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806 FIELDS LINES AND TREMORS
Dominic Di Mario presents three very different VLF detectors for examining your environment. One senses electric fields, one examines magnetic lines and one records unfelt vibrations in the earth.

812 POST-OFFICE WAR RECORDS
Andy Emmerson has been investigating some of the Post Office's strange activities during the war - one of which involved producing vinyl disks.

815 SPEAKERS' CORNER
John Watkinson explains why good loudspeakers are so difficult to produce.

817 KILLING NOISE
Ian Hickman looks at one of the most important nuisances facing designers of oscillators for rf comms - phase noise.

825 DESIGN WIDEBAND ANTENNAS
An impedance-loaded antenna capable of continuous coverage from 6 to 150MHz without tuning or matching illustrates Richard Formato's design procedure.

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Have you splashed out on the world’s best hi-fi loudspeaker cables? If so, the last thing you need is Doug Self’s revealing article.

837 A PROBLEM PHASE
Use textbook equations for a Wien oscillator and you could end up with frequency errors of up to 70%. Bryan Hart explains why.

840 HANDS-ON INTERNET
Find better mosfet models, new amplifier designs, improved simulation software and faster search methods on Cyril's pages.

843 WORKING WITH MICROWAVES
Boris Sedacca shows that circuits for working at microwaves need few components and are quite easy to design - provided you can afford the software.

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Efficient psus save energy, but in many applications, power down mode can save even more. Phil Darrington reports.

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CAN is a versatile serial bus designed for automotive applications. But it is finding new uses, as Svetlana Josifovska explains.

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15% reader discount on temperature-controlled soldering stations – see page 817.

Read about a nanoguitar, an electronic guide dog, and this autonomous robot in Research Notes starting on page 802.

Doug Self has been analysing cables for connecting loudspeakers. See his conclusions on page 801.

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Research or eco folly?

The High-Frequency Active Auroral Research Project - HAARP - has been a source of increasing concern among scientists and the general public alike, especially given the recent advances in ionospheric research. As the US military continues to expand its HAARP system, concerns about the implications of such high-intensity electromagnetic emissions rise.

"HAARP is much more than a weather modification experiment, it is a potential weapon of mass destruction," says Dr Bertell, a researcher in the field of quantum biodynamics. "The ionosphere is a complex system that is delicate and easily disturbed. No one really knows how ionospheric phenomena will affect the balance, or what the Earth will do in response to try to restore balance." These words are echoed by many scientists around the world, raising questions about the safety of such activities.

The HAARP project may be the test run for a ground-based 'Star Wars' defence system. With powerful pulsed radio frequency beams, radiation, in some cases, can change the surface of the planet in a way that affects everything from basic communication to physical stability. Radiation, in some cases, can change the surface of the planet in a way that affects everything from basic communication to physical stability.

Individuals participating in the European Parliament are among the growing number of people worldwide who have been startled to hear about HAARP. Some are concerned that this system and others like it could affect the health of humans and other biological systems. Others are concerned about the potential for interference with natural systems, such as weather patterns or navigation systems on the earth and high above it.

The concern that this system and others like it could affect the geophysical stability of the planet has also been expressed by Brooks Agnew. Brooks is a radio chemist who discovered that some very specific frequencies when broadcast into the Earth could trigger earthquakes even at relatively low power levels.

This same sentiment was expressed by William Cohen, the United States Secretary of Defense on 29 April 1997. Commenting on the possibility of terrorists states he said, "...engaging even in an eco-type of terrorism whereby they can alter the climate, set off earthquakes, volcanoes remotely through the use of electromagnetic waves causing destruction "when interacting with protective layers of the earth and its gravitational field."

The concern that this system and others like it could affect the geophysical stability of the planet has also been expressed by Brooks Agnew. Brooks is a radio chemist who discovered that some very specific frequencies when broadcast into the Earth could trigger earthquakes even at relatively low power levels.

The answer is that the US military wants to communicate with its submerged submarines by penetrating the ocean with extremely low frequency radiations. It also wants to penetrate the land with ELF waves in order to search for hidden tunnels or other sites of military interest - a process known as earth-penetrating-tomography. This application is funded under the counter proliferation and non-proliferation of nuclear weapons in the United States Defense budget.

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Assault on batteries

Battery maker Duracell wants battery specifications tightened to avoid the increasing risk of appliances such as smoke alarms failing.

The specifications, defined by the International Electrotechnical Commission (IEC), cover the shape, size and discharge currents of batteries in various applications.

According to Philip Smith, sales manager for Duracell's industrial products division, battery performance is being stretched beyond the IEC's tests and standards. Discharge tests run from 1 to 500mA.

"That doesn't take account of datalogging equipment or cellular phones," said Smith. "These drop off either end of the scale, he claimed.

Because the specifications are lax, a battery manufacturer can change the design to a point where it causes problems. For example, at extremely low currents, perhaps tens of microamps in a smoke alarm or datalogger, the battery might even stop working.

But the battery passes the IEC tests because they only go down to 1mA discharge current.

"The IEC is very slow to reflect the changes in the applications," said Smith. "We want the IEC to reflect what is happening in the marketplace with a wider range of discharge currents. A smoke alarm test is needed. Peoples' lives depend on it."

An IEC spokesperson said:"We are trying to respond to the industry's needs. Duracell participates in the process of producing standards and has access to the lobbying procedures."

The IEC is already extending the tests to cover products using very low currents. However, a test is needed for high current products such as mobile phones.

Richard Ball, Electronics Weekly

Driverless cars on the road to reality

Researchers in the US say they have successfully completed a series of tests involving cars that drive themselves. The tests were conducted along an eight mile stretch of California motorway.

The cars were equipped with onboard computers and video cameras, and were guided to stay within their lanes by small magnets embedded into the roadway at four foot intervals. One driver said that the experience was thrilling for a few seconds but then became "really dull" as he had nothing to do.

The experiments are part of a US national effort to develop fully automated intelligent vehicle highway systems as mandated by a law passed in 1991. A group of high-tech companies have joined with car manufacturers to form the National Automated Highway System Consortium. The group estimates that current motorways could be adapted to handle automated vehicles for as little as $10000 per mile.

There are still trials in various parts of the US awaiting completion before standards for building cars and automated highways are defined.

1996 saw patents increase

The latest annual report from the UK patent office shows the number of published patent applications, at 11 452, increased by three per cent in 1996. Telecommunications remains the most innovative technological sector: 908 patents were published in the financial year 1996 compared to 819 the previous one. Other patent segments include measuring and testing, with 655; and electric circuit elements and magnets, which at 577, grew by 12 per cent.

ATM switch markets are set for growth in both the local area and the wide area

Five times lower power than cmos?

A breakthrough in the way chips consume power, reported by the University of California's Information Sciences Institute (ISI), could dramatically cut power consumption by 80 per cent and make possible new types of highly integrated chip designs.

Researchers within the ACMOS group at ISI have patented a prototype microprocessor called the AC-1, which consumes just one fifth of the power of a similar c-mos processor. The design uses pulsed power and adiabatic charging techniques which recycle some of the power used in the chip's clock cycle.

Researchers are unsure if the same techniques can be applied to commercial microprocessors and other chips, but chip companies will be able to buy licenses for the technology.

Low power consumption is critical to building large, high performance microprocessors where problems of heat dissipation are limiting designs. More exotic types of chips built as a cube could be made possible with low power technologies.
Government warned on traffic policy

Government is urged to establish integrated transport framework to prevent UK missing out on growing telematics markets.

The UK is set to lose out commercially if the government doesn’t hurry up with a policy to establish an integrated transport framework based on electronics, telecommunications and IT. So warns ITS Focus, the UK’s industrial and institutional voice on transport telematics.

ITS Focus believes that the government must speed up its policy as well as making a direct investment to allow the UK to take a lead and participate in the growing markets of telematics (see ‘At a glance’ box). Otherwise, countries such as the US will overtake the UK in seizing commercial opportunities.

"The main thing is that this is a rapidly moving area that offers scope for business opportunities and employment but which will continue to stall unless we have organisational, institutional and technical frameworks," said John Miles, public policy adviser to the European Commission’s DG XIII and consultant to the ITS Focus.

ITS Focus recommends that an overall framework and analysis of different transport and traffic services is tailored to UK needs for a successful implementation – as has already been undertaken in the US under a legislative $30m funding.

ITS Focus has welcomed the government’s research programme on Urban Traffic Management and Control (UTMC) announced last week, and a white paper on integrated transport policy from the newly merged Department of Environment, Transport and Regions (DETR). The research is aimed at developing intelligent control systems to manage traffic in cities.

"We see this this as an important start – it’s along the lines we’ve been setting. However, the resources for the UTMC are quite small for a big ambition. They are £5m over five years," said Miles.

The DETR’s white paper will follow an assessment on the motorway tolling trials recently completed at the Transport Research Laboratory in Berkshire.

What is Intelligent Transport System?

- It can be applied to all means of transport: road, rail, air and sea.
- It relies on electronics, telecommunications and IT – collectively known as telematics.
- It combines traffic management and control, travel and traffic information, electronic fee collection, automatic vehicle location, road safety and route guidance and navigation – amongst others.

ITS report, Tel: 01344 770757
New camera-on-a-chip venture

Specialist flash, E2 and PLD house Atmel is to develop a digital-camera-on-a-chip with Polaroid. It hopes to sell the single chip solution on the open market next year.

The ability to integrate this on one chip comes from ES2 (European Silicon Structures), which Atmel bought in 1995.

Atmel’s task is to implement imaging sensors in c-mos and integrate them with a dsp and both E2 and flash technologies on a cell-based IC. “Our E2 and flash are easier to mix than other solutions,” Atmel’s European boss Bob Henderson said.

“Other solutions” means ccds, which are a high power bipolar technology.

Instead of ccds, the Atmel chip will use low power c-mos sensors.

Polaroid will contribute proprietary technology for colour recovery signal processing, colour filter processing and pixel sensing.

As well as its memory technologies, Atmel will be contributing its a-to-d and d-to-a converter technology, compression circuitry and dsp cores.

Last year, Atmel licensed the Oak and Pine dsp cores licensed in turn from DSP Group in California.

Smart cards in Boots

Boots the Chemist confirms September launch of loyalty smartcard – the first of its kind in the UK. Initially the card, which carries a chip rather than a magnetic stripe, will be used for bonus points but it is expected to be developed further.

“The smartcard may be used as in Germany, for carrying medical information, name, address, doctor’s details and national insurance of the person that carries it, and it can be used as a social security card,” said one Boots spokesperson.

The smartcard, dubbed Advantage, will carry Siemens’ SLE4442 memory chip with 256byte eeprom and a programmable security code. Up to 80 per cent of the cards will be manufactured by GPT and 20 per cent by Gemplus. The smartcard readers will be supplied by Dione Communications.

Boots says it could have used a magnetic stripe like any other retailer, but it wanted to have more than just a loyalty bonus card by using a chip.

“It allows a greater security, it gives an advantage that the card can be used in any store, the holder can spend the points in any store and the card provides a good platform for other services in the future,” said the spokesperson.

Boots is expecting to have up to eight million Britons carrying Advantage by the end of next year.

In 1995, Peter Lilley, then social security minister, told a party conference that the government would adopt a chip-based card in the near future. But since, the government has abandoned the idea of a chip-based card and instead opted for a magnetic stripe in order to save money in the short term. Its plans are to start rolling out its limited-functionality card in three years time.

Svetlana Josifovska, Electronics Weekly

Low distortion a-to-d

Crystal Semiconductor is claiming the lowest total harmonic distortion achieved in an integrated circuit a-to-d converter. Its CS5396, aimed at professional audio use, yields greater than 105dB total harmonic distortion plus noise with a 120dB dynamic range.

Mark Taylor, a company spokesman, said: “The only converters that have a similar performance are rack-mounted units.”

Within the integrated circuit are two chips, one digital and one analogue. Sample rate of the stereo 24-bit device is 96ksample/s and noise-shaping parameters can be down-loaded during use.

○ Asahi Kasei Microsystems, a company that used to have a tie-up with Crystal Semiconductor, has introduced the AK5352, a 20-bit stereo a-to-d converter with a 104dB dynamic range.

Euro PC growth is on a high

The European pc market grew by 16 per cent in the second quarter, says a Dataquest report. The quarter growth, measured against the corresponding quarter last year, is the highest in the last 18 months. “It’s been a very successful quarter but there are too many counter-indicators to predict another boom,” said Steve Brazier, Dataquest’s European PC group’s associate director. The growth will be welcome news for semiconductor companies such as Intel which, earlier this year, warned that it would not meet second quarter expectations, attributed in part to the downturn of the European pc market.

Svetlana Josifovska, Electronics Weekly

798 ELECTRONICS WORLD October 1997
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<th>PARTS NAME \ SPEC.</th>
<th>I.L.</th>
<th>TPY</th>
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<th>Fo(+oM) (dBc)</th>
<th>3dB(on) MHz</th>
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**40GHz** - £1000 + PI only - £600. MF only - £250.
Nanoguitar hits the highest notes

Many people claim to have music in their blood, but how about having a musical instrument in your blood! That claim is at least theoretically possible following an announcement from scientists at Cornell University that they have manufactured the world’s smallest guitar.

It has six strings, each about 50nm or the width of about 100 atoms, and if plucked by an atomic-force microscope, for example – the strings would resonate, but at inaudible frequencies. The entire structure is about 10µm long, about the size of a single cell.

The nanoguitar has actually been made for fun rather than function, and is just one of several similar structures. While the guitar makes a good photograph – it won an award for the best micrograph – and illustrates the progress being made in microelectromechanical devices, or mems, it is the other structures and devices being made at the Cornell Nanofabrication Facility that will be of real utility.

Researchers have made interferometers for example using parallel mirrors, one of which moves relative to the other. These electrically driven devices can be used to modulate the intensity of the reflected light. "This could be of interest for light displays," says Harold Craighead, Cornell professor of applied and engineering physics and former director of the Cornell Nanofabrication Facility, a national resource.

"You could have arrays of these things because they’re so small, with each one independently drivable. We have tremendous flexibility in what we can build."

In the near term, such nanostructures also can be used to modulate lasers for fibre optic communications. These researchers already have demonstrated the ability to make large amplitude modulation of light signals at high speeds.

"We can make reflected light pulses at a rate of 12 million/s," says Craighead. Such a rate is faster than the bit rate of most ethernet connections.

Most microelectromechanical devices are made by photolithography and chemical etching and have minimum feature sizes of slightly less than 1µm. To build devices with dimensions of nanometres rather than micrometers requires a new fabrication approach.

"I know we can go smaller than this. The question is how small we can go and still have dependable and measurable mechanical properties. That is one of the things we would like to know," admits Craighead.

Using high-voltage electron beam lithography, the Cornell researchers sculpted their structures out of single crystal silicon on oxide substrates. A resist is used to pattern the top silicon layer. The oxide that is underneath this layer can be selectively removed using a wet chemical etch. The result: free-standing structures in silicon crystal.
Robot gets ready to hit the road – on Mars

The Nomad rover will be expected to roam much further than existing planetary rover vehicles.

The Nomad rover will be expected to roam much further than existing planetary rover vehicles.

Nasa scientists were rightly excited at the recent performance of their Mars rover vehicle. But to open up exploration of other planets, researchers really need vehicles able to cope much more happily with travelling relatively large distances. Now, they hope they may be on the path to developing one in ‘Nomad’ — a rover that has just set a record by travelling farther than any remotely controlled robot has before over rough territory. The robot’s four wheels logged more than 215km across Chile’s rugged Atacama Desert, during a field experiment designed to prepare for future missions to Antarctica, the Moon and Mars.

Although the straight-line distance on a map was only about 20km, Nomad had to weave through very difficult terrain, and it made numerous side trips for science and to test the meteorite sensors.

The 750kg robot, developed at Carnegie Mellon and funded by Nasa, also validated the use of colour stereo video cameras with human-eye resolution for geology. Autonomous driving is critical for planetary exploration because the communications delay between Earth and planets can be many minutes. With autonomous driving, a robot can explore a much greater distance because it doesn’t have to wait for a person to decide a safe route. The rover is able to see obstacles and recognise them on its own.

Another first for Nomad is use of an on-board panospheric camera to provide live 360° video-based still images of the robot’s surroundings. The high-resolution video camera focuses up into a hemispheric mirror and takes a 360° picture — one frame per second. The video view includes all of the ground up to the horizon in the circle surrounding Nomad.

