call    @bardly
        mov     ra,#0           ;black
        mov     count,#4        ;1.5uS
        call    @dly
        retp
;8 grey bars
gbars
        call    @linesync
        mov     count,#17
        call    @dly
        or      port,#%00011000 ;short colours
        mov     rb,port
        mov     ra,#7           ;white
        call    @bardly
        mov     ra,#15
        call    @bardly
        mov     ra,#3           ;lgrey
        call    @bardly
        mov     ra,#11
        call    @bardly
        mov     ra,#1           ;dgrey
        call    @bardly
        mov     ra,#9
        call    @bardly
        mov     ra,#0           ;black
        call    @bardly
        and     port,#%11100111
        mov     rb,port
        mov     count,#2        ;1.5µs
        call    @dly
        nop
        nop
        nop
        retp
```

CIRCUIT IDEAS

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

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

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

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

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

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

£75 winner

Isolated voltage and current measurement for mains powered equipment

Sometimes it is necessary to measure the current and voltage supplied to a variable load. This can be done simply by connecting a voltmeter in parallel with and an ammeter in series with the load.

However, if you want to record the current and voltage passing to the load, with for example a data logger, then things can become a little more difficult. Firstly some form of isolation between the AC mains circuitry and the test equipment is necessary. Secondly, most data loggers require a common ground.

The following simple circuit provides both optical isolation of the test equipment and a common ground.

The part of the circuit at mains potential consists of a dropping resistor, R_1, and diode D_1. These pass current to the reservoir capacitor C_1.

DC voltage on the reservoir capacitor is about 18V. The voltage on the reservoir is then filtered by the network R_2 and C_2 to remove any residual ripple. Resulting voltage is used to bias the photodiodes of the optocouplers OC_1 and OC_2 via the series resistors R_3 and R_4.

Photodiode current is about 1mA in each case. The DC voltage on OC_1 is modulated by an AC voltage derived from resistor R_5, in series with the load, via the resistor R_6 and the capacitor C_3. This modulation signal is thus directly proportional to the current flowing through the load.

DC voltage on OC_2 ,is modulated with an AC voltage derived from the voltage across the load via R_7. This modulation signal is thus proportional to the voltage applied to the load.

On the isolated side of the circuit, the phototransistors of optocouplers OC_1 and OC_2 are connected via load resistors, VR_1 and VR_2, to the positive supply. The output signals are taken from these load resistors via the capacitors C_4 and C_5.

Output from OC_1 is directly proportional to the current through the load and the output signal from OC_2 is directly proportional to the voltage applied to the load. The outputs are grounded via the resistors R_8 and R_9 to remove any DC component from the signals. These signals can be applied directly to the high impedance inputs of a data logger.

Power supply to the isolated part of the circuit is not critical. The phototransistors in the optocouplers are essentially constant current devices and simple stabilisation is provided by ZD_1 and ZD_2 and their series resistors R_{10} and R_{11}.

Current consumption of the circuit is around 2mA – excluding the zener current. With the values shown, the circuit will provide an output signal at OP_1 of about 2.5V with a current through the load of 1A . It will provide a similar output from OP_2 when the voltage across the load is 240V.

To calibrate the current measuring circuit, the current is measured through the load using a multimeter and the variable resistor VR_1 adjusted to give the required output at OP_1. The voltage measuring section is adjusted in a similar way using a multimeter to measure the voltage applied to the load and then adjusting VR_2 to give the desired output at OP_2.

It is possible to adjust the measuring range of the circuit for other values of voltage and current. This is done by changing the series resistor R_5 to alter the current range or resistor R_7 to alter the voltage range. The value of R_5 in ohms is given by $0.5 \div I_{max}$. Here, I_{max} is maximum current, in amps, that it is desired to measure.

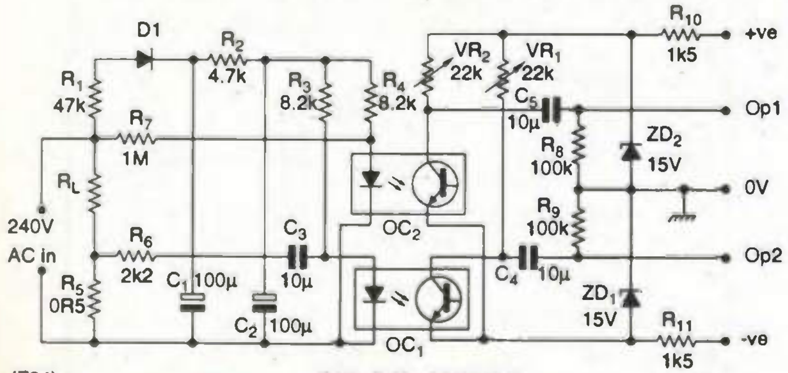
Power dissipation, P, in watts in this resistor is $0.5 \times I_{max2}$ and the power rating of the resistor should be chosen accordingly. The value of R_7 in megaohms is $V_{max} \div 250$ where V_{max} is the maximum voltage, in volts, that it is desired to measure.

Several of these circuits have been constructed and they have proved very reliable, with very good linearity between the input current or voltage and outputs and with almost no crosstalk between the current measurement and the voltage measurement.

One application of the circuit has been to monitor the current and voltage of motors under variable load conditions. Once these values have been logged it is a simple matter to compute the power consumption as the product of the current and the voltage.

Mike Cox
Bankeryd
Sweden

Recorder interface for mains power measurements.



(F34)

OC1, OC2 - SFH618-2

Back-off for capacitance meter

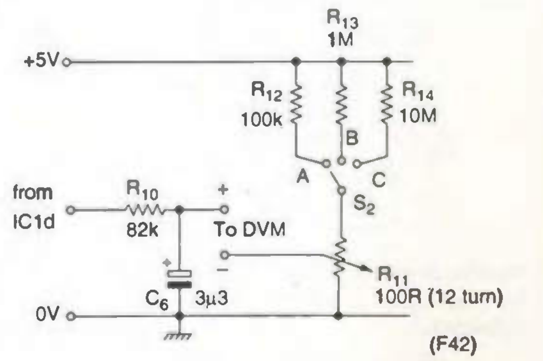
In a recent article by Hickman[1] describing a simple wide-range capacitance meter, the problem of coping with the 'residual reading' (C_o) was mentioned. On the most sensitive range, this value has to be subtracted from the reading displayed on the DVM. For example, when measuring 100pF the DVM reads 130pF, but you then subtract the residual, C_o=30pF.

An easily implemented method for backing-off this error electronically is shown in the Figure. It involves adding 3mV to the negative end of the DVM. Input resistance of most such meters is 10MΩ or more, and so can be ignored in comparison with the value of the 12-turn trimmer potentiometer R_{11}.

On the two less sensitive ranges, called B and C in the original article, proportionately less back-off is required, and so the bias-feed resistor from +5V needs to be increased from 100kΩ to 1MΩ and then 10MΩ.

Switch S_2 can conveniently be ganged with the original S_1 by using a standard four-pole three-way wafer. A similar method can often be applied to other instruments, avoiding the necessity of an additional op-amp to perform the subtraction.

C J D Catto
Cambridge
F42

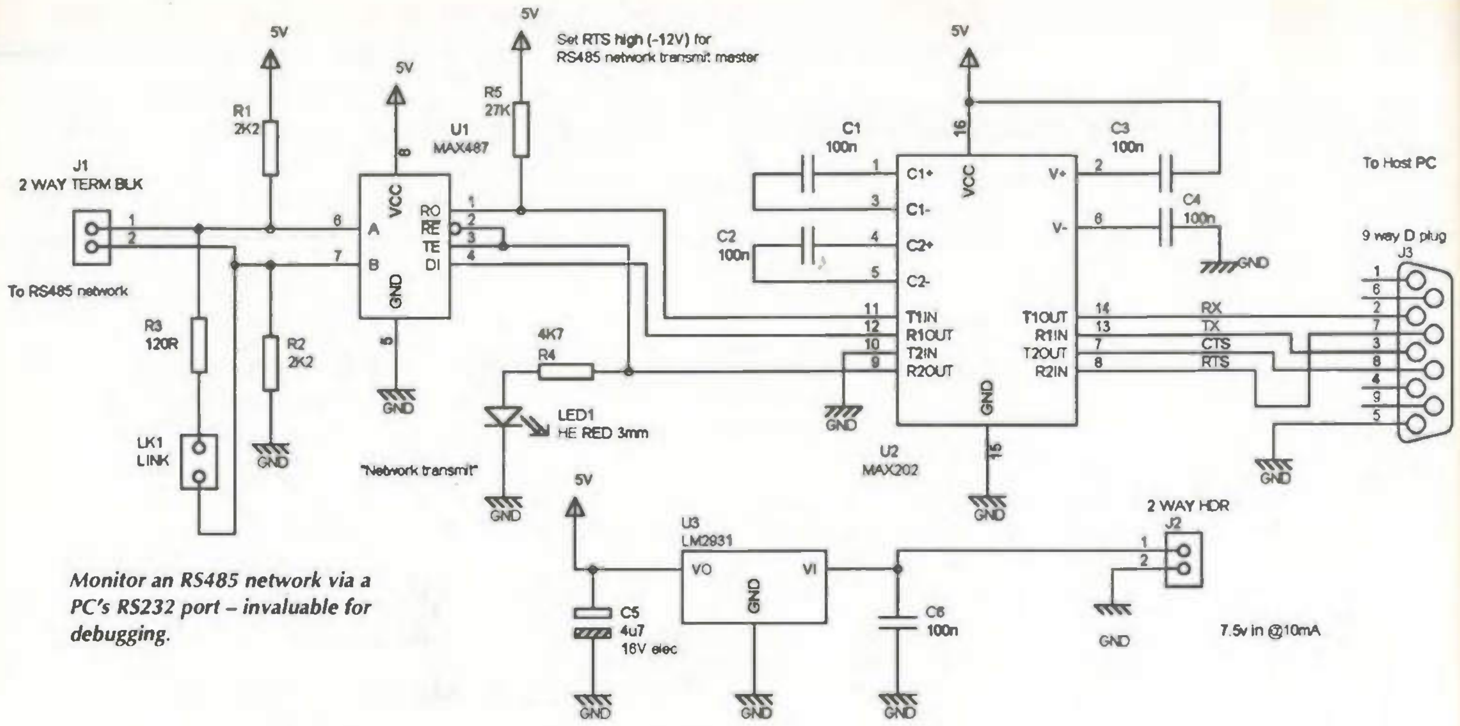


(F42)

Improvement to 'Wide range capacitance meter'.