Electronics take over house calls

The prototype for an electronic house call system that could reduce the number of visits to chronically sick patients made by nurses and GPs has been developed and tested in the US. The system uses established hardware and cable communications to set up a two way link between patient and medic.

Researchers fashioned the prototype from existing computer hardware, with additions such as a multi-function patient monitor — like one used in intensive care units — into which blood pressure cuffs, stethoscopes and other medical devices are plugged.

A commercially available video-conferencing program enables the nurse and patient to see each other and talk throughout the examination. The system also accepts data from a variety of medical devices, such as those registering blood pressure and blood oxygen levels, so that the nurse can listen to heart and lungs and perform an electrocardiogram.

Development and initial testing of the electronic house call system has so far been funded primarily by grants from the Department of the Army and the Georgia Research Alliance.

For the most part, patient feedback has been positive. Patients seem to like the idea of being monitored at their homes. Though the system would mean less face to face contact between medics and patient, the interaction with nurses through the system could actually be greater.
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Electronic guide dog steers round obstacles

You can’t pat it and it won’t nuzzle up to you. But a new sonar-equipped navigation aid for the blind being developed at that University of Michigan College of Engineering’s Mobile Robotics Laboratory claims to be able to detect obstacles in the user’s path as well as any dog, and automatically steer around them.

The device, invented by Johann Borenstein U-M research scientist in mechanical engineering and applied mechanics is called a GuideCane, and a preliminary version of a working prototype has already been tested by visually impaired individuals. Users reactions are said to have been extremely positive, though more development will be required before the device is ready for widespread commercial use.

The 3.5kg GuideCane consists of a long handle with a thumb-operated joystick for direction control, an array of ultrasonic sensors and a small on-board computer mounted on a two-wheeled steering axle. Users push it ahead with one hand. When the device’s ultrasonic sensors detect an obstacle in its path, the computer automatically turns the wheels to steer around the obstacle and resume the original direction of travel.

Steering changes are experienced as a direct physical force through the handle, which should make it easy to follow the GuideCane’s path without conscious effort. The body automatically follows the trajectory of the guide wheels just as a trailer follows a car.

Once the obstacle is cleared, the guide wheels resume their original direction.”

The Michigan researchers hope that, after a brief adjustment period, most people will become so comfortable that they will be able to navigate around obstacles at their normal walking speed.

Clock watching can save chip energy

Scientists at the University of Southern California have successfully demonstrated a chip that uses as little as 20% of the energy of conventional models. The new microprocessor – designed by research professor William Athas and colleagues at the USC School of Engineering’s Information Sciences Institute (ISI) – recycles energy from the chip’s clock, the timing signal normally used to synchronise computing functions.

In conventional chips, the clock consumes a large fraction of the total energy supplied to the chip, all of which eventually winds up as heat.

ISI’s experimental chip, called AC-1, has two different clock circuits. It can work with an ordinary clock mechanism, or a flip of a switch will activate circuits that briefly convert the energy of the clock’s electric signals into magnetic form. This captured energy is then reconverted into electrical form and returned to power the data-processing sections of the chip. The energy savings range from 75% to 80%, depending on how fast the clock is set to run (with slower clock settings yielding greater energy savings). Yet in its energy-recycling mode, the chip is able to perform the very same computing tasks it performed using the conventional clock circuit.

The idea of using the clock to reduce power consumption was first proposed, in 1967, and in the mid-1980s, ‘hot clock’ chips were proposed and designed for fabrication by the then-standard nmos technology.

Prototypes were made, but nmos was soon eclipsed by cmos, and many researchers thought the hot-clock system would not work. But the ISI scientists hope their success will spark new interest in the idea.

Initial applications of the energy-recycling chips will probably be at the lower-performance, cost-driven end of the chip market. But applications could include portable computers, digital watches, cell phones or GPS position finders. The only difference would be that the batteries would last much longer.

Eventually the researchers expect to double the energy savings with chips that consume only one-tenth the power of conventional chips.

For more information contact: William Athas, School of Engineering's Information Sciences Institute, University of Southern California, 3620 South Vermont Avenue, Los Angeles, CA 90089-2538. Tel: 213 740 7600.
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Fig. 1. This anomalous train of waves, detected in Italy, lasts for 14 seconds and repeats at irregular intervals. At times it is rather frequent – one set of waves every 5-10 minutes – but at other times it is absent for days. The frequency is 1.87Hz and it is relatively strong if compared with the other signals in the background. Its origin is anybody’s guess, but it is likely that some kind of electric machinery is at work.

Detecting ultra and extra low frequency signals needs something different from a tuned LC circuit. Designs incorporating coils weighing several pounds and unyielding capacitors are occasionally implemented by brave experimenters. But there are alternatives if the electric, magnetic and possibly mechanical waves are dealt with separately.

Aperiodic detectors do not tune to a specific frequency and leave the job of frequency determination to other equipment like spectrum analysers. As a consequence, the design can be optimised for best performance in a given application.

Electric fields

Figure 1 shows an oscillation of unknown origin at 1.87Hz. I occasionally detect it with a 3m long antenna just outside my home.

Suspects for the source of the signal are an appliance from a neighbouring house and electric industrial machinery in the area. But, so far I have been unable to discover the source.

When I pushed the detection circuit’s sensitivity to its limit, I was able to see all sorts of apparently unrelated signals. The only signal that could be positively identified was a sharp, clean positive peak showing that someone in the area had switched on a television set.

These low-frequency signals are detected by means of a high impedance circuit, Fig. 2, based on the OPA124P. This differ op-amp is on the expensive side, but its characteristic of stability and low noise make it the ideal choice.

Problems met in the design of this circuit were twofold: interference from the mains and the connection to the antenna. The mains interference problem was solved by inserting a low-pass filter followed by a 50Hz notch filter. Together, these eliminate all mains induced signals – including harmonics.

I took care to make the circuit provide good global transient response with little or no overshoot. Removing or changing one or more of the filter components will change the bandwidth, which is likely to worsen the transient response. Because of the filters, the bandwidth extends to about 10Hz. The system is still useful up to 16Hz, where gain is -10dB, at which point starts to fall off rather rapidly to 50Hz.
SENSORS

showing, on the simulator, an attenuation of -91dB.

The low-frequency -3dB point is 0.1Hz. This means that dc signals are not amplified. In fact the amplifier is ac coupled. This eliminates all kinds of problems relating to dc stability and coupling, as well as any effect of large electrostatic fields present around the antenna.

It is, however, necessary to balance the dc point of the OPA124P in order to compensate for any change induced by the setting of the potentiometer. This is the only adjustment required by the circuit.

When the implementation is complete, switch it on, wait five minutes and then adjust the 100kΩ potentiometer until there is no dc variation when rotating the 220kΩ potentiometer near its maximum sensitivity.

Catering for high impedance

The second problem was that, due to the very high impedance of the circuit, any length of wire or coaxial cable connecting the input to the antenna would introduce noise and microphonic.

After several tests and trials, I found that the best solution was to connect the antenna directly to the input of the amplifier without any intervening wire or cable. This meant that the first IC had to be mounted directly next to the antenna. Connection to the rest of the circuit is via a five-core shielded cable carrying power, ground, signal and the two connections for the potentiometer.

This solution works very well. The distance between the antenna head-end and the rest of the circuit could be quite long, although I only tested it up to 10m.

Both the setting of the 220kΩ potentiometer and the length of the antenna influence the overall sensitivity of the circuit. It might be convenient to install a telescopic antenna and make the unit portable. This is why I designed the system with two PP3 9V batteries in mind.

Useful signals are detected even with a 5cm antenna, although the interference generated by just walking around is strong enough to bury the signal. A quick test can be carried out using the set up shown in Fig. 3. Spikes should be detected whenever one of the poles of a 9V battery is alternatively brought in contact with the screwdriver. The field is enough to generate 1V peaks at the output.

The input of the circuit is protected by low leakage diodes and a series resistor. This is very effective against any overvoltage or mishandling of the input circuitry.

The output goes first through Schottky diodes. This gives a convenient threshold against very low level signals. The result is a zero signal that looks very clean - quite effective if you are using a pen plotter.

Crossover distortion introduced by the diodes is very low and is evident only at very low signal levels. As an alternative you might like to connect the output lead directly to the output of the amplifier. This bypasses the rectifier bridge and gives the full signal - noise included.

Fig. 2. Complete circuit for the electric field detector.

Components marked with an asterisk should be 1%. If 5% components are used, then one of the 47kΩ resistor should be changed to a series of a 39kΩ resistor and a 22kΩ trimmer adjusted for the minimum mains noise. Values between brackets are suitable for a 60Hz mains. An American equivalent to the BC337 for the audio oscillator is the 2N2222.

continued on page 808...
An acoustic signal is available from a frequency modulated audio oscillator. This is useful if you intend to use the meter as a portable instrument.

Note that mains induced noise can be quite high and could easily saturate the preamplifier. Consequently there is a practical limit to the length of the antenna. In the metropolitan area where I tested the meter, a 3m enamelled copper antenna generated 5V pk-pk at the output of the preamplifier, before the filters. This means that probably a 6m antenna is the longest that could be used under these circumstances. In a rural area the situation will certainly be better, allowing longer antennas.

**Implementing the circuit**

On a practical note, if you have difficulties obtaining the 500MΩ resistors, connect 100MΩ resistors in series. I recommend that you use high quality input capacitors, mylar or similar, but definitely no ceramic capacitors.

The Burr-Brown data sheet for the OPA124P suggests using a printed ground ring around the input lead and connection to ground of the substrate, pin 8. It also suggests direct soldering of the IC to the circuit board.

The preamplifier was assembled in a small metal box with the shielded cable on one side and the antenna wire on the other. If it is to be permanently placed on the outside, a convenient shelter should be foreseen against weather extremes. The temperature range of the IC, from -25° to 85°C, should not cause any problems in normal operating conditions.

Some form of data logger or a pen plotter is essential part of the system. I used the ADC42 data logger from Pico Technology, connected to the parallel port of a PC. The ADC42 software also includes a spectrum analyser virtual instrument, so it is a convenient way to analyse the signal.

I now have several traces of strange signals, oscillations, spikes — even square waves. My next task is to correlate them with known events such as earthquakes or signals that are supposed to make the round trip of the world generating a 7.5Hz oscillation.

**Magnetic lines**

The sensor that first springs to most people's mind when they think of implementing an electronic compass is a Hall-effect device. But there are alternatives — like the one suggested in Fig. 4. This is a classic oscillator running at around 1MHz.

The ferrite rod is almost identical to the one normally found in medium wave receivers — the longer the better. The only difference is the strong magnet attached to one end of the rod.

With the magnet attached, the exact frequency of oscillation now depends on the prevailing magnetic field. If you have a second oscillator, it is possible to detect the beat note which depends on the orientation of the magnetised oscillator with respect to the Earth's magnetic field.

An easy test requires a medium-wave radio tuned to a station, preferably a weak one. Next, tune the oscillator in such a way to generate an audible beat note in the radio. Any minute movement of the rod will now change the audible beat frequency.

I plotted the frequency variation against rotation of the rod, Fig. 5. The first curve shows the frequency of the beat note starting from 200Hz and the rod pointing north. The second curve shows the beat frequency, starting from 200Hz but with the rod pointing east.

In both cases, the rotation was clockwise and extended for 180°. For this test, I used a second oscillator because the interference of the transmitting station was too strong.

A complete circuit with two oscillators and ancillary circuitry was the subject of a circuit idea published in *Electronics World+Wireless World* in December 1994.

When prototyping the circuit, take care to use adequate electric shielding around the
oscillator. Metal boxes are not suitable. The metal would create loops that would kill the oscillation. A solid plastic box with copper wires or strips, glued on the inside, running parallel to the rod and soldered to ground at one point only is the best solution. A small hole is made in order to adjust the trimmer capacitor with a plastic screwdriver.

As this detector reads the effective direction of the magnetic lines, you will have a change in frequency for any movement of the oscillator both in the horizontal and vertical planes. Listening to the beat note shows an incredible variety of signals, Fig. 6. The situation is much quieter in rural areas, but in a large town you have to wait until late at night - between 1 and 5 o'clock in the morning - if you want to see less frenetic magnetic activity, Fig. 7.

Where all this activity comes from is a mystery to me, although I am sure that most of it is man-made. Frequency changes, hence magnetic variations, take place at a very low rate; 4-5Hz is the fastest you may expect. This detector operates from 0Hz, so it will track changes lasting hours, limited only by the long term stability of the oscillator.

The 1MHz oscillation frequency was chosen for convenience. Its radiated field will not be detected by a radio beyond a 2m radius. Nevertheless, a more suitable frequency could be used. Bear in mind that a higher frequency will increase sensitivity but stability might become a problem. Conversely, lowering the frequency makes the circuit less sensitive but more easily stabilised.

Seismic waves
The OPA124P and the idea behind the electric field detector was the basis also for the seismic detector, Fig. 8. The circuit is straightforward and requires no adjustment. The amplifier is local to the detector element. This solves the problem of connecting a high impedance source to the amplifier without introducing any further degradation.

A three-core shielded cable connects the amplifier to the other elements of the circuit and to the power supply, which is ±15V. The detector is the piezoelectric element of a kitchen gas lighter. Although its electrical characteristics were unknown to me, it performs well. The problem with this circuit lies in the proper mechanical construction of the detector. It should be on a solid base stuck in the ground, with the detector placed vertically with a 1-1.5kg weight on top. This weight should not have a proper resonant frequency, so a loose package of sand, fine gravel, or even salt, sealed against humidity, works fine.

The weight should be in contact with the detector only and kept in place using vibration absorbing material such as shock absorbant rubber, plastic sponge and the like. The whole set up should be screened against electric fields with a non magnetic screen such as aluminium or copper, and should be enclosed in a sound proof box, shown over.

Of course, you do not have to go to such an extent to see the circuit's basic operation. You can place it on a table if you are happy to see...

Fig. 6. Printout of from data logger shows the variation of the beat frequency over a period of 113 seconds. Samples where taken at 1s intervals at 6pm in a built-up area. In open country the response is much quieter with several minutes without appreciable changes.

Fig. 7. Data collected from the magnetic detector over a period of 16 hours, from 6pm to 10 am the following day. Large variations take place during the whole day except for a short lull around 4am. This is also the best time to carry out testing of the unit. Data were collected in a large city. The same device shows an almost flat curve only in open country, away from human activity.

Fig. 9. Seismogram of a lorry, barely heard in the silence of the night, passing in the area and measured over 10s. The frequency range of the signal is mostly between 10 and 50Hz and is bound to interfere with a real seismic wave. This demonstrates that careful choice of the installation site is important if you want to record natural phenomena.
Fig. 8. There are no settings or adjustments in this seismic detector. The screened portion is placed together with the piezoelectric detector and connected to the remaining part of the circuit with a three-core shielded cable. Piezoelectric gas lighters are economical and readily available. The element from any such lighter should be suitable for the detector.

the resonant frequency of the table—and probably your house. Two LEDs flash alternatively when a seismic wave is detected and the overdriven meter gives a clear reading, even with small signals.

Road and railway traffic are the main interfering signals. They could mask a real seismic wave. You can learn to distinguish the signature of interfering signals but it will still be a problem to recognise a seismic wave simply because they do not happen frequently enough—fortunately.

The frequency range of the circuit is from 0.44 to 11Hz with a -10dB point at 24Hz. In seismological terms, this is a narrow band detector, as opposed to a 0.1 to 100Hz alternative, which is defined as a wideband detector. This detector has a smooth filter so signals outside the nominal bandwidth will be detected. But as most of these signals are human artefacts, Fig. 9, a certain amount of attenuation is preferable.

Morning quiet
As with the magnetic detector, I noticed a quiet period during the early hours of the morning with the seismic detector, Fig. 10. At other times, it is as frantic as the magnetic or the electric detector. There is probably a relationship that only a multichannel recorder could confirm.

Sensitivity seems to be at the right level. With an overall gain of 60dB, the circuit amplifies a hiccup in close proximity, but I did not attempt to properly calibrate the device.

The seismic waves detected are the vertical component of the surface wave. Adding two further detectors, rotated by 90°, would also allow you to record the horizontal components.

Take care when handling the detector. Hitting it even gently will generate hundreds of volts—thousands of volts in the worst case. So take safety precautions to protect yourself, and the input circuit. The only protection is via the 3.3MΩ series resistor. This protection is sufficient for normal handling of the detector but it is advisable to keep the input shorted during installation.