Reference

1. Hickman, I, 'Wide range capacitance meter', *Electronics World*, Nov. 2000, p.905.



Monitor an RS485 network via a PC's RS232 port – invaluable for debugging.

RS485-to-RS232 transceiver

For industrial control systems, half duplex RS485 networks offer minimal distributed wiring. The remote stations share just a pair of copper wires as the communications medium.

For this to work, only one station can transmit data (single 'talker') at any moment in time to up to 31 receiving stations (multiple 'listeners'). Software protocols such as 'peer-to-peer' or 'master-slave' will ensure that there is no contention when one device needs to issue data to another addressed RS485 network device.

When developing or troubleshooting RS485 networks, the ability to connect up a PC to either monitor the data flow or emulate the action of a network master or slave unit is invaluable. RS485 interface cards are available for the PC but these require the use of an ISA slot and are generally impractical if the using a portable PC.

The circuit shown has been used extensively to test and debug several RS485 network designs both on the bench and in the field on various PC models. It simply plugs in to an RS232 COM port of the PC. This allows the PC to interface transparently to the RS485 network provided that the data direction is set according to whether the PC wants to send or receive data.

The RTS line is used to control data flow. If RTS is asserted (logic '0', line voltage +12V) which is the default state for most terminal emulation software then the interface will act as a listener. Generally, dedicated software would be required for the PC to implement the network data protocol. This code would then need to clear RTS (logic '1', line voltage -12V) when the PC is to act as the network talker.

The circuit has worked comfortably at 38.4kbaud over a short network connection but does need an external supply in the form of a bench PSU or a 9V battery. Current consumption will typically be about 9mA giving up to 15 hours use from an alkaline PP3.

No reverse polarity protection has been provided. An LM2941 low dropout regulator is used to extend battery powered operation.

The LED is useful as a visual indicator for the interface being set into the transmit state. Link LK_1 is normally left open unless the PC is at the end of a long line. Resistors R_1 and R_2 ensure that the interface does not get spurious start bits when all devices are set as receivers.

A basic set of Windows serial communications routines, written in Delphi, are available as a starting point for experimentation on request.

Huw Jones
Llantrisant
Mid Glamorgan
F31

Buck/boost regulator

Requiring a 25mA 5V regulated rail from a supply of 3 to 20V occupying minimum board space, I used a blocking oscillator to provide headroom for a conventional series regulator. Efficiency was of secondary importance.

The blocking oscillator transformer uses a Siemens B64290A37XI ring core of 6.3mm outside diameter, wound with 36 gauge 0.2mm wire. For a 3V supply, Tr_1 boosts the voltage to the 5V regulator, the total current then being 100mA for a 25mA load. As the supply voltage increases, current decreases as Tr_2 maintains the regulator input at 7.5V. Above 8V, Tr_1 is disabled and supply current is approximately 30mA.

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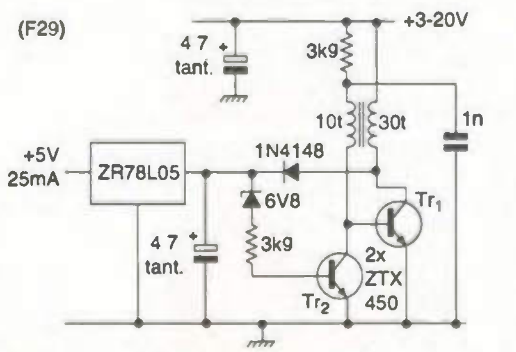
Salisbury
Wiltshire F29



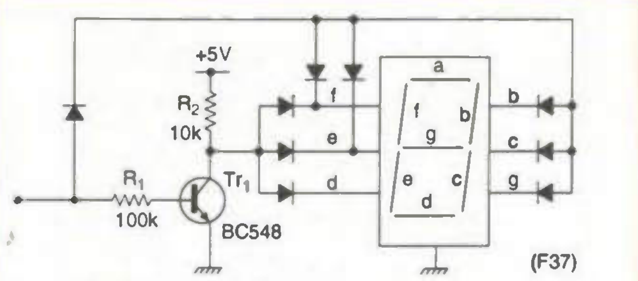
This switch-mode/linear regulator features a wide input range.

Cheap and cheerful logic tester

Here is a simple logic tester with which it is possible to check stuck high, stuck low or pulsing. The high or low state is indicated by the seven-segment display showing 'H' or 'L' respectively, while if the logic line under test is pulsing, both are indicated.

The transistor acts as an inverter while the diodes select the appropriate segments to light.

Raj K Gorkali
Via e-mail
F37



Simple circuit indicates logic states.

Free line indicator

When only one phone line is available for two phones: one in the office and the other in the workshop say, each time you want to make a call, someone else is often using the other phone... Even at home if you have one phone in the living room and the other one in your bedroom, when you try to make a call, you cannot see if the line is free.

This simple circuit can help you, lighting a LED to tell you if line is free or not. No battery is needed: the phone line powers the circuit and an accumulator saves energy for an 'in use' indication.

A phone line has different voltage between its terminals, depending on whether the line is free or not. The second diagram shows three possible states. As line terminal polarity is often unmarked, absolute value is represented. Rectifier bridge D_1 is used to ensure that voltage is positive for the circuit.

Phase 1: Line is free, voltage is continuous, at *circa* 50V DC. Series zener diode D_2 drops the voltage by 12V while R_1 and D_3 limit the voltage to 8V. Current goes through the NiCd accumulator, R_5 and D_6.

At this point, diode D_6 lights, and the voltage across D_6 turns on Tr_1 on. Hence Tr_2 is off and there is no current in D_5. The current is limited by R_5 to 3mA, which is enough to charge the accumulator. D_6 is a low current type. D_7 protects the accumulator against over voltage. An extra 2mA goes through D_3, giving a 5mA consumption from the line.

Phase 3: When answering the call, the voltage falls to a value close to 10V. Voice modulation rides on this continuous voltage. In fact, a phone line is considered as 'in use' by the exchange if a current close to 30mA is drawn through the 300Ω equivalent impedance of the phone. These current and impedance values are not critical.

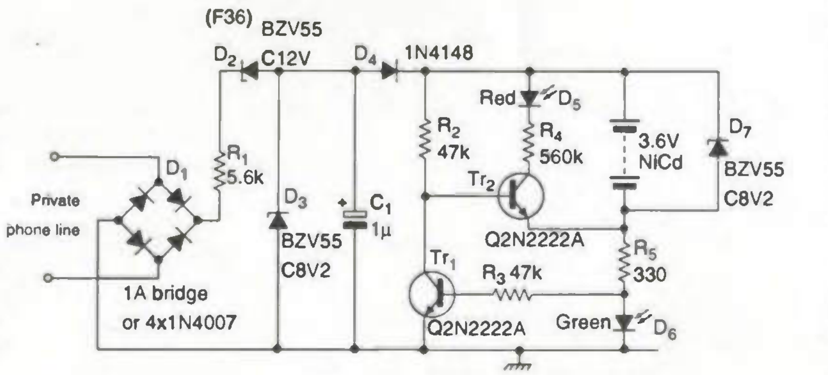
As the voltage is below 15V, D_2 is off and the voltage across D_3 is near zero. Now no more current goes through R_5 and D_6 is off, so Tr_1 is also off and Tr_2 conducts. Current is drawn from the accumulator through R_4 and D_5 is on. The current value is limited by R_4 to 3mA, which is enough for the low current LED D_5.

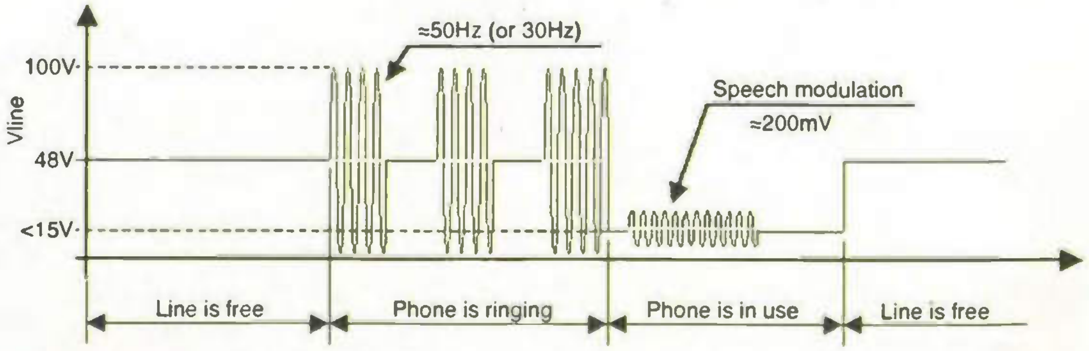
Phase 2: If the line rings, an extra AC voltage with an amplitude of 50V is added to the 50V DC. In this case, the value of C_1 *is* critical. If C_1 is 1µF, then both LEDs are on, the voltage being alternatively above and under 15V. If C_1 is 47µF, the voltage stays over 15V and D_6 is on.

The 300mAh, 3.6V accumulator is a phone type. Autonomy is more than any phone call duration. If the circuit is unplugged, the 'in use' LED D_5 will light until the accumulator is discharged.

This circuit can be used safely on an private phone line, but before connecting it to your operator's line, you need authorisation.

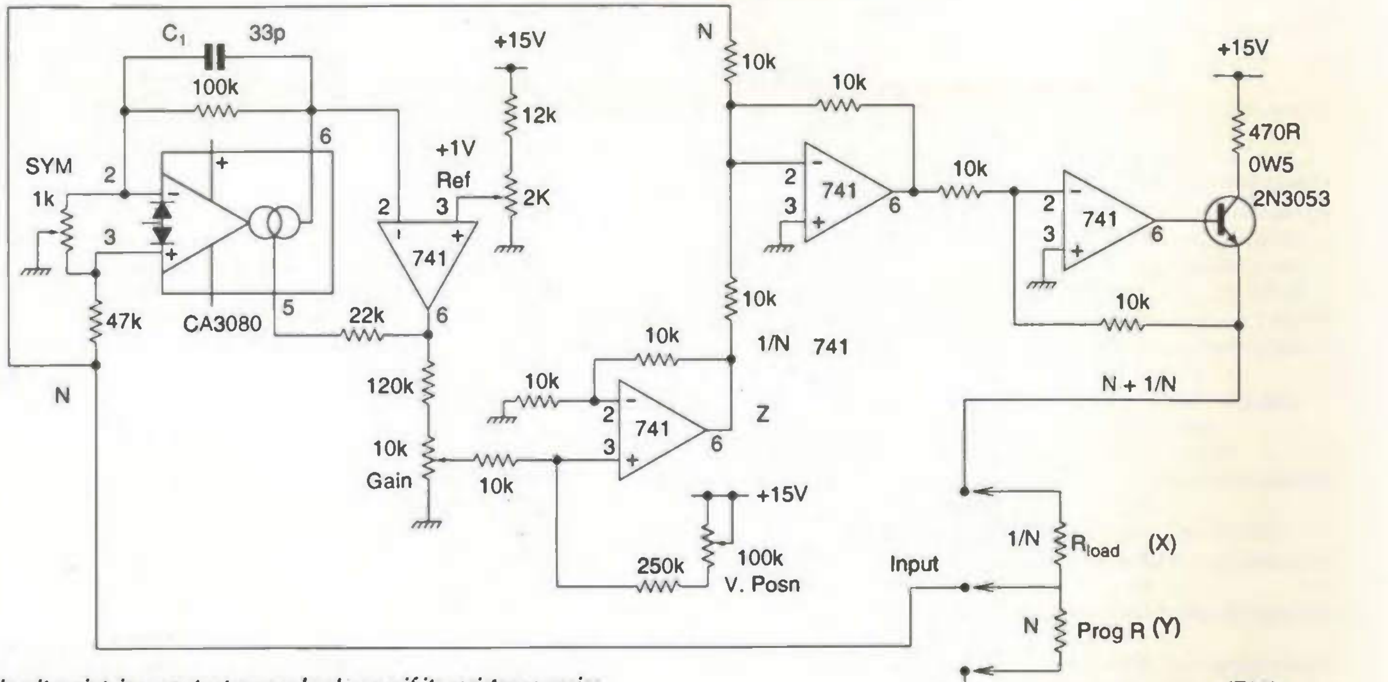
J M Terrade
Clermont-Ferrand
France





Busy/free line indicator, above, and its signal characteristics, below. In the circuit, the red LED indicates that the phone line is use while the green LED indicates that it's not.

£100 winner



Circuit maintains constant power load, even if its resistance varies.

Constant-wattage monitor

Constant current and constant voltage circuits are well known. This is a circuit that gives constant wattage, and when scaled up might be useful, for instance, for incandescent lamp stability.

The circuit depends on a reciprocal relationship between the load and sensor voltages The wattage delivered is dictated by a programming resistor.

The design presented revolves around one volt. Say the load is 100Ω and a constant wattage wanted is 0.01W. The regulator relationship is 1/N×N=1V, i.e. the reference voltage.

Input voltage is at Y and is represented by N. Resultant output is: $(1/N+N)$V. To find the programming resistor, you'll need to know load voltage, which is $1/N$ and W, which is 0.01 load or 100Ω. Then, $A{\times}A{\times}100{=}0.01\,W$, where A is 0.01 (current), and load voltage is 100×0.01, or 1V.

Programming resistor voltage N is 1/1V, i.e. 1V. Current is common so the programming resistor is 1V/0.01, or 100Ω. In this case both load and programming resistor are equal $(X+Y)$ and the voltage is 2V across them.

If the load were to change to 200Ω then for 0.01, $A{\times}A{\times}200$ is 0.01, so A is 0.00707. Also,

$200{\times}0.00707{=}1.414{=}1/N{=}X$V,
$100{\times}0.00707{=}0.707{=}N{=}Y$ volts

Output=X+Y=2.12V.

To calibrate the meter, adjust the SYM nulling potentiometer to mid-way and the reference voltage to +1V. With both the programming resistor and load absent, feed the input with a triangular voltage waveform of +2V top and +1V bottom from a function generator at 1kHz. Adjust both 'gain' and 'V position' potentiometers for an output of +1V top and +0.5V bottom from point Z. Note that with a reasonable layout, C_1 (33pF) capacitor should not be needed.

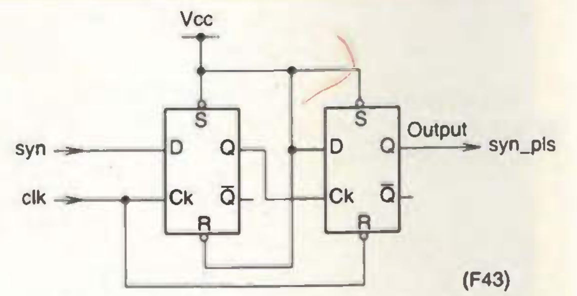
J R Hawkins
Chingford
London
F38

Single-pulse detector

This simple single pulse detector uses a 7474. It generates a single pulse of width same as that of 'high' time of the clock 'clk', on every leading edge of the input signal 'syn'. The first bistable device synchronises the syn with clk , while the second generates the pulse. Output Q_2 goes high on the leading edge of the output of Q_1, the first bistable device, and goes to low on the trailing edge of the clk.

I found this circuit useful when the leading edge of the 'syn' signal is jittery and has glitches. Any jitter or glitches with in the clk 'high' time is disregarded. The detector can also be used to generate fixed pulse width from any incoming edge.

S Vijayan Pillai
Kochi
Kerala
India
F43



This circuit produces a single clock-defined pulse following the leading signal edge.

£75 winner

Universal filter for audio frequencies

This universal filter can be used as a low-pass, band-pass, high-pass and notch filter at audio frequencies. Actually it can be adjusted to be any type within the second-order category. The transfer function is,

$$H(s)=\frac{k_{HP}s^2+k_{BP}\frac{\omega_0}{Q}s+k_{LP}\omega_0^2}{s^2+\frac{\omega_0}{Q}s+\omega_0^2}$$

Here, $s=j\omega$.

The filter can be characterised by five constants.

k_{HP}	high-pass voltage gain
k_{BP}	band-pass voltage gain
k_{LP}	low-pass voltage gain
ω_0	characteristic angle frequency
Q	Q value

The circuit shown has five potentiometers. Each one adjusts one respective constant independently, within at least one decade.