If the purpose of the above circuits is to detect natural phenomena, then they should be operated away from cities or industrial sites, unless of course, you are looking for man-made activity. But wherever you are, you are bound to discover many of the mysteries, quirks and anomalies of the world of ultra-low frequencies.

Practical implementation of the seismic detector. Most of the circuit is housed together with the detector. There are no critical components. The rubber that keeps the weight in place must not transmit vibrations coming from the ground. Spongy rubber works fine. The sound proof enclosure prevents low-frequency sound—typically from helicopters—from interfering.
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More by accident than by design, the British Post Office became involved in many strange activities during World War II. Its work devising the Colossus, Cobra and Tunney devices for breaking enemy codes at Bletchley Park is now fairly well known, whereas similar work producing high-speed computers for co-ordinating anti-aircraft gunnery and for timing bomb release has not received the same recognition.

Other less well-known episodes include the freight trains run by the Post Office tube railway in London. These carried spoil from new bunkers under construction and brought it to the surface away from the sensitive location. And there was the assistance the Post Office submarine cable team provided for the design of 100 miles of hollow three-inch cable which carried vital fuel across the Channel to support the Allied armies in the invasion of Europe.

But perhaps the most unlikely of all unusual assignments was the Post Office's involvement in the development of hi-fi recording techniques. During World War II, materials for recording radio programmes were so scarce in Britain that the Post Office was forced to commandeer a factory to produce blank recording discs. In so doing, its boffins made significant improvements to recording techniques.

"The BBC alone used 7000 blank discs a week"

The BBC relied on vast quantities of blank discs for sound recording. The corporation alone used 7000 a week for making transcriptions of its own programmes and the enemy propaganda monitored by its listening stations. The armed forces – and the secret services, if rumoured – also made heavy demands on the supply of direct-cut discs; they were used for covertly recording the conversations of prisoners-of-war and also for capturing weak radio signals. By replaying the weak signals many times, it was hoped that they would eventually be read correctly.

The reliance on these discs was so great that it was felt necessary to put their manufacture under closer direction. To safeguard supplies, the Post Office took control of one of the two factories which produced them. They and the BBC also improved record cutting head technology and a microgroove recording cutter using these blanks was first used on VE-Day.

This is to anticipate events, however. The story really starts back in the mid-1930s when Cecil Watts invented a sound recording technique using a metal disc coated with cellulose nitrate lacquer. People often call these discs 'acetates' but this title is a misnomer.

"With its high signal-to-noise and power versus bandwidth ratios it has not been improved upon"

This direct recording disc was and is a remarkably high quality sound recording medium. With its high signal-to-noise ratio and power versus bandwidth ratio it has not been improved upon. Surface noise is low and in fact the technique is still used today for mastering some albums. The discs themselves were fairly soft, the disc cutter being a sapphire cutting tool.

But this was not the only method. In 1939 the commercial record companies were still using wax for mastering – an inferior process. The BBC, however, use these cellulose nitrate discs for short-term sound archiving. In addition, they had two other recording systems at their disposal. These were the Marconi-Stille – which recorded magnetically onto steel tape – and the Philips-Miller – which made an optical sound track on a material similar to cinema film.

With the war under way, these other two systems hit problems. The special steel for the Marconi-Stille recorders came from Sweden but supplies ceased in August 1940, while the optical film for the Philips-Miller machines came from Gevaert in Antwerp. After this city had fallen to the Germans in May 1940, supplies of this tape also ceased.

Alternative sources were tried for both technologies but, none was ideal. This left the BBC dependent on the direct recording cellulose nitrate discs alone at a time when the demands on recording for broadcast purposes was mushrooming.

"In effect, the manufacture of these discs was nationalised"

Unprepossessing they may look but blank discs – similar to these – for one-off recordings played a vital role during WWII. So important were they to the war effort that the Post Office commandeered a factory to safeguard supplies (photo from National Sound Archive).
There were only two sources of the blank discs, EMI and Cecil Watts’ firm, MSS Recording Company. In the summer of 1941, it was decided that the MSS factory should come under the control of the Post Office, which also provided additional capital to boost output. In effect, the manufacture of these discs was nationalised.

Two scientists from the Dollis Hill research station, J.F. Doust and F.G. Hopwood, were put in charge of this activity. In a separate initiative, the BBC devised an improved recording cutter and this cutter used these MSS blanks. The frequency response extended beyond 10kHz and so it could be called high fidelity, later on it could and did cut microgroove records. The new system was first used on VE-Day.

As well as blank discs, Cecil Watts’ MSS company had made disc-cutting machines since the mid 1930s. These also came under the control of the Post Office when they took charge in 1941.

Improvements to the MSS disc cutters were also made during the period of Post Office management. It can truly be said that the Post Office played a part in improving recording technology—even if it was more by historical accident than by any intentionally planned intervention. The Post Office did not, however, invent the microgroove record, as some people have alleged.

“...the story of the GPO’s wartime involvement in acoustics doesn’t end here...”

The story of the GPO’s wartime involvement in acoustics doesn’t end here, though. There were many other interesting projects. Special noise-cancelling microphones and headphones were devised by the Post Office for tank use.

Deep below the Post Office’s Dollis Hill research station, a special chamber with deafening sound effects was constructed to simulate a tank’s interior. Producing the sound effects with which to test these devices was another matter; recordings on 78rev/min gramophone records would not last long enough and the chosen solution made good use of other Post Office technology.

First of all, recordings of a tank rumbling past a microphone were made on direct-cut disc. Ten-second recordings made in this way were transferred to 35mm film by the Crown Film Unit—a new name for the old Post Office Film Unit—and a two-second long section finally transferred to glass disc, for playing continuously on a speaking clock machine.

Apparently the joint in the recording and the second-repetition were not noticeable in use. Similar machines were made for the Royal Air Force; these generated continuous background aircraft noise effects for training radio operators. Four different aircraft types were covered, at normal speed and at absolute maximum.

Yet another machine was developed for training fighter pilots; this was a twin-channel simulator, providing continuous aircraft noise and spasmatic machine-gun effects, as and when required.

At the end of hostilities all this development activity came to an end. The MSS factory was handed back to Cecil Watts and all these matters were effectively forgotten. At this distance in time nonetheless they form a fascinating sideline in the history of sound recording and reproduction.

My thanks go to Peter Copeland of the National Sound Archive for his assistance in preparing this article. Some details were taken from the book ‘BBC Engineering 1922-1972’ by Edward Pawley, and others from an IEE paper by Arnold Lynch entitled ‘Some Derivatives of the Speaking Clock.’
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Loudspeakers are everywhere. So it's not hard to study and compare them. Mostly I compare them with the original sound. Even casual inspection reveals that, with rare exceptions, most of today's loudspeakers fall far short of what is desirable — and look awful into the bargain.

There is no technological reason for this state of affairs; in fact we have never been better served with advanced materials and the means to form them.

Looking on the bright side, if the vision exists, it only requires some good industrial design to have much better sounding and looking loudspeakers than are generally available. In this series I intend to explore some possible avenues.

The only fundamental criteria we have for loudspeaker performance are purely subjective. It follows that a loudspeaker designer must understand psychoacoustics, quantitative as well as qualitative. As human beings vary, we have also to understand statistics in order to appreciate that psychoacoustic data are not absolute like classical physics but are subject to variation.

If psychoacoustics helps us to define what performance our loudspeaker must have, it does not tell us how to achieve it. We can stir the air in a variety of ways, but seldom directly. Generally we need to cause some object to vibrate, and subsequently transmit those vibrations to the air. The study of what happens next is known as acoustics.

Causing precisely controlled vibration requires a knowledge of how masses move. Control theory tell us enough about electromagnetic and electrostatic actuation of diaphragms. To follow what goes on inside the speaker we need to know thermodynamics because we are compressing a gas, and it matters whether we do that isothermally or adiabatically.

As listeners we are primarily interested in what happens outside the speaker; in the far field to be precise. It is the job of the cabinet to separate these two results: it must be utterly inert and resist the reactions from transducers and the internal pressures without any movement or flexing. Most of today's loudspeaker cabinets are very poor in that respect, yet a basic knowledge of structural engineering and materials science will help to design a suitably rigid structure.

In order to anticipate and control the high frequency sound output of a loudspeaker, it is important to understand the wave or diffraction theory of acoustics. The concepts of coherence, phase and interference, even chaos theory, become crucial to stereo imaging, yet are widely neglected. In fact light and radar behave in exactly the same way as sound — although on different scales — because they are wave motions. A knowledge of wave optics is extremely useful in loudspeaker design because it allows the diffraction behaviour of cabinets to be predicted.

As passive speakers simply cannot meet modern criteria, the ability to design electronic systems has become essential. This encompasses not only amplifier design, but also active filter and analogue computing implementations. Emc awareness is important too, but knowledge of speaker cables can be safely set aside as we won't be using any.

Although I consider the above list of skills to be a minimum for the development of advanced loudspeakers, such combinations are rare. Little wonder that the average loudspeaker today is so primitive. Many are developed using pure empiricism. The mechanism is not understood but the parameters are changed until it sounds best.

Unfortunately a speaker designed in this way may only sound good on a certain type of music. The design may then appear in a cookbook to be copied parrot fashion with even less understanding of the principles.

Often the designer is forced by economics to use sub-optimal techniques. Here the home constructor may be at an advantage because he can fabricate complex shapes which could never be produced economically.

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CIRCLE NO. 117 ON REPLY CARD
Oscillators with very low phase noise are crucial in modern, fast data radio links with high spectral efficiency. Relatively easily achieved in a crystal oscillator, low phase noise is a much more daunting design task in a wide range vco, as Ian Hickman explains.

How do you make a ‘quiet’ – i.e. spectrally pure – oscillator? In an earlier article¹, I highlighted the need for oscillators with very low close-in phase-noise sidebands. The characteristics of an oscillator to fill this requirement are well known, and include the following:

- a high-Q resonator – tuned circuit, crystal, SAW resonator or microwave cavity;
- a low-noise transistor or other active component as the maintaining amplifier – especially with low $1/f$ noise, as this cross modulates onto the output;
- a high-gain maintaining amplifier stage so that it can be lightly coupled to the resonator which thus works at close to its unloaded $Q$;
- a buffer stage for the output, again enabling the resonator to work at close to its unloaded $Q$;
- a large amplitude of oscillation to ensure the stored energy at the desired frequency is greatly in excess of circuit noise;

My earlier article showed that when the resonator is an LC ‘tank circuit’, with the limited value of $Q$ thus obtainable, the phase noise is lower if the design is such that the transistor maintaining amplifier is not allowed to bottom. I was curious to know if this was still the case with a much higher $Q$ resonator, such as a quartz crystal.

To enable a direct comparison to be made, I used the same oscillator circuit as in Fig. 2 of the previous article, but modified to incorporate a 10MHz crystal, Fig. 1. In the earlier circuit, the junction of the two capacitors tuning the inductor in a Colpitts oscillator was grounded directly. In the new circuit of Fig. 1, it is grounded via a 10MHz crystal.

Given the low equivalent series resistance of the crystal, the circuit operated much as before, once the tank circuit was retuned from its previous value of 10.4MHz to the series resonant frequency of the crystal – a shade below 10MHz. The series resonant frequency $F_r$, the equivalent series resistance $R_e$ and other parameters of the crystal were determined beforehand, as detailed in the panel.

Quartz to the rescue

The output of the oscillator, now crystal controlled, was connected to the phase noise discriminator as shown in Fig. 2 of the earlier article. The only change made to the phase noise discriminator was the addition of an extra 2.7m of coaxial cable to the delay cable. This was to bring its length back to $15/4$ at the lower 10MHz operating frequency.

Incidentally, the length of the cable used for the earlier 10.4MHz oscillator was quoted in error in the previous article as 108.4m. That was in fact the equivalent length in air: the physical length calculated out as 72.3m. Thus the cable represented the best part of what was originally a 100m reel.

I connected the crystal oscillator of Fig. 1 to the phase noise discriminator. Everything was set up and left working for a while to reach equilibrium. The phase noise out to 5kHz offset from the oscillator frequency was then measured using the HP3580A audio frequency spectrum analyser, and the trace stored in its digital store.

As in the previous article, a 56kΩ resistor was then connected in parallel with $R_1$, resulting in some 10dB increase in output level, with the oscillator transistor bottoming each cycle. The phase noise out to 5kHz was again measured and displayed, this trace and the stored trace both appearing simultaneously on the HP3580A.

The resulting display was photographed, Fig. 2a). As you can see, the two traces are virtually identical. Indeed, it is only the two
odd spikes of mainsborne interference on one of them, three quarters of the way across the screen, that show there really are two separate, superimposed, traces.

Furthermore, both traces show lower phase noise than the non-bottoming version of the oscillator without crystal control – at least above 1kHz offset. This was shown in the previous article, and is reproduced here as Fig 2b), lower trace.

The conclusion, then, is clear. A resonator with a Q of 40 000 can make up for the deficiencies of a mediocre maintaining oscillator. But on the other hand, the tuning range of a crystal controlled oscillator is limited to a few parts per million at most.

Where a greater tuning range is needed, as is usually the case in the voltage-controlled oscillator of a synthesizer, then a non-crystal controlled LC oscillator is required, and a non-bottoming maintaining amplifier is much to be preferred.

Avoiding the noise?
Even when the maintaining transistor of an oscillator does not bottom, it still operates in a non-linear mode, cutting off for part of each cycle. This results in noise – particularly the device’s white noise – cross-modulating onto the desired sinusoidal output. The lower order non-linearity of the non-bottoming circuit results, as has been shown, in lower close-in phase noise.

Normally, oscillators rely on nonlinearity to stabilise the amplitude of the oscillation. Initially, at switch-on, as the amplitude builds up from nothing, the transistor operates linearly. The loop gain at the operating frequency is in excess of unity. But as the amplitude increases further, the device is driven into non-linear operation, and the loop gain at the fundamental falls to unity. This is illustrated in Fig. 3b), which shows the open loop gain of the stage at the centre frequency of the tuned circuit.

In principle, this can be plotted by opening the loop and applying a gradually increasing external signal at the closed loop frequency of oscillation. But of course, to be representative, the signal must be applied from the same source impedance which the transistor sees in closed loop operation.

Likewise, the loading of the transistor’s output impedance on the tank circuit must also be representative of closed loop operation. Both of these conditions could be met by breaking the loop and embedding the oscillator in a long chain of identical stages, as shown in Fig. 4.

The highlighted stage sees the appropriate source and load impedances, and if the chain is long enough, it will operate at unity gain. For the earliest stage will selectively amplify the noise at the centre frequency of the LC tank circuit, and subsequent stages will further amplify and band limit the signal, so that the boxed stage runs in limiting at the same amplitude as if it were operating closed loop.

If you think such a long chain of identical circuits is too expensive a proposition, even conceptually, a much shorter chain – with the output of the last stage connected back to the input of the first – will do as well or better. For by definition, each stage operates at unity gain and zero phase shift. A ring of just three, or even two such stages illustrates the idea.

Figure 3a) shows the open loop gain versus amplitude of a surefire oscillator, one where the gain at low amplitude is well in excess of unity. A badly designed oscillator with a characteristic like Fig. 3b) is occasionally encountered, and infuriating it is, too. It usually starts up all right, due to the switch-on transient kicking it into life. But just occasionally it fails to start – usually just when it’s needed most.

It is tempting to think that an oscillator
A bright idea tested

In my earlier article, I mentioned the Comlinear CLC5523 variable-gain amplifier from National Semiconductor, together with its possible application as the maintaining amplifier in an rf oscillator. The scheme proposed was to use the device enclosed in an automatic level control loop, so that the amplitude of oscillation was controlled without any non-linear operation on the part of the maintaining amplifier.

Within the CLC5523 are a closed-loop input buffer, a voltage controlled gain cell and an output amplifier. The input buffer is a transconductance stage, whose gain is set by the gain setting resistor $R_g$, Fig. 5a). The output amplifier is a current feedback op-amp and is configured as a transconductance stage whose gain is set by, and equal to, $R_g$.

With its 1800V/µs slew rate, the CLC5523 offers a 250MHz bandwidth and adjustable gain. Figure 5a) shows the device connected as a non-inverting amplifier, the maximum voltage gain $-1$ with $V_{gs}=2V$ being set by the ratio $R_g/R_o$.