The first two operational amplifiers provide buffered non-inverted and inverted signals. There are three routes. Each route has a switch and a potentiometer. These form the polarity and magnitude of the low-pass, band-pass and high-pass gains.

A pseudo-logarithmic effect is provided by the resistor connected between the centre tab of the linear potentiometer and virtual ground. The main circuit includes two integrators and some gain and summing blocks. Component values are selected so that the characteristics, depending on the potentiometer rotation, follow the Table.

A signal generator and level meter are needed in adjustment procedure. The first thing is to adjust the frequency f_o. It is easiest to do in band-pass mode.

Block the low and high-pass signals by turning the respective potentiometers fully counterclockwise. Set the band-pass signal potentiometer to the middle position.

Now route the frequency f_o from the signal generator to the filter input and measure the output. Adjust the level to maximum with fo potentiometer.

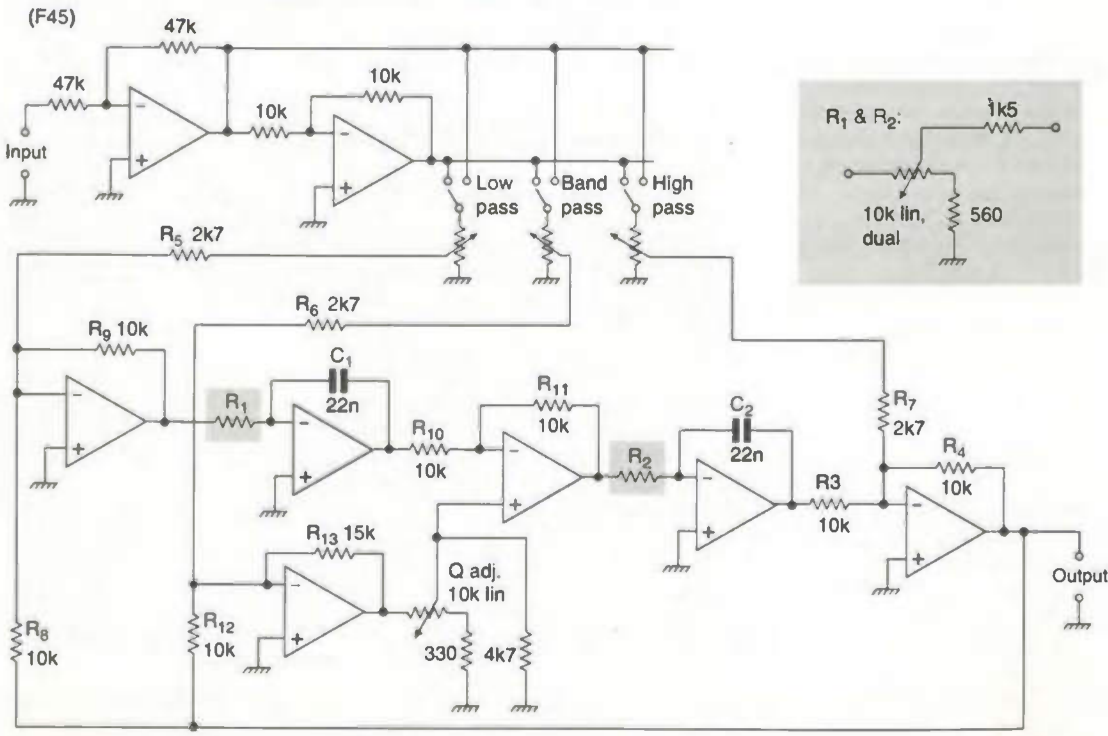
The next thing is to adjust the Q value potentiometer. Use band-pass mode and –3dB frequencies. The last thing is to adjust the level. If the target mode is notch, turn the band-pass potentiometer counterclockwise and use the same polarity and level for the low and high-pass signals. Restoring the band-pass component, in the appropriate polarity, then provides an all-pass characteristic.

Frequency f_o and Q value range can be easily changed without affecting the other characteristics. The frequency depends inversely on C_1 and C_2, which should have same values. The Q value depends inversely on R_{13}. ■

Ilkka Marttila
Espoo
Finland
F45

Table. Characteristics of the versatile audio filter.

	min	middle	max
f_o (kHz)	0.2	1	5
Q	0.3	1	11
gain (V/V)	0	1	4



Select characteristic, gain, polarity and Q with this circuit.

Letters to the editor

Letters to "Electronics World" Cumulus Business Media, Anne Boleyn House, 9-13 Ewell Road, Cheam Road, Surrey SM3 8BZ
e-mail j.lowe@cumulusmedia.co.uk using subject heading 'Letters'.

Phono preamp for the CD era

I must be one of many readers who spotted Douglas Self's oversight in his comments upon Norman Thagard's phono preamplifier design.

Given the figures that Douglas states for the resistance and inductance of the M75ED II cartridge, it is certainly true that 'the total cartridge impedance is much higher than the resistive part'. However, that impedance is still much lower than 47kΩ in the critical mid-band region where our ears are the most sensitive to noise. The noise current generated by the 47kΩ resistor in this region therefore is indeed shunted away by the cartridge.

This is easily proved by carrying out a series to parallel conversion of the cartridge components at, say, 2kHz. The new resistance, R_p, calculates out to about 58kΩ, but the inductive reactance, at 5.9kΩ, remains about the same as its series value. Therefore, while the noise due to the resistive component of the cartridge is of the same order as the 47kΩ resistor, adding 3dB to the available noise current, the resulting noise voltage is reduced substantially by the cartridge's inductive reactance, which it very largely controls.

At some much higher frequency where resonance occurs between the cartridge's inductance and its parallel capacitance and other strays, the cartridge can no longer control the noise. Because R_p becomes very much higher, the noise is dominated by that of the 47kΩ resistor. So, the National Semiconductor application note referred to by Dr Thagard is correct when it states that the 47kΩ resistor is a significant noise contributor only towards resonance when the source impedance is high.

It is easy to take as gospel what someone as distinguished as Douglas Self says in print and I really can't think what came over him this time. Being the present *EW* audio guru, maybe he was asked (cajoled?) to put his oar in but he was short of time and didn't really want to, so he did it without much thought.

And by the way, we seem to have forgotten an earlier audio guru, John

A lesson in new technology

The comment piece 'A Lesson in New Technology' by John Hindhaugh in the July 2001 issue, regarding the shortage of qualified electronics engineers, was interesting. I have followed the on-going debate concerning engineers' statuses, lack of pay, shortage of engineers, etc., for about ten years now, since I graduated with an honours degree in electronics and electrical engineering.

In my opinion there is one single reason why these problems exist. Low pay. If you opt for engineering as a career – not including contracting – in Britain, generally you sentence yourself to an existence on a pathetic salary. No amount of investment in fancy equipment for schools is going to change this fact.

I'll give you an example. While at the polytechnic, I shared a room with a friend who was doing a business studies HND. I was following a 3 year Beng course in engineering.

My friend left college after finishing his course a year before me, and got a job with Wandsworth council. I continued studying hard, for my final year, and graduated with a good honours degree.

After college, I went back to Plessey, where I had done my apprenticeship, which in the meantime had become GEC. Two years later, I met up with my friend again. He was doing well. He had been promoted and was now earning £40 000 a year.

I remembered how we both laughed long and hard, when I told him rather embarrassingly, that I was on a meagre £12 000 a year.

Ten years on, I am now working for a small electronics design consultancy, and things have improved. I enjoy my job and find it varied, challenging and satisfying, but I still don't earn enough to get a mortgage on a one bedroom flat!

Anyone for engineering?

***Nigel Dean**, Design Engineer*
Southampton

Valve ICs

Further to Jim McDermott's aside in the July issue on page 558 about early German valve 'Integrated circuits', I actually have one of these, a Loewe 3NF. It is in my collection of pictures, which can be seen on line at www.electricstuff.co.uk

One motivation for producing these masterpieces of glass blowing was a 'per valve' tax on radio receivers at the time – mid 1920's. These devices allowed a 'one valve' loudspeaker set to be made.

Mike Harrison
Via e-mail

Photos from Mike's web site, which features not only the Leowe valves but also numerous other valves and general science items.

Linsley-Hood, who long ago published in your pages a design for a phono preamp which used shunt feedback and a series 47kΩ resistor. I no longer have this design to hand, but I'm sure I recall that although it was admitted that that configuration was rather noisier than the parallel resistor one, it was still perfectly acceptable in that respect.

In those days, the emphasis seemed to be on the lowest distortion rather than the lowest noise; and in my view, if noise is not obtrusive at loud listening levels, we shouldn't bother about it and concentrate instead on correcting the more annoying circuit aberrations.

Mike Hall
Somerset

*Mike, what have I to gain from asking someone to criticise a design that I had chosen for publication? Ed.

I have some comments for Dr Thagard on his article, 'A Phono Preamplifier for the CD Era,' published in the April issue. First, let me thank you for sharing so many examples of audio work with us. This applies to not just the present article, but also those of the past on power amps, etc.

I have a relatively minor contextual but important point on a statement within the first preamp article. Under the heading 'Complementary symmetry', last paragraph, you say: "Thus, this design is mostly a JFET-based one" – a point that is obviously valid. You then continue, saying: "It is entirely JFET-based, if you accept the notion that the BJT active-load devices aren't really in the signal path."

Actually, I do question that comment, as follows. Taking the upper-half schematic portion of the circuit diagram as an example, clearly the signal carrying drain current of J_{1A} drives diode-connected Q_{1A}, which in turns drives Q_{1B}. Thus, this bipolar transistor signal current path also appears at the collector of Q_{1B}, along with the signal current from J_{1B}. In doing so, it flows through the p-n-p path.

A similar reasoning applies for the J_{2A}-Q_{2A}-Q_{2B} n-p-n-to-JFET signal path at the bottom. The statement with regard to the bipolar devices as not present in the signal path would, in my estimation, only be true if Q_{1A} and Q_{2A} were AC-bypassed to their respective rails. Or, if bipolar devices Q_{1B} and Q_{2B} were biased with a non signal-dependent voltage – i.e., like Q_5 and Q_6, respectively.

Having said that, I'd like to note that these technical points shouldn't be regarded as any overall criticism of the design's utility, but only as a further clarification of how it works. In my opinion at least, there is no inherent harm to an audio signal passing through bipolar transistor current mirrors, which can be operated quite linearly on a current-in, current-out basis.

Earth-leakage issues

In the July issue, Chris Miller asked if cabling and filter capacitance would cause an earth leakage circuit breaker to trip. The answer is yes they will.

Capacitors in EMC filters can also cause shock hazards if the earth goes open circuit. This is why special 'low-leakage' filters are used on medical equipment. Patients tend to have their skin penetrated allowing lower resistance paths to the heart.

The problem is worse at higher frequencies. Servo motor drives operating at a few kilohertz have so much stray capacitive leakage they can require an isolation transformer to prevent RCDs from tripping.

Robert Atkinson, G8RPI
Via e-mail

There is a possibility that the situation described in Mr Miller's letter in the July issue could indicate the presence of a serious electric shock hazard.

I think that 5mA leakage current from the aerial connection of a TV set is indicative of a fault in the set. The applicable safety standard limits the current to 0.5 mA and I would expect all sets in good condition to be inside this.

The TV should be taken out of service until the fault is rectified. If Chris measured the 5mA by connecting the ammeter between the TV aerial socket and the aerial down lead, this could indicate a low-impedance fault in the TV, with the leakage current being limited by the impedance of the return path from the antenna to earth. This could account for the nuisance tripping as the impedance back to earth is reduced in wet weather. I have seen one problem caused by the down lead rubbing on metal guttering.

Note that in this situation the aerial and everything connected to it could be at or close to mains voltage!

If Chris measured the 5mA by connecting the ammeter between the the TV antenna socket and earth – e.g. the earth contact of a socket-outlet – then, unless you included a series resistor to protect the ammeter, the current is clearly being limited by the impedance of the fault in the TV. In this case I suppose it is possible that the fault impedance could be varying – complete breakdown for example – so as to cause tripping of the RCD.

In reply to the rest of your letter, all installations and all earthed equipment will have steady state earth leakage currents. In general these may be due to

- degraded insulation in the installation or equipment, (true leakage);
- stray capacitance between phase conductors and earth;
- leakage current in some types of surge suppressors; and
- capacitors in RFI filters on the mains input of equipment.

The latter are probably the main source of leakage current when insulation and surge suppressors are in good order. (RFI filters are fitted to many types of domestic appliances and the like, as well as to electronic equipment.)

If a protective-earth conductor becomes open circuit, leakage currents may be available on the frames of all earthed equipment downstream of the discontinuity, clearly causing a shock hazard. To minimise the possibility of serious electric shock the safety standards for electrical equipment specify limits for leakage current. For stationary earthed equipment they are typically 3.5 mA and for equipment without a protective-earth connection, 0.25mA.

Note that connecting equipment via extension leads with multiple socket-outlets can accumulate their leakage currents beyond acceptable limits.

Tracing the source of earth-leakage current by measuring the differential current in the supply conductors is the correct method for installations since, as you discovered, earth leakage current does not always return directly via the protective-earth conductors. Of course a current transformer using a linear metallic-tape toroid would provide a fairly accurate means of measuring differential currents.

If there is a fair proportion of electronic load on the installation the leakage-current waveform may show some harmonic distortion.

Nuisance tripping of RCDs is often caused by leakage-current surges due to a few items being switched on simultaneously near the peak of the voltage waveform. In the absence of switching, perhaps mains drop-outs could produce similar surges. However, tripping may be due to true faults. In this case I would consider the TV as prime suspect.

If the tripping continues with the TV disconnected I recommend that you determine the cause. The manufacturer of the RCD is probably best placed to provide its tripping characteristics.

For more details on leakage current see:
IEC 335-1
IEC 60950
BS 7671 Section 607

Ted Smith
Via e-mail

I'm also confident that the circuit works as Dr Thagard has described, and readers should have few problems in replicating it. Thanks again for all of those interesting articles!

Walt Jung
Fallston MD
USA

I find it strange that Douglas Self continues with his negative letters concerning all things audio that don't happen to fit in to his idea of good practice and design. I am sure that the Dr Thagard design sounds admirable and it may well be that it betters anything that Mr Self has designed.

Without continued experimentation how can we hope to get higher fidelity? The knocking copy that Self produces can only be a disincentive to other designers to try to get their humble exploits published. Surely we need more designers – not fewer – and a diverse input of new ideas?

Unfortunately very few amateur electronics developers have access to or indeed want to use Pspice and again it is a shame that this is always brought into the equation. Pspice may be very clever but it is rare for there to be anything innovative to come from a piece of software. It takes the human brain and lateral thinking that we should all embrace as the way forward.

If Mr Self feels so strongly that he is right then perhaps he should tell the audio manufacturers directly; Pioneer, Akai, Creek, Chord, etc., rather than picking holes in others carefully considered designs. How about being a little more constructive and lighten the tone?

David Tutt
Via e-mail

One hesitates to comment on any audio discussion whatsoever, even a beauty like this one, for fear of being flamed in pseudo-religious hellfire.

First off, Self asserts that, "the modest resistive component... is completely dominated by its huge inductance." I hope that readers are not deceived into believing that the inductance generates thermal noise by this very unfortunate wording. However, it is indeed true that the inductance is so high that it will shunt off only part of the high-frequency hiss from the damping resistor.

There is another way to deal very effectively with that, which I will get to shortly.

Secondly, Thagard replies, at tedious length, "I ran some secret proprietary computer codes that exhibit many strange properties but are closed to all outside inspection, so it follows that all my conclusions must perforce be unassailable". Alas, this is a drearily tiresome line of argument nowadays.

Unfortunately, the television

Modelling the ground plane

In his letter 'Mysterious EMC' in the July 2001 edition, Philip Williams makes some complimentary comments about my article in the May issue. I thank him for those comments, and for raising a point that I can respond to.

Philip argues that the ground plane should be treated as an equipotential conducting surface. I disagree.

The method of images exploits the symmetrical nature of the field distribution round two parallel conductors, to derive a value for the loop inductance – or capacitance – of a conductor over a flat conducting surface.

The conducting surface is replaced by a non-conducting mirror, and the field distribution is simulated by the action of an image conductor below the surface.

By analogy with the circuit model of the two-conductor configuration, an equation can be derived for the loop inductance of real and image conductors. Since a voltage source is assumed to exist in series with the image conductor, the voltage applied to the wire/image loop is twice that applied to the wire/plane loop. Hence the loop inductance of the wire/plane loop is half that of the wire/image loop.

This does not mean that all the inductance of the wire/plane loop should be assigned to the wire, and none to the plane. The fact remains that a circuit model has been created, with the inductive (and capacitive) properties of the plane assigned to an image conductor.

The model is valid as long as the plane in the thought experiment is assumed to be non-conducting. For a single wire over a ground plane, half the voltage developed by a current transient in the loop should be assigned to the wire, and half to the plane.

The technique can be developed to create a circuit model of the coupling between two wires over a ground plane. This involves the use of three inductors, one for each wire, and one for the plane. It includes three capacitors and three resistors. See 'Grounding on a different plane', EW Aug. 1998.

This three-conductor model allows the interference coupling between the wires to be analysed, either in terms of frequency response, or in terms of transient response. If the ground plane were assumed to be a conductor of zero impedance, then the circuit model would predict that there is no coupling between two circuits sharing the ground plane as a return conductor. Such a circuit model cannot possibly be used to analyse interference.

To insist that the ground conductor be treated as an equipotential surface is to deny to oneself the use of a powerful analytical tool, and to give up hope of any clear understanding of the coupling mechanisms.

Ian Darney
Via e-mail

Philip Williams comments on the EMC article by Ian Darney were timely and needed. It shows the care needed when using equivalent circuits to represent an actual situation. Often they can be more of a hindrance than a help.