As $V_g$ is reduced to zero, the gain falls to a minimum, with up to 80dB of gain reduction possible, depending on the operating frequency. In linear terms, i.e. gain measured in V/V, most of the gain change occurs over the range 0.8V $<$ $V_g$ $<$ 1.2V, while in logarithmic terms, i.e. gain measured in decibels, the gain variation with $V_g$ is linear from 10dB below maximum gain, i.e. +1V, downwards.

Figure 5b) shows the device pin-out. This is the same for both the DIP and SOIC versions, but note that when using the DIP version, pin 4 must be grounded via 25Ω rather than directly, Fig. 5a).

After some considerable time and development effort, I arrived at the circuit of Fig. 6. The CLC5523 is set for a maximum gain of $x_{10}$, determined by the ratio $R_{22}/R_{20}$. Oscillation is maintained via a two turn feedback winding $L_2$, $L_1$ having some 14 or 15 turns.

Considerable care is needed where the automatic level control, or alc, loop is concerned.

Quartz crystals for frequency control

A quartz crystal makes a high-Q resonator, which also exhibits a very high degree of frequency stability with time and changes in temperature.

When excited at its resonant frequency, the crystal vibrates — rather like a tuning fork but usually at a much higher frequency. Like a tuning fork, when the excitation ceases, the vibration slowly dies away, over many cycles. The higher the 'Q' or quality factor, the slower the decay.

In fact, Q is the ratio of the energy stored to the energy lost per radian. Thus if the Q of a crystal were say 10000 — a very modest figure — then 0.01% of the stored energy would be lost per radian. After 10 000 radians, or 1592 cycles, the energy left would be down to $1/100$ or 37% of the original amount.

A high quality crystal resonator would have a Q of 100 000 or more; the very best approaching 500 000. The best raw natural quartz exhibits a bulk Q of over a million, but adding the electrode arrangements always reduces this.

The plate of quartz behaves like a series tuned resonant circuit, the resonant frequency being determined by the values of the motional inductance $L_m$ and the motional capacitance $C_m$. These are the electrical equivalents of the mass and compliance — springiness — of the quartz.

Equivalent inductance and capacitance are shown in Fig. A, together with $R_s$. This represents the lossiness of the vibratory material, determining how quickly the vibration dies away in the absence of excitation. It is known as the equivalent series resistance $E$.

The crystal can be maintained in a steady state of vibration by applying an electrical signal across it. The piezo-electric effect causes minute changes in the crystal dimensions at the frequency of the applied signal. The amplitude of the vibrations is negligible except at the natural resonant frequency of the plate of quartz, the resonant frequency being determined mainly by the size and thickness of the plate.

In addition to $L_m$, $C_m$, and $R_s$, Fig. A also shows the capacitance $C_q$. This represents the capacitance between the two areas of metalisation, one on either side of the plate. These serve two purposes. They couple the electrical stimulation to the crystal, and there are also wires bonded to them which serve to support the crystal.

The mounted crystal is usually enclosed in a container of glass or metal, filled with air, dry nitrogen or preferably — a vacuum.
Fig. 6. Circuit of a 10.4MHz oscillator with a linear maintaining amplifier, the amplitude of oscillation being controlled by an ALC loop.

Any attempt at adding an rf bypass capacitor following the diode detector D3 resulted in 'squegging', i.e. oscillation of the level-control loop resulting in bursts of rf, each followed by a recovery dead period. Any low-pass filter, whatever its time constant, must follow the loop error amplifier TR1. Thus the dc level out of the schottky diode detector D3 still shows a considerable level of rf ripple.

Both the ripple and the dc level—the latter adjusted by an offset control R14—are applied to the loop error amplifier TR1. This compares the adjusted dc level with 0V ground and reduces the gain-control voltage Vg applied to pin 1 of IC1 to maintain the desired level of oscillation. Note that even if TR1 cuts off completely, the potential divider formed by R17 and R18 prevents Vg exceeding +2V.

A requirement for proper operation of the device is that Vg should not exceed +2.5V. If it does, the gain control circuitry may saturate, and the gain may actually be reduced. With the circuit shown, the loop was very effective. There was no measurable change in output.

Measuring crystal parameters

Crystal manufacture is a specialised business, although traditionally amateur radio enthusiasts used to grind their own, to get the exact frequency they required.

A crystal manufacturer has the specialised equipment to measure all of the parameters shown in Fig. A, and others—such as temperature coefficient, ageing, etc. Such equipment is not generally available in the usual electronics development laboratory. But if you do want to know the parameters of a given crystal, as I did in connection with a 10MHz crystal oscillator described in this article, it is not a difficult matter to measure them. The usual laboratory rf instruments, plus a simple jig involving a little purpose-built circuitry, will do the job.

Figure B shows the arrangement I used to characterise a 10MHz crystal dredged from my stock. It was made by an East Anglian crystal manufacturer which is no longer in business. It was salvaged in the early sixties from a naval missile tester. There, it formed the reference in a purpose-designed digital voltage/frequency/period/etc. meter that was the heart of the system. Naturally, in the circumstances, detailed information on it is no longer available; only the maker’s name, the frequency ‘100000’ and a batch number stamped on the solder-seal metal can.

I measured the series resonant frequency Fs using the crystal as the shunt leg in a tee-attenuator pad, the series legs being 100Ω resistors. These resistors should be as small as possible and certainly not wirewound. Chip resistors would be fine.

For accurate results it is important that the impedance seen looking ‘each way’ from the crystal, should be purely resistive. To the left, this is ensured by the 10dB pad, the 50Ω splitter and the 50Ω output.
level as the frequency was tuned, by means of the core of $L_1$ over the range 9.5-11.5MHz.

Note the apparent similarity in Figure 3 between a) and d), even though they are in fact very different. In a) the fall in gain with increasing amplitude is due to the device cutting off and probably also bottoming – it is physically incapable of supplying as much gain at the larger amplitude. In d) on the other hand, the gain falls only because the level control loop causes it to: the device could provide more gain at that amplitude if asked to. Thus the characteristic crosses the unity gain line at a steep angle as in a), resulting in a closely defined amplitude. And yet the amplifier is working linearly, more so even than the device illustrated in c), which exhibits very poor amplitude control.

The circuit of Fig. 6 was designed to supply a 50Ω output load directly, and this is connected across $C_{18}$. The reactance of this capac-

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Fig. 8. The spectrum of the output of the linear rf oscillator using the CLC5523. Reference level (top of screen) 0dBm, 10dB/division vertical, Span 0 - 100MHz, IF bandwidth 100kHz, video (post detector) filter max, analyser front-end attenuator set to 20dB.

Fig. 9. The spectrum of the output of the oscillator using the CLC5523. Reference level (top of screen) 0dBm, 10dB/division vertical, centre frequency 10MHz, 500kHz/division horizontal, IF bandwidth 10kHz, video (post detector) filter max, analyser front-end attenuator set to 20dB.

Fig. 10. The spectrum of the output of the linear rf oscillator using the CLC5523, with C22 in circuit in place of R21, C21. Reference level (top of screen) 0dBm, other settings as for Fig. 8 except analyser front-end attenuator set to 30dB.

Fig. 11. Lower trace: spectrum of the output of the phase noise discriminator with its input grounded. Upper trace: spectrum with the output of the CLC5523 oscillator connected as shown in Figs 6 and 7. Both cover 0-5kHz, reference level -60dB, 10dB/division vertical, resolution bandwidth 30Hz, smoothing maximum, 100Hz/division sweep speed.

How about the phase noise?

Clearly, as an oscillator with low harmonic content, the circuit is a great success. But of course, the whole raison d'être of the work was to seek an oscillator with reduced close-in phase noise. This was measured using the phase noise discriminator shown in Fig. 7. The circuit used was similar to the earlier
version, but with the addition of an active smoothing circuit, $R_{12}, C_{14}$ and $T_{r2}$. The intention was to reduce the mains related harmonics which can be seen on phase noise measurements taken with the circuit. However, the improvement proved small.

On investigation, I found that although the mains related spectral lines are 100Hz apart, they are not harmonics of power supply full-wave rectifier ripple, but odd harmonics of 50Hz. They are therefore doubtless due to the leakage flux of mains transformers. The phase noise discriminator was surrounded on the lab bench by various instruments - spectrum analysers, oscilloscope, frequency counter and power supplies.

I recorded a run covering 0-5kHz with the oscillator connected to the phase noise discriminator. Both were powered up, but with the $C_{16,17}$ input of the phase noise discriminator connected to ground.

This set-up is shown as the lower trace in Fig. 11, which represents the lower limit of phase noise measurement capability. As such, it is a more meaningful measurement than the lower trace in Fig. 4(a) of the previous article. That showed the level with the supplies switched off. It thus represented the noise floor of the HP3580A low-frequency spectrum analyser working from a 4.71uS1 resistive source, representing a noise figure of about 5dB.

The oscillator was then connected to the phase noise discriminator as indicated in Figs 6 and 7 - with $R_{21}, C_{21}$ in circuit and $C_{22}$ disconnected - and another run recorded. Both traces are shown in Fig. 11.

**Did it work?**

In a word, no - at least, not as well as hoped. Compare the linear oscillator performance, upper trace in Fig. 11, with that of the LC oscillator covered in the earlier article, reproduced here as Fig. 2(b). You can see that at 2.5kHz and upwards the former is as noisy as the bottoming LC oscillator, and worse than the non-bottoming version.

At low frequencies however, it does not show such a marked rise in noise. Indeed, below 500Hz, it is better than the bottoming LC oscillator and comparable with the non-bottoming version. This gives me hope that with some further development, an improved performance may be achieved.

A difficulty arising from using the CLC5523 in this application, is that it has a low impedance output. This is quoted as typically below 30mS2 at dc and still not much more than an ohm at 10MHz. In the circuit of Fig. 6, this has been padded up to 3351. But via $L_2$ this still reflects damping across the tuned circuit, resulting in an estimated working Q of only 10.

The next task clearly is to achieve a tank circuit working Q of close to the unloaded Q of the inductor, resulting - hopefully - in an oscillator with very low close-in sideband noise. Watch this space.

**References**

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Design wideband antennas

Richard Formato's guidelines for helping you optimise your impedance loaded wideband antennas include a design example that gives continuous coverage from 6 to 150MHz without resorting to tuning and matching.

Adding impedance loading - resistance and reactance - to an antenna is one of the most effective ways to increase its bandwidth. Since the early 1980s, this approach has received progressively more attention, and today it is state of the art for wideband systems.

Choosing a suitable loading profile can maintain high radiation efficiency while providing a remarkable increase in bandwidth. One of the dipole designs in this article, for example, provides continuous coverage from 6 to 150MHz with no tuner or matching network.

The question is, of course, how to select an optimum loading profile. Selecting the wrong profile can produce dismal results. Indiscriminately adding resistance to an antenna obviously causes performance to deteriorate, and the reduction can be severe.

This article analyses several key design parameters for selecting wire antenna loading profiles. Design guidelines are then developed to achieve the greatest antenna bandwidth.

Some of the results are rather surprising, but they can lead to substantial performance improvements for loaded antennas.

Resistive loading

The simplest loading scheme is to insert a resistor in the antenna. In the early 1960s, Altshuler\(^1\) built a centre-fed dipole, or cfd, whose input impedance was, for practical purposes, flat over a 2:1 frequency range as a result of added resistance. The antenna was loaded with a single 240Ω resistor in each arm a quarter-wavelength from the end.

Because resistance reduces the radiated power, the dipole's bandwidth was increased at the expense of radiation efficiency, which was reduced by about 50%. This is the inevitable trade-off in designing impedance-loaded wire antennas.

Bandwidth is best in heavily loaded antennas, but the resulting penalty in radiation efficiency may be too high to provide acceptable power gain - i.e. the product of directive gain and efficiency.

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\(\text{Fig. 1. Elements of the centre-fed dipole, comprising two wire radiating elements with half-length } h \text{ and radius } a.\)

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October 1997 ELECTRONICS WORLD
Narrow rings were burned away using an argon laser to separate the conductive segments, thereby providing the reactive component of the loading profile. Although the loaded centre-fed dipole was useful as a field probe, it was not useful as a transmitting antenna because it was so inefficient. Its transfer function was typically below -22dB.

Attempts have been made to increase the efficiency of impedance-loaded antennas. Rama Rao and Debroux 2,7 for example, developed a more efficient hf monopole antenna by using a 'fractional' Wu-King profile. Efficiencies of 15-36% with a standing-wave ratio of two or less were achieved from 5-30MHz in a 35ft-high including power gain and pattern. But these are important measures of antenna performance, including power gain and pattern. But these are important measures of antenna performance, and number of antenna segments, i.e. the number of discrete resistors. Certain parameters are more important than others – in the sense of having a relatively greater impact on performance – and for some parameters the results are unexpected.

These four design parameters are discussed below for typical hf/lf centre-fed dipole antenna designs. Radiation efficiency and standing-wave ratio, or swr, are examined for rf sources between 2 and 150MHz – the upper limit of the computer model. There are, of course, other important antenna parameters, such as power gain and pattern. But these are not examined in detail because they are usually acceptable in an impedance-loaded antenna with half-length $h$ and radius $a$. The total dipole length is $L=2h$, and its diameter is $D=2a$.

Amplitude of the current profile is plotted schematically along one element's length. Maximum current occurs at the rf source at the feed point, and it decreases along each arm until reaching zero at the end.

The centre-fed dipole's bandwidth is increased by symmetrically loading it with an internal impedance profile, i.e. resistance and reactance. The profile is given by:

$$Z'(z) = R'(z) + jX'(z),$$

where $Z'$ is the complex internal impedance per unit length, in $\Omega$-m, consisting of lineal resistance $R'$ and reactance $X'$, and where $j = \sqrt{-1}$.

The resistance and reactance per unit length for the improved power-law impedance profile are given by:

$$R'(z) = 60(z - |z|)^{-2} \left\{ -1 + \frac{(1 - \nu) \psi'}{2k(h - |z|)} \right\} \left(1 - \frac{\mu R}{k} \right)$$

$$X'(z) = 60(z - |z|)^{-2} \left\{ -1 + \frac{(1 - \nu) \psi'}{2k(h - |z|)} \right\} \left(1 - \frac{\mu X}{k} \right)$$

where $\nu$ is the wave number and $\lambda_0$ is the free-space wavelength corresponding to the design frequency, which is designated $f_0$, via the power-law exponent ('profile exponent') for a travelling-wave current distribution that minimises resonance effects.

The first equation recovers the 100% Wu-King profile when $\nu = 1$. The general case corresponds to $\nu < 1$. The derivation of equation 1 and its relationship to previous work are discussed in Formato 6.

The reactance computed from equation (1b) can be positive, i.e. inductive, or negative, i.e. capacitive. The lineal inductance in H/m and lineal capacitance in F/m are given by $L = \psi'(z)$, and $C = \psi''(z) / \lambda_0^2$. For $X' > 0$ and $X' < 0$, respectively, where the design frequency $f_0$ is in hertz.

In equation 1, $\psi = \psi(f_0)/\psi(1)$ is a complex quantity known as the expansion parameter, 1,2 its real and imaginary parts being subscripted $R$ and $I$, respectively. Symbol $\psi$ represents the ratio of the centre-fed dipole's vector potential to current, which is approximately constant along its length. The expansion parameter is defined in the panel headed 'Expansion parameter equations'.

Because $\psi$ is frequency dependent, it is usually evaluated at the antenna's fundamental half-wave resonance, that is, $\lambda_0 = \lambda_0$ (see Wu and King 5 for details). As discussed below, however, this choice does not necessarily provide the best antenna performance. The design frequency $f_0$ in hertz, and the wavelength $\lambda_0$ (in metres) are related by $f_0 \lambda_0 = c$, where $c = 2.998 \times 10^8 \text{m/s}$ is the free-space velocity of light.

The improved loading profile specified by equation 1 in general contains both resistance and reactance. But adding reactance to a wire antenna – especially capacitive reactance – can complicate construction. As a consequence, many practical designs

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Fig. 1. Demonstration of how radiation efficiency varies with different values of $v$ for the centre-fed dipole example.

A few years after Altshuler's work, Wu and King 5 published a theoretical model of the loaded centre-fed dipole. Unlike the discrete resistor approach, the Wu and King loading profile varied continuously along the antenna and could be implemented with a conductive surface layer of different materials – aluminium and carbon, for example – of varying thickness. Profiles based on the Wu-King theory, which requires a travelling-wave current mode with a power-law amplitude decay, instead of the linear decay required by the Wu-King profile. The derivation of the improved profile and its relationship to the Wu-King profile have been developed in the March 1997 issue 6.