In this case, surely it is better to consider the effect of the ground plane from first principles?

Throughout the following, in the practical case, for "no flux" read "negligible flux". The voltage drop per unit length along the wire above the ground plane is due to its impedance, and the current flowing in it. The impedance is determined by the frequency of the signal, and the resistance and inductance of the wire.

The resistance will be higher than the DC resistance, due to skin effect, which ensures there is no flux within the wire, as there would be at DC. The inductance is due to the lines of magnetic flux surrounding the wire.

Both effects are relevant in considering the ground plane. No flux at RF can penetrate it, due to the skin effect, so if it really is infinite in extent, no flux can encircle any part of the ground plane: therefore it has zero inductance. But the ground plane will have resistance, and the return current will distribute itself between the send-end and receive-end ground plane connections.

The current density will be greatest along the shortest path, i.e. the straight line between the connections, falling off on either side, on roughly elliptical paths further and further away from this line.

Assuming that flux really cannot penetrate, and thus encircle any part of, the infinite ground plane, I see no reason why the ground plane current distribution should differ from the distribution that applies at d.c. Current will therefore leave or arrive at the connection points over a full 360°, the lines of current flow being everywhere orthogonal to the equipotential lines, as in the DC case.

Along any path, the current density is then proportional to the 'closeness' – inversely proportional to the spacing – of the equipotential lines. All of this, of course, depends on the applicability in the practical case, of "no flux" meaning "negligible flux".

Ian Hickman

EMC and the DIY PC

I would point out to Mr Williams, who had a letter published in the July 2000 issue, that it was not me that said that a home-built PC had to be 'CE certified' (whatever that means – it is a concept unknown to the Directives!), or indeed the other remark attributed to me. The author was Mr R Atkinson, G8RPI.
John Woodgate
Via e-mail

generations find it persuasive, what with the nearly insurmountable difficulties they have distinguishing their screens from reality. And yet it's no answer at all, now is it?

We might try adding – fancy this – a bit of actual thought, a simple plausibility analysis. Pick a moderate frequency at which noise – hiss – will certainly be audible. 2.5kHz will do. Self's 610Ω coil will generate about $3.2nV/\sqrt{Hz}$ at this and any frequency.

The 470mH coil inductance will have an impedance of about 7400Ω. This will pass the 3.2nV through to a 47kΩ damping resistor and high-impedance JFET gate virtually unattenuated.

Modern ears raised on deafening music will lack the capacity to notice the slight roll-off at the very top of the audio band. Beyond that, the damping resistor will generate about $28nV/\sqrt{Hz}$. The coil will shunt this down to around $4.4nV/\sqrt{Hz}$. This component will double with each of the next couple of octaves as the inductive impedance rises.

Overall, in audible terms, this noise from the damping resistor will well overmatch the the coil noise, confirming Self's assertion. This may seem to contradict AN-346 – available at www.national.com – but AN-346 is immune to contradiction since it does not state its assumptions concerning the cartridge coil, which may include a lower inductance. If it contradicts one or another of the Spice runs, then I suggest correcting the Spice model, which may not be easy.

Note also, that AN-346 does give excellent reasons to use a two-amplifier topology. Self's note demonstrates that he did not trouble to do them justice, and Thagard's reply demonstrates that he did not trouble to understand them. Neither action was helpful to readers, so, may a plague of wet noodles be visited, share and share alike, upon both.

Above 2.5kHz, a reasonable JFET preamp, at 2 or $3 nV/\sqrt{Hz}$, will add only negligibly to the total noise. At low frequencies its contribution will eventually become slightly more significant, but any effect will be well masked by the higher frequency hiss. Indeed, you could nearly as well use a good JFET op-amp at $4 nV/\sqrt{Hz}$, if your religious bias permits it.

By the way, Bob Pease – also available at www.national.com – gives various good tricks to improve distortion performance of op-amps. For example, careful loading to V– (or V+) will eliminate crossover distortion.

Bob also maps out some of the garden paths down which even very bright engineers have been led by the strange properties of the secret proprietary codes of Spice. Anyone thinking about doing designs of this sort should read his notes. All of them.

Finally, to reduce the noise further, as promised, while retaining the damping and frequency response intact, one must resort to 'electronic cooling' of the 47kΩ resistive impedance.

Rather than connect a 47kΩ resistor as a shunt from the JFET gate to ground, connect, say, a 470kΩ resistor from the gate to a circuit node having a gain of –9. And, of course, see to it that there is enough bandwidth and proper roll-off at that node, so that the arrangement will not oscillate, and will have small phase shift up to whatever frequency one's religious beliefs might require.

This may require a very surprising amount of bandwidth if an op-amp is used, since op-amps use bandwidth inefficiently. All the same, as long as the amplifier input is quiet enough, it will reduce the noise.

The reduction will even show up in a properly constructed Spice model, provided that you can coerce it into attaining correct DC levels and converging – matters that you should check every time without exception when making Spice runs for publication.

Note, by the way, you can't 'cool' the coil's own resistance this way – so no mail about that please.
Paul Schick
Madison
Wisconsin
USA

High gain and Class-A amplifiers

Here are some comments about the high-gain and Class-A amplifiers published in the June 2001 issue.

Regarding 'High-gain amplifier uses medium-power MOSFET' on page 466, the circuit, as it is drawn, does not seem to have an input impedence of 10MΩ; rather, the input impedence is 10MΩ/(1+gain), where 'gain' is the voltage gain at the emitter of the BC109. This gain can be up to 30, so the input impedance will be about 300kΩ for the minimum overall gain setting.

It seems better to connect the right side of R_2 to the plus side of C_3, thus avoiding the AC negative feedback on R_2, while retaining the DC feedback for the self-bias.

About the 'Low dissipation Class A amplifier' on page 467, assuming that the "Op-amp A2", not marked in the drawing, is the lower OP2134, the phrase, "As it does so the non-inverting input of A2 goes positive..." should read "As it does so the non-inverting input of A2 goes negative...".

The lower TIP142 should be labelled Tr_2, not Tr_1. Apart from that I wonder if the circuit should be considered truly class A. Transistor Tr_1 is always in conduction and amplifies both positive and negative voltage semiperiods, as in Class A circuits. On the other hand, Tr_1 supplies to the load only the positive half waves of the current, and Tr_2 supplies the negative ones, as in Class B circuits.

Is the 0.01% distortion figure measured or calculated?
Ezio Rizzo
Via e-mail

What kind of lightning?

In an american magazine from 1994, I read of 'Lightning sensors' intended for detecting an impending storm from its electromagnetic activity. The sensor would give an alert, or directly connect the antenna of an amateur radio to the ground.

Some of them where said to detect 'specific signal profiles and lightning signatures' to avoid false alarm due to man-made statics.

Does anyone knows how do they discriminate between lightning and man-made static electricity?
Ezio Rizzo
Via e-mail

How accurate is vinyl?

Norman Thagard's phono preamp article reminded me of something that has been a puzzle for a number of years.

While it seems quite a simple task to build – as opposed to design – a phono preamp with the RIAA curve accurate to 0.1dB or better, I have often wondered how accurately records were cut in the first place. I say were as I haven't bought vinyl for 15 years.

The EMI Technical Test Record TCS101, for example, according to data on the sleeve is anything from +0.5dB to +1.5dB (ref. 1kHz) at frequencies of 6kHz and higher and has up to 1dB difference between the left and right channels at 4kHz and above. It seems reasonable to assume that some considerable care would have been taken in producing this test disc.

So just how good was the run-of-the-mill record production? If by chance there ever was a perfect test disc that when played gave an output as flat as the proverbial pancake, was it still perfectly flat after fitting a new stylus into the pickup cartridge?
Barry Taylor
Via e-mail

Ice alert

Cyril sure dropped a clanger in 'Ice alert' on pages 215 and 217 of the March issue. Differential amplifiers respond to the difference of signals at the inputs, not the ratio. A ratiometric amplifier is a divider.
Phil Denniss
Sydney
Australia

Cyril replies:
I'm not certain who the 'clanger' that Phil refers to belongs.

The LM3900 is of course a 'Norton' input amplifier which responds to current inputs only and not voltages.

Re-looking at my schematic, I see I omitted to include the 'arrow' indicating current flow between its input pins, often used to indicate a 'Norton' input. For this I apologise. National Semiconductor takes care not to refer to the LM3900 in their data as a differential amplifier for this reason.

The circuitry around 'A1' in my schematic provides no input bias voltages, so a conventional difference amplifier would not work in this circuit.

We commonly refer to a signal having two equal time periods or voltages as having a 50:50 ratio – not pedantically correct perhaps but well understood by all.

In my text, for brevity, I was describing the action of the complete circuit not just 'A1' on its own. This LM3900 circuit has as its output two LEDs. These illuminate for time periods that relate to the input currents from the thermistor and resistor chain.

At the median temperature of 2.5° the LEDs are each on for one half the oscillator period, or in the parlance I was using have a 50:50 mark-space ratio. In this respect, this circuit's action differs substantially from most other ice warning indicators.