The nonlinear amplitude decrease of the current along the antenna results in a higher average antenna current. In turn, this increases the radiated fields and total radiated power. The antenna's radiation efficiency is higher because it radiates more of the input power.

Several parameters influence how well a particular loading profile performs. There is no one 'best' profile. Important design parameters include: value of the power law exponent; design frequency, wave length-to-diameter ratio; and number of antenna segments, i.e. the number of discrete resistors. Certain parameters are more important than others – in the sense of having a relatively greater impact on performance – and for some parameters the results are unexpected.

The improved loading profile specified by equation 1 in general contains both resistance and reactance. But adding reactance to a wire antenna – especially capacitive reactance – can complicate construction. As a consequence, many practical designs

Expansion parameter equations

$$\psi = \sin h^{-1} \left( \frac{h}{a} \right) - C(2k, a, 2kh) - jS(2k, a, 2kh) + \frac{j}{k} \left[ 1 - \exp(-j2kh) \right]$$

$$C(b, x) = \int_0^1 b \cos W \, du$$

where $W = (a^2 + b^2)^{1/2}$

Table 1. Resistance-only profile for the centre-fed dipole example.

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Electronics World October 1997
employ only resistive loading, because excellent results are often achieved even without the loading profile's reactive component (see Rama Rao equation (1a)). This approach provides a piecewise linear, or step, approximation to the continuous loading profile. Because of these advantages, only discrete loading profiles are considered in this article.

A discrete profile may be determined by first dividing the centre-fed dipole into an odd number of equal length segments, N. The centre segment, which contains the rf source, is not loaded. All other segments are loaded with a lumped resistance placed at the segment centre. The value of the resistor is computed as the product of the segment length – in metres – and the value of the continuous loading profile evaluated at the segment centre, \( R \) in \( \Omega \) from equation (1a). This approach provides a piecewise linear, or step, approximation to the continuous loading profile. There is, of course, any number of other discrete approximations, but only this uniform step approximation is considered here.

As an example of a typical discrete profile, consider a centre-fed dipole with the following design parameters:

- \( L = 22 \) m
- \( D = 10 \) cm
- \( f_0 = 7 \) MHz
- \( v = 0.4 \)
- \( N = 29 \)
- \( \varphi = 8.822 - j2.464 \).

There is no reactive loading or feed-point loading, and the full 100% profile is used. The discrete resistance-only profile, as described above for \( v = 0 \), appears in Table 1. Distance is measured from the origin, and the loading is symmetrical in each arm of the centre-fed dipole.

The loading resistance increases slowly. It rises from 3.86Ω on either side of the rf source, ± 0.76m from the centre, to just over 43Ω in the last segment, located ± 0.62m from the source. This very gradual increase in resistance is typical of more efficient loading profiles.

In the sections that follow, several antenna design parameters are investigated by examining the computer-modelled performance of a typical 22m-long centre-fed dipole using discrete resistance-only loading.

### Power law exponent

Probably the most important design parameter in determining radiation efficiency is the value of the profile exponent, \( v \), which determines how quickly the current amplitude decays along the dipole.

Slower decay, i.e. lower values in a higher average antenna current. This increases the radiated fields and consequently the efficiency. The improvement in efficiency can be quite dramatic. The trade-off is that decreasing \( v \) increases the peak standing-wave ratio, or swr, and causes it to fluctuate more with frequency.

The influence that \( v \) has on radiation efficiency and swr is illustrated in Figs 2 and 3. These plots are based on computer-modelled data for a 10cm diameter, 22m-long centre-fed dipole with \( N = 29 \) and a 100% resistance-only loading profile computed at a design frequency of 7MHz, which is approximately the fundamental resonance frequency of the rf source frequency from 2 to 150MHz. Calculations were made every 1MHz.

The profile exponent \( v \) has a very significant effect on radiation efficiency, with lower values resulting in higher efficiencies. The curve in Fig. 2 for \( v = 1 \), which corresponds to the 100% Wu-King loading profile, shows that the efficiency increases from about 1% at 2MHz to about 54% at 150MHz.

The variation with frequency is smooth and monotonic. But as \( v \) decreases, the efficiency increases progressively more rapidly, especially at lower frequencies. When \( v = 0.2 \), the efficiency increases from about 14% at 2MHz to more than 65% at 10MHz, a 51% increase in a span of only 8MHz.

Beyond 10MHz, the efficiency fluctuates more or less periodically, with a gradually increasing trend until it reaches a maximum above 75% at 150MHz. For \( v = 0.05 \), the efficiency exhibits a pronounced quasi-periodic fluctuation, but its minimum value is more than 68%, and the maximum is well above 80%.

Figure 3 plots swr parametric in \( v \) for an rf source characteristic impedance of 37Ω. If a different feed system impedance is used, an appropriate broadband transformer would be required. For \( v = 1.0 \), the swr varies smoothly from a maximum of greater than 2:1 at 12MHz to a minimum of about 1.45 near 67MHz. It then increases gradually above 67MHz, with a slight dip near 150MHz. The curves for \( v = 0.8 \) and \( v = 0.6 \) show the same general trend. But, significantly, the swr is generally lower with decreasing \( v \), even though it fluctuates more at lower frequencies.

Figure 3(b) plots swr for \( v = 0.4 \) and \( v = 0.2 \). The swr is generally lower for \( v = 0.4 \) than it is for \( v = 0.6 \), but the variability with frequency is much greater, and the peak values are higher at some frequencies.

For \( v = 0.04 \), the swr exceeds 2:1 between about 12 and 16MHz, but it is below 2 for \( v = 0.6 \). As \( v \) decreases to 0.2 and then to 0.05, Fig. 3(c), the swr fluctuation becomes more pronounced, and the peak values are higher. The minimum swr values, however, are generally lower, and, on the average, the swr is still well below 2:1.

The best choice for \( v \) is evidently the lowest value that provides acceptable swr at frequencies of interest. Choosing \( v \) in this way ensures the highest possible radiation efficiency, and the improvement is usually very substantial.

### Design frequency

The design frequency \( f_0 \) is another important parameter in determining a good loading profile. Although it appears to be accepted practice to choose \( f_0 \) close to the centre-fed dipole half-wave resonance frequency (see Wu and King for example), this choice is not necessarily the best.

Because the expansion parameter, which plays a major role in determining the loading profile, is frequency-dependent, the actual choice of design frequency must be based on how much a given loading profile improves bandwidth while still providing good radiation efficiency.

There is no other sensible scheme for determining \( f_0 \) because there is no theoretical basis for choosing one value over another. The best approach is therefore empirical, which is the
RF DESIGN

greater provides better performance. Even more important.

Because the efficiency increases with decreasing \( v \), the influence that \( f_0 \) has becomes less pronounced at the high frequencies. The less heavily loaded profile (\( v=0.4 \)) is much better - especially at the design frequencies. The less heavily loaded profile computed at a design frequency off_0=7MHz.

In order to investigate the effect of \( \text{LD} \), I computed the radiation efficiency and swr from 2-150MHz for radiating element diameters of 0.1cm, 1cm, and 10cm. A 22m long centre-fed dipole was modelled, yielding \( \text{LD} \) ratios of 220, 2200, and 2200, respectively, which represent antennas ranging from 'extremely thin' to 'thin'. Profile exponents of \( v=1.0 \) and 0.4 were used with 29 segments and a 100% resistance-only loading profile computed at a design frequency of \( f_0=7MHz \).

The larger diameter centre-fed dipoles have better radiation efficiency at all frequencies for both values of \( v \). The improvement in efficiency becomes progressively greater at higher frequencies, and it approaches a factor of two at very high frequencies. Its swr is below 2.5 at all frequencies above 6MHz.

Thus, even though the 22m centre-fed dipole has a fundamental resonance near 7MHz, choosing a design frequency that is ten times greater provides better performance. For example, as the curve for \( v=0.4 \) shows, the radiation efficiency at 10MHz is about 45% when \( f_0=7MHz \), but it increases to 67% when \( f_0 \) is increased to 70MHz. Choosing a higher design frequency thus results in a much better antenna.

The advantage of a higher design frequency is also evident in the swr plots of Fig. 5. It is quite significant that selecting \( f_0=70MHz \) when \( v=1.0 \), Fig. 5(a), results in the lowest swr across the entire 2 to 150MHz band. When \( v=0.4 \), Fig. 5(b), choosing \( f_0=70MHz \) results in swr \( \leq 2 \) across most of the band. The variability is greater, and the swr is not consistently lower with increasing \( f_0 \) as it is when \( v=1.0 \). These effects are minor, however, and better overall performance usually results from higher values of \( f_0 \).

Radiator length-to-diameter ratio

Increasing the element diameter is a standard broadbanding technique for wire radiators. It is therefore not surprising that a larger diameter, impedance-loaded centre-fed dipole exhibits better overall performance than its thin counterpart.

In order to investigate the effect of \( \text{LD} \), I computed the radiation efficiency and swr from 2-150MHz for radiating element diameters of 0.1cm, 1cm, and 10cm. A 22m long centre-fed dipole was modelled, yielding \( \text{LD} \) ratios of 220, 2200, and 2200, respectively, which represent antennas ranging from 'extremely thin' to 'thin'. Profile exponents of \( v=1.0 \) and 0.4 were used with 29 segments and a 100% resistance-only loading profile computed at a design frequency of \( f_0=7MHz \).

The larger diameter centre-fed dipoles have better radiation efficiency at all frequencies for both values of \( v \). The improvement in efficiency becomes progressively greater at higher frequencies, and it approaches a factor of two at the high end of the band. For the profiles with \( v=0.4 \), increasing the element diameter also reduces fluctuations in the radiation efficiency, but this effect is not evident in the heavily loaded profiles when \( v=1.0 \).

The largest diameter element provides the best standing-wave ratio performance, especially at lower frequencies. Its swr is below 2.5 at all frequencies above 6.5MHz, and below two above approximately 12MHz. The swr decreases quickly up to about 30MHz and flattens out at less than 1.5 for most of the rest of the band.

By contrast, the standing-wave ratio for the very thin element with \( D=0.1cm \), is high, being above two throughout the band, and above 2.5 below 30MHz. The very thin radiator thus fails to provide acceptable standing-wave ratio even though it is very heavily loaded.

Similar standing-wave ratio behavior is evident when \( v=0.4 \). The fattest element provides the best performance. Its standing-wave ratios below 1.5 over most of the band, and below 2.15 at all frequencies above 6MHz.

Decreasing the diameter to 0.1cm increases the standing-wave ratio, but not as much as it did for the more heavily loaded profile with \( v=1.0 \). When \( v=0.4 \), however, the standing-wave ratio variability becomes much more pronounced for smaller element diameters.

Building an antenna with a low \( \text{LD} \) ratio, that is, making it 'fatter', may be difficult if too large a diameter conductor is required. A continuous cylindrical surface can sometimes be approximated by a sufficient number of parallel wires uniformly spaced around the cylinder's circumference. A large number of wires may be required, however, depending on how good the approximation must be.

Such a dipole structure, sometimes called a 'cage dipole' because of its resemblance to a bird cage, can offer a convenient and effective alternative to large diameter cylinders for low \( \text{LD} \) designs.

Segmentation

The segmentation used, that is, the value of \( N \), also influences how well the loaded centre-fed dipole performs. I studied its effect by computing radiation efficiency and standing-wave ratio for three segmentations. The results were somewhat unexpected. A 22m long, 1cm diameter centre-fed dipole was modelled with \( N=29 \), and 119 segments. A 100% resistance-only loading profile with \( v=1.0 \) and 0.4 was computed at \( f_0=7MHz \).

The efficiency data show that the least segmented antenna, \( N=29 \), provides the best overall performance. As segmentation increases, radiation efficiency decreases, although the change is not great from \( N=29 \) to 119. This result is somewhat surprising, since increasing segmentation presumably provides a better approximation to the continuous loading profile. However, the data show quite convincingly that the net effect of adding more discrete resistance is to increase the \( R^2 \) - Joule heating losses more than the radiated power, resulting in lowered efficiency.

This effect occurs for both values of the profile exponent. As is typically the case, the efficiency fluctuates more with frequency as \( v \) decreases, and the variability is greatest at the low end of the band. One effect of increasing \( N \) is to reduce the fluctuation somewhat, but the change is not pronounced, and it occurs only when \( v=0.4 \).

The standing-wave ratio data are not as clear cut, but the general conclusion is still that the lowest segmentation probably provides the best overall performance. For \( v=1.0 \), the ratio is lowest for \( N=29 \) at frequencies below approximately 90MHz - about a factor of 12 greater than the fundamental resonance. In the same frequency range, it is not significantly different from the \( N=59 \) or 119 values even when \( v=0.4 \).
Above 90MHz, the antenna with N=59 performs best for both values of v, but the 119-segment design is very close. Nevertheless, in that same frequency range, 'eyeball average' standing-wave ratios for N=29 are about 1.6 for v=1.0 and 1.65 for v=0.4, which are very good indeed.

Thus, the least segmented antenna provides very robust swr performance at all frequencies from 2-150 MHz.

In summary
This article has investigated four design parameters for impedance-loaded wideband wire antennas. Using the following guidelines to select these parameters when designing an impedance loading profile should provide a near-optimum wideband antenna.

First, the profile exponent v is very important in determining radiation efficiency. Lower values of v result in higher efficiencies, and the improvement is very significant. The optimum value for v is the smallest value that provides acceptable swr at frequencies of interest.

Secondly, because the design frequency f0 is usually chosen close to the fundamental centred dipole resonance, it is somewhat surprising that higher frequencies generally result in much better radiation efficiency, especially for loading profiles with smaller values of v. Higher values of f0 also usually give better overall swr performance. The optimum value for f0 is the highest value that provides acceptable swr at frequencies of interest.

Thirdly, the lower the L/D ratio, the better.

Large diameter radiating elements provide much more bandwidth than thin ones, even without impedance loading. A large diameter radiator makes it easier for a loading profile to provide the greatest possible bandwidth. If necessary, large diameter conductors can be approximated by multiple parallel wires.

Finally, reducing the segmentation, that is, the number of discrete resistors used to approximate the theoretical continuous loading profile, results in slightly better radiation efficiency. The swr is not particularly sensitive to segmentation at 'low' frequencies, and it is slightly better with increased segmentation at 'high' frequencies.

For a thin antenna, 'low' frequencies are less and 'high' frequencies are greater. The design guideline for segmentation is to use the smallest number of discrete resistors that meets the swr objectives.

Software on disk
Richard's PC software relating to this article, for computing loading profiles, is available on a 3.5in disk. Send a postal order or cheque for £10, payable to Reed Business Information Group, to Wideband Disk, Electronics World Editorial Offices, Quadrant House, The Quadrant, Sutton, Surrey SM2 5AS.

Please don't forget to mention your name and address - some do!

References
4. Rama Rao, B., and Debroux, P. S., Wideband HF Monopole Antennas with Tapered Resistivity Loading, IEEE Military Communications Conference, Monterey, California, 30 September to 3 October 1990 (MILCOM '90)
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ELECTRONICS WORLD October 1997
Audio expert Douglas Self reveals why he uses Woolworths' mains cable to drive his speakers in this, the second of three articles on the interface between the power amplifier and loudspeaker.

Loudspeaker cables are relatively simple things - notwithstanding the haze of controversy that often surrounds them. In the drive to adopt ever more complex pulse-testing techniques, most of them of very questionable relevance, the basic cable properties appear to have been overlooked. This is emphasised by the striking fact that the proponents of unlike-cable-hypotheses rarely even mention fundamental parameters like the resistance and inductance of their test specimens.

I looked at the first-order effects first, and only then examined the details. This approach at least has the merit of novelty. Firstly, it is beyond doubt that an audio cable is not a transmission line. Transmission line effects can have nothing to do with audio, unless you're in the telephone business.

In real life work, such effects are usually considered negligible until the frequency is high enough for the length of the line to reach one-sixth of a wavelength. The wavelength of 20kHz is 15km, so unless your loudspeakers are over a mile away from the amplifier, there is no need to fret.