Finally it would be more accurate to describe the 'A1' circuit as a current comparator with hysteresis. That is exactly how it acts, combined with the other stages. Above 5° the 'Green' LED illuminates continuously, the 'Red' LED being extinguished. Below 0° the circuit toggles the red LED continuously on, the green LED is then fully off.

Finally it was Phil Dennis who introduced the word 'ratiometric' – not me.

Low-consumption Class-A

I read with interest the circuit idea by Jeff Macaulay in the June issue entitled 'Low-dissipation Class-A amplifier'.

The topology described in that idea is the subject of patent application number 9804063.7 filed in February 1998. A 50 watt prototype has been completed with a quiescent consumption of only 16 watts. It is presently being auditioned.

Initial reports suggest that it is indeed possible to obtain Class-A performance without the daunting penalty of enormous quiescent power.
Clive Read
Brighton

Stranger than fiction indeed

Although much is made of the better drag coefficient (Cd) of a VW bus compared with a Jaguar E type, the figure one is really interested in is Drag (Force).

Drag is proportional to the product of Cd and Frontal area, so the VW bus may well produce more drag than the E type.

I've never understood why car manufacturers make so much of a fuss about drag coefficient in their publicity and never mention the frontal area.
Robert Royall
Via e-mail

Low-pass filter distortion

I have been following the correspondence about Graham Maynard's claim for distortion introduced by a low-pass filter. I have done so with interest, and a mixture of amusement and despair. He says that the introduction of a low-pass filter at 72kHz into a 'filterless' audio stream is audible, and gives a waveform plot that appears to demonstrate this distortion.

Any real audio signal will have already received considerable mechanical or electrical low pass filtering at lower turnover frequencies — and probably higher orders — than this from microphones and transmission or storage methods. Mr Maynard's filterless audio source does not exist in reality, so I guess his 'distortion' is inaudible with real signals.

His waveform plots simply demonstrate that a sine wave switched on at t=0 is a different signal from a continuous sine wave, and so gives rise to a different output after a low pass filter. His input signal really has two components: a sine wave and a switching pulse. The sine-wave part comes out simply delayed, or phase shifted — the two descriptions are equivalent ways of saying the same thing. The switching pulse comes out lengthened and attentuated. What he sees as initial distortion on the sine wave is actually the exponentially decaying output from the pulse. See any textbook on signal processing for the maths.

A more intuitive method is to regard the input as the sum of two waveforms: A is a 0.5V continuous sine wave, B is a 0.5V sine wave at the same frequency but inverted in sign before t=0. So before t=0 they add to give zero, after t=0 they add to give a 1V sine wave, at t=0 they are both zero anyway. The low-pass filter is a linear circuit so we can use superposition.

The A wave comes out simply delayed/phase shifted. The B wave looks the same — albeit inverted — up to t=0. But the input B wave has a sharp cusp at t=0 where the two opposite polarity sines meet. The low-pass filter smooths off this cusp. You can sketch what this might look like. The waveform then settles down again to simple delay/phase shift.

Add A and B together and you get Mr Maynard's 'distortion' — it is simply the smoothed-out cusp from B.

I am surprised that so much correspondence has resulted from such a simple phenomenon. ■
Dave Kimber (G8HQP)
Davant Technology Limited
Via e-mail

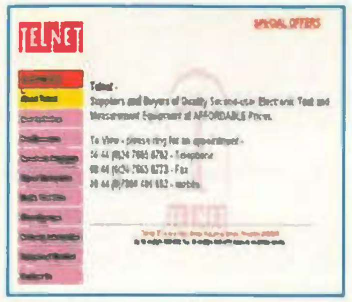
Oops

Despite proof reading, one or two misprints have still slipped into my article 'Pseudo-random bits', in the July 2001 issue. In the paragraph 'Polynomials', in the first expression, a mysterious 'K' has appeared.* It was three dots in the proofs, indicating other terms – not printed – in the general polynomial.

Lower down in the same column, 'octal to decimal conversion' should read "octal to binary conversion" (my error, not a misprint).

At the end of the paragraph "Time-reversed sequences", the 'r' and 'n–r' should be superscript, indicating powers of x, and in the last two lines of Table 2, 'n–4' and 'n–3' likewise should be superscript, indicating powers of 2.
Ian Hickman
Via e-mail

*The mystery 'K' was an ellipsis at every stage while the article was being produced up to the point where it left our offices to go to the people that make the film for us. They couldn't have altered the equation, which is an individual EPS file, because they don't have the software to do it with. I've even gone so far as to check whether the ASCII character for the ellipsis could be a K in another font. We're looking into the matter at the moment but if anyone has any ideas, we will be pleased to hear from you. Ed.



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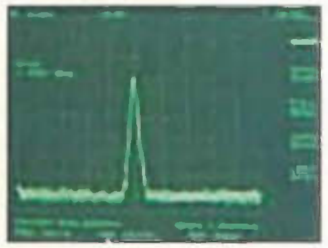

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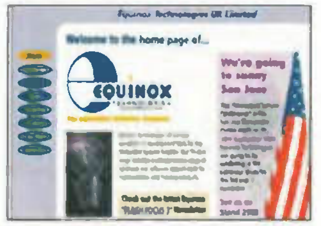

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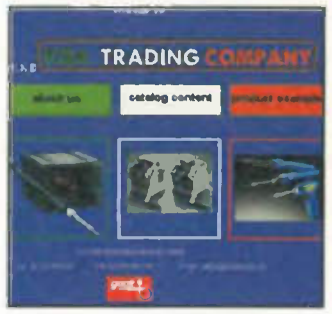

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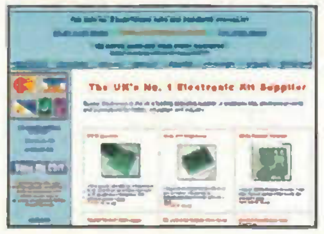

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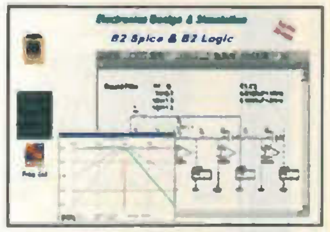

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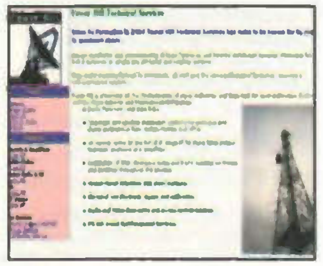

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Following on from the newsgroup discussion last month there is a UK Email group for TV technicians where you can send an Email to everyone in the group. There's just over 30 people in the group at present. For more details and how to register look at the egroup home page. Just a general comment though - you do have to be careful who you give your Email address to so that you can avoid "spamming" - that is getting lots of unwanted Email about dubious Russian site (amongst others).

REPAIRWORLD
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Repairworld is a sophisticated US based fault repair database which is updated bi-weekly. It operates on a subscription basis and describes itself as an "affordable solution for all technicians". You can see some samples of the material for free, monitors, VCR, DVD and Camcorders being of particular relevance to UK users. The site also provides a "chat room".

Beginners' corner

Transforming power

Having covered the basic theory in a previous article, Ian Hickman now reveals how to design a power transformer.

A previous article[1] of mine laid the foundation for an understanding of transformers, of which there are several very different types. All can be approximated by 'the perfect transformer'. This is a useful concept that would be even more useful if such a device existed.

A perfect transformer is illustrated in Fig. 1. This transformer exhibits an exact relationship between its primary-to-secondary turns ratio, and the ratios of primary to secondary voltages and currents, as shown.

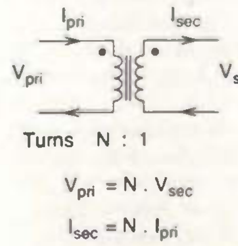

Fig. 1. A perfect transformer with a turns ratio of N:1. N may be greater or less than 1.

If the primary to secondary turns ratio N is greater than 1, it is a 'step-down' transformer. An example is a mains voltage with 220V AC in and 12V AC out. In a 'step-up' transformer, N is less than 1 and the secondary voltage is higher than the primary.

The AC voltage applied to the primary causes a current to flow. This alternating current creates an alternating flux on the core and that in turn generates a back emf in the primary winding. In a perfect transformer, this back emf exactly equals the applied voltage, and very nearly also in a practical one, at least when off-load.

Warts and all

Figure 2 shows a representation of a practical transformer, with winding resistances, leakage reactances and the primary magnetising reactance. The dots indicate the start of each winding, both assumed wound in the same direction. Thus the voltages at the dot ends will be in phase.

What Fig. 2 doesn't show is the self- and inter-winding capacitances, which become very important in a wideband signal transformer. So assume for the moment that it is a mains transformer, where the capacitances can usually be safely ignored.