A length of cable has series resistance, series inductance, and shunt capacitance as its major parameters. A simple cable model is shown in Fig. 1 and in Fig. 2a), where the series resistance and inductance, and the shunt capacitance, are represented as lumped components. The capacitance is shown concentrated after the resistance and inductance, where it will cause the most response variation; this is actually a very slight rise rather than a roll-off. This is due to interaction with the inductance.

In practice the effect of this capacitance is tiny compared with the roll-off due to the series inductance. A typical response change at 20kHz is a fraction of a thousandth of a decibel, so the effect is negligible anyway. The amplifier is assumed to have zero output impedance unless otherwise stated.

It seemed worthwhile to confirm that this position for lumped capacitance really is the worst case, and this was done by simulating the cable models in Fig. 2. A first approximation to a distributed model is Fig. 2b), where the cable is split into two halves. A slightly more accurate version is Fig. 2c), with the cable now divided into four. Simulation confirms that Fig. 2a is the worst case for capacitive high-frequency lift; but the effect is tiny in all three cases. Further thoughts on cable capacitance are to be found in a separate section below.

Cable resistance effects

The first basic parameter is cable resistance, determined by the conductor material and its total cross-sectional area. Copper is the obvious material, though silver can be used if you are rich and gullible enough. Silver is the most conductive metal but despite its cost has only 7% less resistivity than copper. It also presents some awkward problems with non-conductive tarnish films, namely silver sulphide.

If you want to reduce your cable resistance slightly, this is not an economical way to do it, compared with a minor increase in copper cross-section. I have seen mercury cables suggested, in the form of quicksilver-filled hosepipes. In view of the insidiously poisonous nature of mercury vapour, one can only hope this wasn’t taken seriously by anyone.

Having rejected silver as ridiculously expensive, and settled for copper, it is then just a matter of using enough area to get a suitably low resistance per metre. The main consequence of unwanted series resistance is frequency response perturbations due to the
speaker impedance varying with frequency.

The second basic parameter is cable inductance, determined both by the conductor spacing and their diameters. The equation for the inductance of parallel spaced cylindrical conductors is:

\[ L = 4 \times 10^{-7} \cdot \left( 1 + 4 \log \left( \frac{D}{R_1 \times R_2} \right) \right) \]  

in henries per metre. Variable \( D \) is the spacing between centres of two cylindrical conductors while \( R_1 \) and \( R_2 \) are the radii of the conductors.\(^3\)

The main consequence of this series inductance is a high-frequency roll-off with an 8Ω load. Cables that deliberately space the two conductors increase the inductance considerably, and this is not a good idea.

Fortunately, inductance increases slowly with spacing. There is the theoretical possibility of some sort of special interaction between cable inductance and loudspeaker behaviour, but the practical inductions are so small - at least two orders of magnitude below crossover inductor values - that this does not seem likely to be a problem in practice.

**Cable investigations**

As usual in audio, if you want some hard facts you have to get them yourself. As a first step, I took myself to the local DIY superstore to find out what was on offer to those who looked no further for their speaker cables - ie 99% of the population.

The data I gathered is shown in Cable Table 2.\(^4\) The main sample was 6A two-core, and it really was cut from a defunct lawnmower. This has a basic go-and-return (ie total loop) resistance of 178mΩ and an inductance of 2.3μH. If we use the amplifier/output-inductor combination examined in the first part of this article,\(^1\) with a total dc resistance of 24mΩ, the extra cable resistance degrades the so-called damping factor from 330 to 40. Note that all cable parameters quoted are for the total go-and-return path.

This gives a flat attenuation of 0.20dB, due to the cable resistance forming a potential divider with the 8Ω load, and a main hf roll-off due to the cable inductance forming an LR low-pass filter with the 8Ω load, giving a further loss of -3dB at 398kHz. This corresponds to -0.01dB at 20kHz, which is negligible.

Add an amplifier output inductor of 6μH to a cable inductance of 3.3μH and the roll-off becomes -3dB at 145kHz, and -0.086dB at 20kHz. This demonstrates that even a maximum-value output inductor combined with twin-core cable does not pose a problem with an 8Ω resistive load. However, increasing either inductance further might be unwise.

**What to do with the spare conductor**

The second lawnmower cable was three-core. I suggest that the thing to do with the spare conductor is to parallel one of the other two; this obviously reduces the total resistance to three-quarters of the original value, and a bit less obviously - to me anyway - also reduces the total inductance. I always parallel the ground return rather than the hot conductor, on the philosophical basis that earths should be more solid than signal paths. Here, of course it makes little difference.

The reduction in inductance is explained by the geometry of the situation in Fig. 3, suggested to me by George Chadwick.\(^5\) If the normal go and return conductors are A and B, then when third conductor C is connected, the return current divides in two. This is equivalent to the original current in a notional conductor at a point D halfway between the B and C.

Distance A-D is 3√2, or 0.97, of A-B, and so the effective conductor spacing is reduced. Putting this into equation 1 shows that inductance is reduced by 1.4 times. Measured reductions in inductance as in Cable Table 2 are 1.32 times, which is reasonable considering that the physical dimensions are subject to tolerances.

Using two conductors only the resistance is 189mΩ and the inductance 3.9μH, and the loss and roll-off much as the two-core sample. Connecting the third conductor reduces the round-trip resistance to 142mΩ and the inductance to 2.5μH. The damping factor improves from 40 to 50.

The total shunt capacitance between go and return conductors is 488pF for the two-wire connection, and 655pF for the three-wire case. With three conductors, the frequency-insensitive loss is reduced to 0.153dB, and the inductive roll-off is now -3dB at 520kHz, which is a mere 0.007dB at 20kHz.

The third sample was 13-Amp three-core cable, as used for connecting electric fires. It is an interesting economic point that 13A cable only has 1.5 times the copper cross-sectional area of 6A, though a factor of 2.2 would be needed for the same current density; presumably a slightly higher voltage-drop per metre is accepted.

Using only two cores of 13A cable, five metres has a resistance of 132mΩ and inductance of 3.3μH. Moving up to the larger cable has reduced resistance exactly as predicted from the increased cross-sectional area, but inductance is unchanged. The flat resistive loss is reduced to 0.14dB, and the roll-off stays at -3dB at 20kHz.

The improvement over 6A cable with three cores used is minimal. Using three cores of...
13A, resistance falls to 98mΩ, the lowest so far, and inductance is 2.5μH; once more unchanged from the 6A cables.

Cook cable
The fourth sample, and the next step in cable sophistication in my view, is the stuff you wire cookers up with. Cooker cable has a flat format with two heavy-gauge conductors and a much lighter earth connection between them. This is usually called 'twin-and-earth'. For 5m, the go-and-return resistance is 21mΩ, which is an order of magnitude lower than the lawnmower cable. The inductance is higher at 4.1μH, due to the greater spacing of the main conductors. Connecting the central earth conductor reduces total resistance slightly to 19mΩ and inductance usefully to 3.1μH. With this rather serious cable, the frequency-insensitive resistance loss is reduced to 0.02dB, and the inductive roll-off not greatly altered.

Finally, I examined 50Ω RG58 coaxial cable, which has been recommended for speaker cabling. Coaxial cable is not made with anything like the weight of copper in cooker cable, so its resistance is relatively high at 158mΩ. The 5m inductance at 3.3μH is less than half that of the ordinary cables used two-core style, though three-core usage closes some of the gap. The point is that the effects of cable inductance appear to be an order of magnitude less important than cable resistance, so the lower inductance of coaxial cable is not the deciding factor.

Referring to Cable Table 2, the loss and roll-off data given here applies only to 8Ω resistive loads. This is unrealistic, but gives a feel for the size of the problems. It would be desirable to produce similar results for a real speaker load, but they come in almost infinite variety — which one would you use?

Amplifier stability
There is anecdotal evidence that when it became fashionable to use allegedly better-sounding cables, several makes of 'high-end' amplifier showed hf instability. This has been ascribed to the greater shunt capacitance of some of the peculiarly constructed cables. We will probably never know the truth of the matter — much as we will probably never know if there really was a 'transistor sound' in the sixties, allegedly due to crossover distortion. But the shunt capacitance of even a very strange cable is so much smaller than the values that cause trouble, usually 100nF and up,
that it is very unlikely to affect the stability of an amplifier.

A much more likely explanation is that all the amplifiers involved lacked output inductors, but the old cables had enough series resistance to isolate the amplifier from the capacitive components of the loudspeaker impedance. The new cables — being in general much thicker — did not.

Omitting the output inductor is not a good idea; so long as it is included, instability from any capacitance should be impossible.

**Skin effect**

An unfortunate complication to the study of audio cables is that their ac resistance is greater than that at dc. This is not due to cable inductance. What we are talking about here is real resistance, not just an increase in impedance magnitude due to additional inductance.

This increase in ac resistance is due to skin effect. The ac current in a conductor is not uniform across its cross-section, but concentrates in the outer part of it. This is because for an elementary filament carrying current, the impedance at the conductor centre is greater than that at the surface, so the current is diverted towards the surface. You will have to take this on trust unless you want to explore some serious vector calculus.

The result is that the ac current distribution across a conductor dies away exponentially with distance below the surface. At some skin depth \( d \) the current has fallen to 1/e of its maximum, and the situation is equivalent to a uniform current distribution in a thin outer shell of the same thickness \( d \). Skin depth \( d \) may be determined from,

\[
d = \frac{1}{\sqrt{\pi \mu \sigma}}
\]

where \( \mu \) is permeability and \( \sigma \) is resistivity.

Having found \( d \) from this well-established equation, it is simple to calculate the area of the annulus representing the equivalent conducting part of a circular cross-section cable, and thus find the increase in resistance, Fig. 4. Note that this is a true frequency-dependent resistance — not an inductance — and does not have the phase shift that would accompany extra inductance.

Taking 13A copper mains cable as an example, equation 2 shows that below 13kHz, the skin depth is greater than the cable radius, and there is no significant effect. At 20kHz, the skin depth is 0.47mm, and the conductor radius 0.6mm. The cross-sectional area is 1.08mm², of which only the central 0.06mm² is unused. Therefore resistance per metre only increases from 152mΩ to 160mΩ, and this can have no appreciable consequences.

Some writers appear to believe that the presence of skin effect must mean that the conductor is acting as a transmission line; this is quite untrue.

Cable cross-sections other than circular give different results, but since the circular version has the minimum perimeter for a given area, all other shapes give more surface, and so less skin effect. This is why high-power rf stages often use flat copper strap connections.

Skin effect is relatively more significant for low-resistance cables, as current can move laterally more easily. For loudspeakers however, the real criterion is the total absolute resistance, which is always reduced by using larger cable. The conclusion must be that even in the worst case there will be no problem, and there is no justification at all for Litz-type cables.

Skin effect is more significant in other technologies; it is vital at rf because the skin-depth is so small, and hence silver-plating is well worthwhile. It is also important in 50Hz power transmission, because of the low impedances involved, and in some circumstances multiple parallel busbars are required to control the extra resistive heating.

**Proximity effect**

One further aspect of cable theory is the proximity effect. This results from the interaction between the adjacent go and return currents. Proximity effect is similar to skin effect, but in this case the interacting currents are moving in opposite directions, causing the current distributions in each conductor to crowd towards each other. As with skin effect, this reduces the effective cross-sectional area of the conductor and increases its resistance. Again, the effect is an increase in apparent resistance as frequency increases, and not an increase in inductance.

The only reference that I have found to this deals rather unconcerningly with resistance in ohms per thousand feet. The ratio between ac and dc resistance varies as,

\[
\text{proximity factor} = \frac{f}{\sqrt{R_{dc}}}
\]

where \( R_{dc} \) is the dc resistance of 1000ft, which for 5/0.020in² copper is 10.85Ω.

At 20kHz, the proximity factor is therefore 42.9, which yields from an ac resistance about 7% greater than the dc value. This is of the same order of magnitude as skin effect, and should be equally negligible.

Once again, the effect is relatively more significant for low-resistance cables, as lateral current mobility is greater, and once more the greatest problems arise in AC power transmission.

**Biwiring**

In recent years, a loudspeaker cabling practice known as "biwiring" has become fashionable. There is no evidence that it does any good, but naturally this has not impeded its popularity.

Assuming we have a two-way speaker with the hf and If sections of the crossover separately accessible, biwiring uses two separate cables from the same amplifier output to drive two separate crossover inputs, as shown in Fig. 5.

The conceptual origin of this practice is disclosed by the similarity of the word to "bi-amping" where separate amplifiers would be used.
for the high and low-frequency sections. Bi-amping has a rationale in terms of reduced amplifier intermodulation, the possibility of sophisticated crossovers, and so on, but none of this applies to solid copper, so biwiring seems pointless.

It is conceivable that a loudspeaker could exist which interacted in some undesirable way with cable inductance, so that separate induc-
tances in separate cables eased the problem. However, the low inductance values shown by normal cables seem to make this very unlikely.

**Cable capacitors**

Cable Table 2 shows that the shunt capacitance in normal cables is low — at less than 150pF/m. Simulations were done on the circuit models in Fig. 2, using some worst-case cable parameters of 0.32Ω resistance, 3.3µH inductance, and 655pF shunt capacitance.

If these R, L, C parameters are treated as 'lumped' components, the worst case is Fig. 2a, with all the capacitance lumped after all the inductance. This gives a rise of 0.00028dB at 20kHz. Figure 2b gives 0.00021dB and Fig. 2c 0.00018dB; as the model more closely resembles a distributed system the effect becomes less.

These tiny response deviations appear to demonstrate beyond doubt that cable shunt capacitance is so small its effects are completely irrelevant. Therefore all the speculation about second-order capacitance effects — such as dielectric absorption — must be completely pointless.

Figure 6 shows the resistive loss of 0.34dB and inductive roll-off resulting from the above cable parameters and an Ω2 load. Both zero and 655pF capacitance versions are plotted. You can see that the two curves are virtually identical.

**Crystal structure**

There has been much debate about the crystal structure of copper, and whether this or its oxygen content can affect anything. Most of it has shown what can only be called an abysmal ignorance of the most basic physics and metallurgy.

Crystal boundaries in copper cannot and do not act as mysterious diodes, and the simplest of experiments proves this beyond doubt. Much of the suspicion directed towards the oxides of copper probably stems from faint memories of copper-oxide rectifiers; their operation relied on a carefully-grown cuprous oxide film in pressure-contact with a lead washer, and this has no relevance to the bulk copper in cables.

**In summary**

All the effects described here are real, though unlikely to ever be audible under even the most critical conditions. The use of a resistive load only is obviously open to objections, which will be addressed in a forthcoming piece on speaker impedance curves and their implications.

It emerges that even a bit of cable can be complicated if examined in detail. For more details on some of the issues raised here, see reference 11. There is, however, no sign of anything mysterious, incomprehensible, or indeed audible. If you examine a real first-order effect, such as response variation due to cable capacitance, and find it far below any possibility of perception, then speculative second-order effects such as dielectric absorption in this capacitance can surely be dismissed.

There is no justification for treating cables as transmission lines, and no need for special cables that cost insulting sums per metre. Silver appears to have no possible justification for its use. The only factors that count are series resist-
ance and inductance. The first can only be minimised by heavy-gauge cables. Inductance offers more possibilities for its reduction; keep the go and return conductors at minimal spacing, use multiple conductors, or resort to a coaxial structure.

From the considerations given here, the best speaker cable would seem to be 13A cable with three cores used. Significant reduction in resistance requires cooker cable, which is bigger and much less flexible, and so harder to hide. If cable inductance is your main concern, then RG58C coaxial cable certainly has less of it, but its resistance is nearly twice that of 13A cable.

Just for the record, my usual speaker connec-
tions are 13A mains cable, with neutral and earth paralleled to reduce resistance and induc
tance, purchased from Woolworths in 1972. Let it never be said that I am not at the leading edge of cable technology.

**References**

4. Mr George Chadwick MIEE, Private communication.

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When designers tackle Wien bridge oscillators, they often overlook phase shift in the amplifier. But simply relying on textbook equations can result in frequency errors up to 70%. Bryan Hart explains the importance of amplifier phase shift and presents a practical way of dealing with it.

The effect of amplifier phase shift on the oscillation frequency of a Wien-bridge oscillator, incorporating an operational amplifier with a single-pole frequency response, is related to the unity-gain frequency of the op-amp by a simple formula. For practical purposes, this formula can be regarded as exact.

In a standard textbook analysis of the Wien-bridge oscillator, which is widely used in the design of commercial signal generators, it is assumed that the amplifier used is ideal in that it exhibits zero phase shift. Consequently the oscillation frequency is shown to be solely dependent upon the passive Wien RC network parameters.