If you assume for the moment that the winding resistances and leakage inductances in Fig. 2 are zero, that leaves just the magnetising inductance L_m. This will have a reactance X_m at the frequency of the mains supply, of $2\pi fL_m$, where f is the frequency of the mains supply, usually 50 or 60Hz. So the transformer will draw a 'magnetising current' I_m of V_{pri}/X_m. Because the winding resistances and leakage

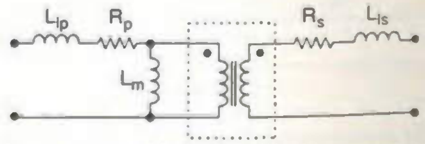

Fig. 2. A practical transformer, warts and all.

reactances have been assumed zero, in this case $V_{pri}=V_{ac}$.

In the more practical case, the windings will have a small but finite resistance, R_p and R_s. Furthermore, while virtually all of the magnetic flux due to the magnetising current will link with the secondary winding, a tiny amount will not. This is represented in Fig. 2 as the primary leakage reactance L_{lp}.

Likewise, if you were to use the transformer 'back to front', applying a suitable voltage to the secondary, a small proportion of the resultant flux would not link the primary, so there is also a secondary leakage reactance.

When the transformer is used the right way round and a current drawn from the secondary, this current must flow through both the secondary winding resistance R_s, and the secondary leakage reactance L_{ls}.

Filling in the detail

When a load current I_{sec} is drawn from a secondary with n_s turns, that current exerts a magnetising force $n_s \times I_{sec}$ on the core. But the flux on the core must determined solely by the magnetising inductance and the supply voltage. So a component of primary current flows, which exactly balances the effect of the secondary current.

The resultant situation is shown in **Fig. 3**. Assume that the current is the rated load current, and that the transformer has been so designed that the magnetising current on full load is 10% of this rated load current.

The relationships between the various currents and voltages can best be shown on vector diagrams. Getting from the information in Fig. 3 to the vector diagram is best done in stages.

The way not to start, is to draw in the applied voltage. Better to start with the current diagram of Fig. 4, beginning with the secondary load current I_{sec}. The resultant component of primary current $I_{pri}=I_{sec}/N$ can then be drawn in. Here, N is assumed to be 2. Note that it flows in the opposite direction: thus the primary and secondary ampere-turns due to the load current cancel out. The only net magnetising force on the core is that due to the magnetising current.

Total primary current I_{pt} is the sum of I_{pri} and the magnetising current I_m. The magnitude of I_m has been assumed equal to 10% of the load current, but being a purely reactive current, it will be lagging I_{pri} by 90° as shown.

Having sorted out the currents, you are now in a position to produce a vector diagram showing the various voltages. The voltage across the primary of the perfect transformer is the same as that across the magnetising inductance, namely $X_m \times I_m$. To this can be added the voltage drops due to I_{pt} flowing through R_p, and L_{lp}, and due to I_{sec} flowing in R_s, and L_{ls}.

The complete voltage vector diagram, **Fig. 5**, assumes the secondary winding resistance and leakage inductance have been referred across to the primary, as explained below.

Designing a mains transformer

Often the mains transformer requirement for any given design can be met by a stock type from a catalogue. Where a suitable model is available, this is always the best option.

Where a suitable type is not available, a special-to-type design will be necessary. Whether you have the mains transformer designed for you by one of the firms advertising in this journal, or you design it yourself, the first requirement is an exact specification of what you want it to do. For example, if you want a small mains transformer to give 1A at 12V into a resistive load from a 240V AC supply, you do not want a turns ratio N of 20:1. The design must take into account the voltage drops in winding resistances and leakage reactances mentioned above.

The first thing to decide is how much magnetising current you are willing to allow. For a transformer as small 12V 1A, one would allow a figure of around 10% of the full load primary current. In a large mains distribution transformer at a local substation a much smaller percentage would be normal, for two reasons. Firstly, in larger transformers a more efficient design is possible, and secondly, a distribution transformer spends most of its time on less than full load – overnight often virtually off load.

The allowable magnetising current, together with the mains voltage on top limit, determines the required primary inductance. On a given toroidal core or stack of E and I laminations, this can be achieved with a certain minimum number of turns n_{pri}. The required primary inductance $L_m=n_{pri}{}^2/S$, where S is the reluctance of the core.

From the data on the laminations a suitable size and depth of stack to provide the required S can be found, or a suitable toroidal core selected. For these ready made cores, an 'inductance-per-turn' figure A_L may be provided, in which case, quite simply, $L_m=N^2.A_L$.

Just providing the minimum L_m is not enough. A check is needed to verify that with the working primary voltage on top tolerance limit, the maximum flux density Φ_{max} in the core does not exceed the maker's recommended maximum figure.

The maximum flux density is determined from the equation $V_{pri}=4.44n_{pri}f\Phi_{max}$, where f is the frequency of the mains supply. If the resultant Φ_{max} turns out to be too large, two courses are open. Either increase n_{pri} or, better, choose a core or stack of laminations with a larger cross sectional area A.

You are now reaching the nitty gritty stage of design, and the next thing to do is to choose a suitable gauge of wire for the primary, knowing the primary full load current. For small transformers up to about 100VA rating, a current density in the wire of about 2000A/in^2, or 2000A/(25mm^2), is reasonably conservative. From this, the area of the winding window occupied by the primary winding is calculated. This should be not more than half the available area, and less if there is more than one secondary to accommodate; otherwise, a larger core size will be needed.

Often, a moulded plastic bobbin can be used, with a central partition. This provides equal areas for primary and secondary, with the necessary safety creepage distance separating the windings ready built-in. At the same current density as the primary, the secondary should fit in the remaining available window space.

Remember that at full load, the effective primary voltage will be slightly less than the supply voltage, and that there will be a further small volt-drop due to the secondary winding resistance. So the actual primary/secondary turns ratio N will be less than the ideal theoretical value.

The difference between the off- and on-load secondary voltage is called the regulation, and a value of 10% is often accepted – even worse in very small transformers of only a few VA rating. Being in quadrature, the voltage-drops due to L_{lp} and L_{ls} are usually negligible on resistive load, but not so on a highly reactive load, whether capacitive or inductive.

For convenience, the secondary winding resistance can be referred, through N^2, to the primary side, to give a single winding resistance figure. Thus if R_s=600mΩ and N=10, then the total effective primary winding resistance, call it R_T, can be considered as 60Ω plus the actual primary resistance, with the secondary winding resistance now considered as zero.

Likewise, the secondary leakage inductance can be rolled up with the primary, in the same way, to give L_{LT}, with reactance X_{LT}. The actual primary-referred total leakage inductance is easily measured by connecting an inductance bridge to the primary, with the secondary short circuited.

For safety, the primary circuit should be fused at a suitable current rating. Use a slow-blowing fuse to avoid nuisance blowing, if the fuse rating is not greatly in excess of the rated full-load primary current. If there are two or more secondaries, with one rated at a tenth or less power than the main secondary, the low power secondary should also be fused. A short circuit on such a secondary would cause only as much primary current as the rated full load on the main secondary. Consequently, the primary fuse would not blow, and a damaged transformer – or even a fire – could result.

Note that when supplying a rectifier circuit, the effective secondary current is the rms value of the rectifier input current. For a full-wave rectifier with capacitor input filter, driven from a centre tapped secondary, where current is drawn from each half of the secondary in turn, the effective current in each half winding is about 1.1 times the DC load current provided. For a bridge rectifier, where the full secondary provides current on both half cycles, this rises to 2.2 times.

Due to the large short pulses of current, drawn at the peak of the half cycle, the voltage drop in the windings will be greater than when supplying a resistive load. Allowance must be made for this when designing the transformer.

Transformers may also be called on to deliver current to a highly reactive load. In this case, the current is sinusoidal, unlike that for a rectifier transformer. But it still has to flow through the winding resistances. Consequently the dissipation in the transformer, at rated secondary current, will be as great as for a resistive load – possibly greater, even though the power supplied to the load may be small. For these reasons, transformers are always quoted with a VA rating, rather than a wattage rating.

I hope to look at other types of transformer in a later article. ■

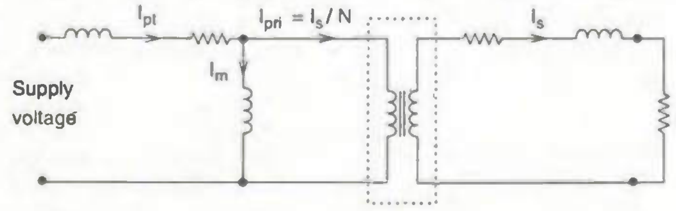

Fig. 3. A practical transformer supplying a resistive load connected to the secondary.

Fig. 4. Showing the currents flowing, assuming N=2, $I_m=I_{pri}/10$. Although the primary and secondary currents are different, their ampere.turns product on the core cancel exactly.

Fig. 5. Showing the corresponding voltages (secondary resistance and leakage inductance referred to the primary). X_{LT} is the total leakage reactance at the supply frequency, referred to the primary.

Reference

1. Hickman, I., 'Understanding transformers I', Electronics World, June 2001 p. 458.

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