But, all practical amplifiers exhibit a finite, even if sometimes small, frequency dependent phase shift.

In this article, I present a simple expression for the oscillation frequency for the common practical case of an amplifier with a single-pole frequency response. I also discuss the significance of the expression in Wien-bridge oscillator design.

Wien bridge analysis

Figure 1 shows the circuit of a Wien-bridge oscillator with a voltage amplifier, $G_v$, whose voltage gain is $G$. Amplifier block $G$, comprises an op-amp, represented by $A_v$, with open-loop voltage gain $A$, and closed-loop gain-setting resistors $R_A$ and $R_B$.

Amplitude stabilisation is achieved by making $R_B$ a thermistor with a negative temperature coefficient. The Barkhausen criterion for sustained oscillations to occur at the output of $G_v$ is,

$$\frac{GZ_p}{Z_s + Z_p} = 1 < 0$$

In this expression, $GZ_p$ is the total gain from the output of the amplifier to the input of the Wien bridge and $Z_s$ is the feedback network impedance.

Initially, $A_v$ is assumed to have an infinite input impedance. As shown by the linearised Bode gain-plot in Fig. 2, the dc and low-frequency gain is $A_0$, the cut-off frequency is $\omega_c$, and the unity-gain frequency is $\omega_y$, which is approximately $A_0\omega_c$.

Hence,

$$A = \frac{A_0}{1 + \frac{\omega}{\omega_y}}$$

Substituting this value of $A$ in (2) gives,

$$G = \frac{1}{b + j\frac{\omega}{\omega_y}}$$

where,

$$b = b_v + \frac{1}{A_v}$$

For the Wien network,

$$Z_p = \frac{R_p}{1 + j\omega C_p R_p}$$

so, after routine algebraic manipulation, it follows that,
Putting it into practice

In practice, equation (10) may be regarded as exact since the non-zero input capacitance and non-infinite resistance of G can be taken as included in Cp and Rp respectively. Also, A\_pc differs from A\_o by less than 0.005% for A>100.

Typically, A>10000. In the case of op-amps with the input-circuit architecture of the classic 741 op-amp, A\_o is very large – at 100000 – but subject to a large uncertainty. Nevertheless, ω\_o is still closely-defined, being determined by the input stage bias current and a capacitor included to give dominant-lag frequency-compensation\(^2\).

Table 1 shows some spot values of ω\_o/ω\_co calculated from (13, 14). Clearly, for a value of u comparable with unity, the effect of amplifier phase shift on ω\_o can far exceed the effect of component tolerances. From (8c), a ±1% resistor tolerance and a ±2% capacitor tolerance leads to a 3% tolerance on ω\_o, and hence ω\_o.

For operation at around 1MHz, it is not possible to meet the condition ω\_o/ω\_co=1 using available voltage-feedback op-amps because it requires an f\_T of around 100MHz.

Discrete/hybrid component amplifiers, or possibly current-feedback op-amps, are required.

When u is much less than unity b is approximately 0.333. In this case, a binomial expansion of (13) that discards all but the first two terms yields the approximation,

\[ b \approx 0.333 - 4.5 \frac{R_p}{R_s} \omega_c \]  

(15a)

or

\[ \frac{\omega_c}{\omega_o} = 1 - 4.5 \frac{R_p}{R_s} \omega_c \]  

(15b)

Table 1. Spot values, calculated from equations (13) and (14), rounded up to three decimal places.

<table>
<thead>
<tr>
<th>u(=(a/\omega_c))</th>
<th>b</th>
<th>(\omega_o/\omega_c)</th>
</tr>
</thead>
<tbody>
<tr>
<td>0</td>
<td>0.333</td>
<td>1.000</td>
</tr>
<tr>
<td>0.001</td>
<td>0.333</td>
<td>0.996</td>
</tr>
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<td>0.1</td>
<td>0.317</td>
<td>0.717</td>
</tr>
<tr>
<td>0.2</td>
<td>0.288</td>
<td>0.570</td>
</tr>
<tr>
<td>0.5</td>
<td>0.185</td>
<td>0.331</td>
</tr>
</tbody>
</table>

Frequency ω\_o represents the resonant point of the Wien RC network and the frequency of oscillation for an amplifier with zero phase shift.

Using (5) and (7) in (1), the condition for oscillation in the present case is,

\[ b + j \left( \frac{\omega_s}{\omega_o} \right) \times m + jn \left( \frac{\omega_s}{\omega_o} - \frac{\omega_o}{\omega_o} \right) = 1 \]  

(9)

From this equation, it follows that oscillations can be sustained at a frequency ω=ω\_o(ω\_o/ω\_co) given by,

\[ \omega_s = \frac{1}{\sqrt{nb + mu}} \]  

(10)

where

\[ u = \frac{\omega_s}{\omega_o} \]  

(11)

Similarly, (12) becomes,

\[ 3b^2 + h(9u - 1) + 3u(u - 1) = 0 \]  

(14)

References
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CIRCLE NO. 129 ON REPLY CARD

October 1997

839
Better mosfet models, new amplifier designs and improved circuit simulation software are the topics of Cyril Bateman’s Internet column this month. Cyril also discusses a new search engine enhancement that should save you a lot of wasted surfing time.

The rapid growth of industrial and commercial Internet web pages, together with the fashion for large, more complex sites, continues apace. This growth has negated the speed recently gained from using faster modem access. Perhaps more importantly it has much increased the difficulty in locating data.

To counter this, the keyword search engines continue to add both storage and functionality to their systems. The inevitable result is that a search now returns with ever more matches.

Most keyword search engines encourage use of Boolean controlled keywords. Thus a search can become well focused if your chosen control words match those required by the search engine. If unmatched, the search result can be utterly useless.

AltaVista, which always provided Boolean controls, has now added their ‘Live Topics’ facility. Live topics, also called ‘Search Wizard’, can be used with text only or Java enabled browsers. It provides a list of matching keywords to select or de-select for a more refined search.

These live topics words are not pre-defined, but are ‘dynamically categorised’ by the search engine, from analysis of all keywords found in the documents identified in your original search result.

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8 LSI Logic http://www.lsiologic.com
9 Pass Laboratories http://www.passlabs.com
10 High Fidelity Engineering http://www.decware.com/zenamp.htm

Using Live Topics
To illustrate the new search tool, I ran a normal AltaVista Boolean search using two keywords, +circuit +simulator. This search found documents containing 541 occurrences of circuit and 115 958 for simulator, spread over 9000 document pages — far too many for me to digest.

While live topics is withheld for finds less than 200 documents, for this search it
was of course offered me. Immediately under the 9000 document statement, I was given three choices, 'tables' for JavaScript enabled browsers, 'text-only' for any browser or 'help'.

Choosing tables resulted in a JavaScript interface of a box showing my original search words and tables of selectable sub-keywords, grouped under major headings. AltaVista advise selecting only one or two words, then repeat searching.

I chose to include schematic and analog, then exclude vhdl, visi, eda, embedded, chip, microprocessor, microcontroller and emulator. This revised search returned only 100 document pages — much more suited to my needs, Fig. 1.

Had I performed a conventional search, I would not have thought of many of these keywords. I would certainly not have deselected. This dramatically demonstrates the value of the live topics words, and the power of this new technique, which is unique to AltaVista.

Using the latest browser versions, a slightly different Java Interface should be available which downloads an applet to your system. It works in a similar way to that shown, but further improves the versatility of the software by providing a choice between table or graph based lists of the live topics words.

Should all the on-line search techniques described to date fail, help is to hand in Internet's equivalent of your friendly local librarian. A new website, HumanSearch2, was started by seven students of the University of Rhode Island.

Intended to overcome the search problems which result from the rapidly increasing Web document base, HumanSearch offers use of its mostly unpaid volunteer team — of researchers free of charge — to search for any topic on your behalf. One particular specialty search they can perform is lost person tracing.

To use this unique search offering, simply fill in the one page questionnaire. Their team of skilled searchers will work on your request and within a couple of days, e-mail you with the results, Fig. 2.

Simulation and design software

My AltaVista live-topics search revealed a possible answer to Ian Hegglun's mosfet question, also three Spice based circuit simulators not previously mentioned.

In the May issue, Ian asks whether level-three mosfet models and/or Electronics Workbench 5, can accurately model in the crossover region. All Spice simulator-engine core software originates from The University of California at Berkeley. The latest Berkeley Spice is 3F5, used in this Electronics Workbench version.

For historical reasons, mosfet models, together with relevant core simulation software, are less well developed than those for bipolar devices. Berkeley currently has two mosfet simulation projects. A research project to investigate energy balance modelling for accurate current drive prediction of deep-submicron mosfets3.

A second program under the generic heading of BSIM34 is responsible for actual BSIM implementations in Spice. While the 3F5 core has been BSIM compatible for some time, the BSIM3 level 3V3.1 mosfet feature was added only in February this year. As a result, it is necessary to carefully check the exact status of 3F5 based simulators.

Silvaco5 produces the SmartSpice simulator and advanced Spice models. SmartSpice fully supports BSIM3V3.1, has enhanced both the physical model used and its numerical implementation. Thus Silvaco claims leadership in advanced Spice model technology.

The SmartSpice simulation results show their BSIM3V3 model eliminates certain negative capacitances found using
the Berkeley version. The company has also enhanced the impact ionisation current model for both short and long channel devices, Fig. 3.  

TopSpice for Windows® is a 32-bit simulator offering analog/digital/behavioural mixed-mode simulations, together with schematic capture. Its built in logic simulator dramatically reduces mixed-mode simulation times. Unusually, this affordable simulator includes a Smith chart post processor able to handle ‘S’ parameters, Fig. 4.  

Version 2 of B2 Spice® is available both for Windows and Macintosh. This simulator helps overcomes model availability problems by including both model and symbol editors. A 4Mbyte download trial version is available. The Windows version requires that either Access 7 be preinstalled, or their DAO program, which is also available to download, Fig. 5.  

Circuit applications
As explained by Leslie Warwick in the July issue, although video still cameras offer the ideal method to input illustrations into engineers reports, those offering better than VGA resolution are expensive compared to comparable quality conventional film cameras. In part this results from digital functions requiring a number of integrated circuits. Following collaboration between the Minolta Company Ltd. and LSI Logic, this looks set to change.  

The DCAM-101 chip incorporates all the major digital functions needed to make a high resolution camera, reducing costs and providing real-time operation rather than the delays inherent in other designs.  

This chip handles high resolution charge-coupled device arrays up to 2000 by 2000 pixels, can capture, compress to JPEG, and store 3.3 million pixels per second at 24 bits/pixel. But it consumes less than 1.2W. It has the potential to allow lower cost cameras but with improved resolution Fig. 6.  

Following the recent strong interest in audio topics and Professor Leach’s low transient intermodulation amplifier design in the September issue, I now include two quite different designs found on Internet.  

Nelson Pass, of PassLabs, has considerable experience in audio amplifier design. His Web page includes a do-it-yourself section offering a free choice from four differing variations on his ‘Zen’ amplifier. These choices are the Class A, A-40, A75 and Citation12 designs.  

Since recent published designs have tended to involve several stages with resultant complexity, I have chosen to show his minimalist ‘Zen’ approach, Fig. 7.  

Finally for those who prefer their active devices to run visibly hot, another novel approach is the ‘single-ended tube amp’ from High Fidelity Engineering. Design considerations and circuit for this ongoing design can be found on Decware’s page. This amplifier trades off the number of active stages, against dc current design complications in the output transformer.
Working with microwaves

Boris Sedacca looks at the problems involved with developing circuits for microwave frequencies – where capacitors and inductors become tracks on the PCB.

The ultimate test of any electronic gadget is whether or not it works when it is switched on. Before it gets to that stage, there is an exhaustive list of equipment to empty the deepest pockets. This is particularly true with microwave gadgets.

But if you can beg, borrow or steal time on a scalar network analyser and sweep generator, you can build something fairly cheaply. Most of the work is in the design and software simulation, and computer time is relatively plentiful.

Obtaining the right software again depends on the depth of your pocket. Microwave design software tends to be expensive, but you can often have a package for a trial period from the supplier.

Introduction

Why would you want to get into microwaves? Well, for a start you can eliminate many components.

If you design a receiver that works at 5GHz and beyond, then you can use microstrip. The advantage of microstrip technology is that the geometric shape of track on a substrate determines whether it behaves as a capacitor, inductor, filter, attenuator, mixer – or even an antenna.

Fewer physical components means lighter weight too. And microstrip is compatible with surface mount technology.

The downside? Producing a design is a lot more complicated. With the right design calculations, you can go straight into a CAD package like AutoCad and produce a mask, but before you do that, a microwave simulation package will prove immensely helpful. Conventional electronics CAD packages are of little use at microwave frequencies.

The two leading contenders in the microwave simulation software market, are Compact Software’s Serenade 7 and Hewlett Packard’s Series IV. First a little background.

Transmission lines

If you strike an ‘A’ tuning fork – i.e. one with an order of magnitude of 440Hz – and place it next to a piano or guitar string that is also tuned to A, the string will vibrate in tune because of the sound pressure waves propagated by the tuning fork. The surrounding air acts as a transmission line.

In a similar way, a pair of open wires carrying an alternating signal can propagate an electromagnetic wave through the air that can excite ac through an aerial at a particular distance away from the current source.

When a transmission line is terminated in an impedance other than its characteristic impedance, you get reflection.

Distributed amplifiers

To widen bandwidth, rf design engineers today have a method available called distributed amplification. Although distributed amplifiers involve microstrip technology, they are easy to set up on monolithic microwave integrated circuits, or MMICs, and much more difficult on hybrid circuits. MMIC technology did not exist of course, when Percival[1] originally patented his distributed amplifier in 1937, but apart from that, very few modifications of the original concept have surfaced.

The distributed amplifier produces relatively lower gain than the multiplicative amplifier but yields significantly larger bandwidths through the ingenious use of the active devices’ parasitics. Unlike conventional amplifiers, the higher the frequency it operates at, the better it works.

The problem is that it needs matched transmission lines and this is difficult if not impossible to implement in hybrid form. It only works efficiently if the parameters of every one of those transistors are identical and that can only be guaranteed by having them all from the same GaAs substrate. Even if they were identical the surface mounts are unlikely to be uniform.

GaAs fets capable of operation up to 100GHz have useful amplifier gains limited to around 10GHz. Ayasli[2] has provided a simplified equivalent-circuit model for a four-stage fet travelling wave amplifier. In this circuit, microstrip lines are periodically loaded with the complex gate and drain impedances of the fets, forming lossy transmission line structures of different characteristic impedance and propagation constant, Fig. 1.

An rf signal applied at the input end of the gate line travels down the line to the other end, where it is absorbed by the terminating impedance. However, the gate circuits of the individual fets dissipate a significant portion of the signal along the way. The input signal sampled by the gate circuits at different phas-
es—and generally at different amplitudes—is transferred to the drain line through the transconductance of the fets.

If the phase velocity of the signal at the drain line is identical to the phase velocity of the gate line, then the signals on the drain line add. The addition will be in phase only for the forward-travelling signal.

**S-parameters**

A microwave transistor’s S-parameter specifications is essentially a table of its behaviour at different frequencies, which allows us to interpret it as a two-port, four-terminal device where \( a_1 \) and \( a_2 \) are waves entering the device at ports 1 and 2, and \( b_1 \) and \( b_2 \) are waves leaving the device, Fig. 2.

The main columns of the individual rows of the table contain the following information where,

\[
S_{11} = \frac{b_1}{a_1}, \quad S_{21} = \frac{b_2}{a_1}, \quad S_{22} = \frac{b_2}{a_2}, \quad S_{12} = \frac{b_1}{a_2}
\]

The first three are the most important, with the last, \( S_{12} \) usually being so close to zero that it can be ignored.

These relationships can be represented in the form of an equation \([b]=[S][a]\) and this can be extended into a matrix,

\[
\begin{bmatrix}
  b_1 \\
  b_2
\end{bmatrix}
= \begin{bmatrix}
  S_{11} & S_{12} \\
  S_{21} & S_{22}
\end{bmatrix}
\begin{bmatrix}
  a_1 \\
  a_2
\end{bmatrix}
\]

For maximum power transfer, we need conjugate matching of complex impedances such that,

\[
\begin{bmatrix}
  S_{11} & S_{12} \\
  S_{21} & S_{22}
\end{bmatrix}
\begin{bmatrix}
  1 & 0 \\
  0 & 1
\end{bmatrix}
= \begin{bmatrix}
  1 & 0 \\
  0 & 1
\end{bmatrix}
\]

As an example, the Mitsubishi MGF1303B GaAs fet which retails around £5.50, has the following S-Parameters in the 10GHz row of the table:
You can even use a spreadsheet but a problem here can be that the equations are normally hidden from the user. I used Mathcad, but for heavy-weight engineering work, you may want to use MATLAB instead.

As I am discussing using parallel-coupled, or edge-coupled, microstrip lines instead of inductors and capacitors, the above values are converted to seven admittance inverters, or J, parameters using a characteristic impedance of 50Ω as follows. For the first coupling structure,

$$j_0 = \frac{1}{Z_0} \left[ \frac{\pi \delta}{2\delta g_1} \right]$$

for the intermediate coupling structures with $$k = 1...5$$,

$$j_k = \frac{1}{Z_0} \left[ \frac{\pi \delta}{2\omega_k \delta g_{k+1}} \right]$$

and for the final coupling structure,

$$j_6 = \frac{1}{Z_0} \left[ \frac{\pi \delta}{2\delta g_{k+1}} \right]$$

so,

$$J = \left( 0.011 \ 8.147 \times 10^{-3} \ 6.833 \times 10^{-3} \ 6.383 \times 10^{-3} \ 6.193 \times 10^{-3} \ 8.147 \times 10^{-3} \ 0.011 \right)$$

Note the symmetry of the coupled sections. Now the seven odd and even-mode impedances can be worked out with $$k = 0...6$$ as follows,

$$Z_n = Z_0 \left( 1 - j \times Z_0 + (J_k)^2 \times (Z_0)^2 \right)$$

and,

$$Z_{o_i} = Z_0 \left( 1 + j \times Z_0 + (J_k)^2 \times (Z_0)^2 \right)$$

giving,

$$Z_{o_6} = (37.58 \ 37.929 \ 38.74 \ 38.87 \ 38.74 \ 37.929 \ 37.58)$$

For a full explanation of these equations, refer to Edwards. The last value gives us an X-axis co-ordinate for reading off a chart of Chebyshev filter characteristics. Matthaei et al have produced charts and tables for passband ripple from (s pseudo 1 dB, and reading off -10 dB attenuation on the Y-axis gives the order of the filter needed. From that we get seven g-values from a table of prototype elements, plus a value for g0=1, as follows:

$$g = (1.21456 \ 1.1041 \ 3.0634 \ 1.1518 \ 2.9367 \ 0.8101 \ 2.6599)$$

At this stage, the calculations become incredibly tedious, so I recommend the use of a maths package like Mathcad or MATLAB. You can even use a spreadsheet but a problem...
copper-clad substrate that can be etched in a similar way to conventional PCBs. Although marketed as a prototyping material for microstrip, RT Duvroid is good enough for many mainstream production applications.

In this case, the substrate's thickness – height \( h \) is 0.635 mm, copper track thickness, \( t \), is 0.008 mm, and relative permittivity \( \varepsilon_r \) is 10.5.

It is assumed that centre frequency \( f_0 \) is 10 GHz, the free space wavelength,

\[
\lambda_0 = \frac{c}{f_0} = \frac{3 \times 10^8}{10 \times 10^9} = 0.03 \text{ m} = 30\text{ mm}
\]

where \( c \) is the velocity of electromagnetic waves in free space.

Now you can work out the odd and even-mode substrate wavelengths as follows,

\[
\lambda_{o,m} = \frac{\lambda_0}{\sqrt{\varepsilon_r}} \quad \text{and} \quad \lambda_{e,m} = \frac{\lambda_0}{\sqrt{\varepsilon_r}}
\]

giving,

\[
\lambda_{o,m} = 0.012 \text{ mm} = 0.012 \text{ mm} = 1.2 \text{ mm} \quad \text{and} \quad \lambda_{e,m} = 0.012 \text{ mm} = 0.012 \text{ mm} = 1.2 \text{ mm}
\]

Hence the average substrate wavelength \( \lambda_{av} = 0.012 \text{ m} = 12 \text{ mm} \).

The microstrip sections are quarter wavelengths and need to be shortened to take account of discontinuities like the open-end effect and step widths, involving more tedious calculations which the reader will probably be relieved to know are not included here for lack of space.

The Serenade 7 schematic in Fig. 3 shows the final dimensions of the filter in the variables block, or VAR. Starting with the control blocks at the top, FREQ sets the sweep generator from five to 15 GHz in increments of 50 MHz, SUB specifies the substrate, and OPT sets the optimisation goals.

In the VAR control block, \( W \), \( S \), and \( P \) are the widths, spacings and lengths respectively of each filter section. The question marks surrounding the \( P \) values signify that these are the variables to be optimised.

The microstrip circuit elements consist of the input and output ports \( P_1 \) and \( P_2 \). There are also two 10 mm lengths of transmission line TRL with characteristic impedance \( Z_0 \) of 50 ohms.

In this case of the substrate, Serenade 7's Transmission Lines utility output a width of 0.576 mm. The STEP elements takes account of step changes in widths, CPL microstrip elements of the coupled sections, and OPEN elements of the open-circuit end effects.

After optimisation, the output shown in Fig. 4 was obtained, showing a near-perfect filter characteristic – in theory at least. The \( S_{11} \) parameter is shown in black, \( S_{21} \) in grey. Now similar results were obtained with Series IV. However, Series IV does not need open-end elements because it has specific coupled filter elements, MCFL, which include open-end effects.

Furthermore it has a layout utility included in the product, whereas although Serenade 7 has its own layout utility based on AutoCAD, it uses a third party add-on called S2A.

At this point, the Smith chart becomes invaluable for transmission line design, both for lumped components and microstrip. This facility is available on both products, but I found it easier to use the conventional hardcopy version. I have not included it here for lack of space.

Polar co-ordinates of \( S_{11} = 0.704 - 163.1^\circ \) correspond to the complex rectangular form 0.18 – j0.145 on the chart, representing the source impedance \( Z_S \), the inverse of which is then converted to a source admittance \( Y_S = 3.37 + j2.71 \). The distance between these two values represents the diameter of a circle with its origin at the centre of the chart.

By extending the diameter from \( Y_S \) to the edge of the chart, you can read off the wavelength as 0.226. To get a resonant match, first move clockwise along the circle until it intersects the unity admittance circle, to get a value for \( Y_F = 1 + j2 \). Drawing a straight line from the centre of the chart through \( Y_F \), you can read off the wavelength as 0.313.

Now the length of microstrip can be found from the transistor to the stub, or distance \( d_{ps} = (0.313 - 0.226) \lambda_0 = 0.966 \text{ mm} \). Next, cancel the imaginary part by drawing another straight line from the centre of the chart through +j2 at the edge of the chart, giving a wavelength of 0.176. Therefore the length of the stub \( l_p = 0.176 \lambda_0 = 1.95 \text{ mm} \).

The same procedure is applied for \( S_{22} = 0.584 + 138.2^\circ \), representing the load impedance \( Z_L \), giving.
You can now simulate this design as shown in Fig. 6. From this the output shown in Fig. 7 was obtained, again with $S_{21}$ shown in blue and $S_{11}$ in red. Note the resonant peak of about 11dB in $S_{21}$ just below 11GHz. This matches published maximum available gain at that frequency for the MGF1303B fet.

Now the filter simulation matched reality quite accurately and the fact that it is a passive design makes all the difference. When it comes to an amplifier, this is an entirely different ball game because as an active device it needs power, which means that the gate and drain lines need to be biased using bias pads.

The measured results are shown in Fig. 8. This is nowhere near as close to the simulation as the results from the filter, but it does show a clear resonant peak at about 9.5GHz. Apart from the biasing, the variance may also be accounted for by the fact that the tracks on the circuit where narrower than those on the mask because of over-etching. This fact was established when the circuit was inspected using an Alpha step machine.

There are various output formats available from Series IV's layout facility including Gerber, AutoCAD and HP Graphics Language.

So much for resonant or narrow-band amplifier matching. As I mentioned earlier, to increase the bandwidth you can use a distributed amplifier setup such as that in Fig. 9.

Note the absence of stubs - just uniform lengths of transmission line from the ports and to the grounds, and twice the length between the amplifier stages - two in this case. For convenience I have used three-terminal, two-port models here to represents the GaAs fets, but the effect is the same.

Now the length and width of the line is crucial, and there are no straightforward analytical design formulas one can apply to derive these measurements. The only practical way is continued on page 871...

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ADC42 is a low-cost, high-resolution a-to-d converter sampling to 12 bits at 20ksample/s. This single-channel converter benefits from all the instrumentation features of the ADC200-50.
Unusually, this capacity meter draws power from the NiCd battery under test. It discharges battery packs of from four to ten cells, a charger being connected after the test. As an example, a four-cell pack applies 3.5V to the circuit after allowing for current sensing and reverse polarity protection.

Latching comparator IC1 starts the discharge when the start button is pressed, discharge ending when battery voltage falls below 1V/cell. In this case, the functions of output and hysteresis pins are reversed, as at low currents their threshold voltages differ by a few millivolts, the output being higher and preventing clean switching. Output current from the hysteresis pin is low and is amplified in Tr4, the resulting switch being capable of supplying over 100mA.

The discharge circuit is formed by IC3b and Tr4 and is switchable to 180mA and 1.8A by S2. Very good power supply rejection in the MC33172 (IC3b) renders the discharge current largely immune to battery-voltage changes; the prototype’s discharge current varies by less than 0.2% for a battery-voltage change of 4-25V.

To measure discharge time, the Maplin FS13P counter accepts its own 512Hz output, divided by the 4040 counter, to increment at two-second intervals while the battery is discharging. This is equivalent to 0.1 or 1mA/h per count on the low and high current ranges. Power for the counter timer comes from IC3a and Tr3, which form a low dropout regulator to supply the relay.

Lower discharge currents may be used; PP9 at 18mA can be handled, but the relay must be omitted. Higher currents or more cells will necessitate a heat sink for Tr4.

John R Hunt
Middlesbrough
Porch light saver

Anyone with a porch having inner and outer doors may have a 40W or 60W light in the porch both to show who is calling and to light up the keyhole of the inner door. It may also be a good idea to leave the light on much of the time to make people think you are at home when you are not. The trouble is, electricity is expensive and lamps last a very short time when on permanently.

You can avoid these problems by connecting a 2pF, 250V capacitor in series with the lamp, which will then be dim and possibly too dim. However, a door-switch mounted so that its contacts are open when the outer door is closed, can bypass the capacitor; when the door is opened, the light goes to full brightness. A 2752 resistor in series with the capacitor will reduce inrush current. There could be a second such switch on the inner door.

H T Wynne
Glasgow

Electronic control for dc motors

Designed for applications in which starting and stopping of a dc motor must be smooth over 0.5-5s this controller costs little and uses pulse-width modulation for speed control. It will take motors of between 5W and 50W although, at low acceleration rates, it is happiest when the load is the same for each start.

For speed control, the voltage on C1 is compared with the sawtooth on C2, produced by the 555 300Hz oscillator IC1, in IC2b. The result is a 300Hz pulse at the output of IC2b whose width is roughly proportional to the voltage on C1.

To start, S1 is set to on; since C1 has not begun to charge, the output of IC2b is high, Tr3 is conducting and power is applied to the motor. The voltage on C1 rises rapidly, charging current coming via S1, Tr1, Tr2 and P1, until it reaches the voltage on pin 3, whereupon power to the motor is cut off. This pulse of power to the motor lasts between 50ms and 200ms, set by P1, to suit the motor in use.

In acceleration, P3 is set to produce a pulse width, after the start pulse, that runs the motor slowly. Since the output of IC2b is low, C1 charges further, but more slowly through P3, to the voltage set by P3. The motor accelerates to a speed set by the adjustment of P3, output pulses now being wide enough at maximum speed to produce, effectively, the full dc level. To make adjustment of P2 and P3 easier, a 47µF capacitor across C1 slows everything down to make acceleration and deceleration take about 1s.

Decelerating, when S1 is off, C1 discharges towards 0V by way of P3, the output pulse width becoming shorter and the motor slowing. When the voltage on C1 reaches the voltage on pin 3 of IC2a, the output again goes high. C1 now discharging rapidly through the same path as for rapid charging. The output pulse is now zero and the motor stops cleanly with no jitter.

As regards emc, the circuit has been tested; emi from the circuit was unmeasurable, since it was masked by noise from the motor brushes, itself within limits due to the use of the L1,2 ferrite beads and C6.

Rolf Schmidt
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A plethora of components include resistors, capacitors, inductors, mutual inductors / transformers, controlled sources, bipolar junction transistors, zener diodes, power MESFETs, JFETs. MOSFETs, voltage regulators, operational amplifiers, out-couplers, voltage comparators, quartz crystals, 10B/10 buffers and switching matrix connectors and much more. All devices and model parameters can be edited to suit your needs. Implement hierarchical circuits in your designs quickly and easily.

**No Limits**

With B²Spice and B² Logic there is no limit on the number of components in the circuit.

**Models**

There are literally thousands of them. The complete Berkeley SPICE model library as well as commercial libraries from manufacturers such as Motorola, Texas Instruments, Burr-Brown, Maxim, National Semi, APEX Comlinear, AMP Elantec, Linear Tech, and many more. Included with B²Spice is a full model and symbol editing package so you can create, import and edit custom models.

**Commands**

B²Spice supports AC frequency sweep, DC operating point, transient analysis, fast fourier Noise, sensitivity distortion, TI small signal transfer.

**Simulation Options**

Added facility for sub-circuits (macro-models). You can set all simulation options, allows you to set initial conditions at all nodes. Allows you to set initial guess at nodes for simulation. Allows "not given" state for all values.

**Total Control**

B²Spice gives full access to Berkeley SPICE simulation control options. For example you can set global defaults for transistor channel lengths and widths. Plus much more.

**Waveform Analysis**

Display and compare multiple results in a single graph at the same time. All SPICE simulation results can be selectively displayed and analysed graphically and in tabular or numerical format as well as exported to other applications. All of B²Spice and B² Logic's display capabilities are completely flexible.

**Devices & Stimulus for Simulation**

B²Spice sinusoidal, constant, periodic pulse, exponential, single frequency or AM, DC voltage, AC voltage, Vcc, Vcc, piecewise linear, exponential, polynomial, arbitrary sources, voltage-controlled voltage, voltage-controlled current, current-controlled voltage, current-controlled current, Lossy and ideal transmission line, MESFET uniform RC, MOSFET uniform RC, current and voltage switches are all available.

**Cross Probing**

Cross probing allows you to display waveform results simply by marking pins, wires and devices on the circuit drawing. Monitor results while the simulation is in progress then plot analogue results on linear or log scales.

**Data Analysis**

Position detection with mouse for data points. Import and export data to and from other industry standard SPICE programs. B²Spice supports Polar, Smith and Nyquist charts.

**Digital Options**

B²Logic is completely flexible. Set up ROM, RAM and PLA to your own requirements. Simulate a whole circuit to a block and use it as a component in a new design. Run the simulations in real time or step by step. Program your own custom libraries. Create your own custom libraries. Create and run pre-programmed simulations.

Professional engineers need software that produces results they can rely on. Anything less is a liability. B²Spice & B² Logic will give you the accurate results you need fast.

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CIRCLE NO. 133 ON REPLY CARD
**Power-on delay**

After several seconds delay following switch-on, this circuit applies power to the equipment and, after a break in the input caused by, perhaps, accidentally switching off, the circuit resets and imposes a further delay.

Figure 1 shows the circuit, R1,2 and C1 being the relevant components. Voltage across C1 increases exponentially to a voltage set by the divider R1.4; these values having been chosen so that the voltage on the capacitor at the output reaches a value large enough to activate a comparator or buffer after the required time. The rest of the circuit is to avoid the necessity to make R2A small for a rapid reset, which would result in a very small voltage on C1.

The Darlington transistors and R3 perform this function by providing an alternative discharge path for C1. Low voltage at the input causes C1 to start discharging through R2A, causing a voltage drop across R2 and turning on Tr1.

Current flows through the low-value R3 and C1 discharges rapidly, Tr2 assisting the process by switching on in response to the voltage developed across R3 and maintaining Tr1 in conduction for the rest of the discharge time.

To obtain a usable delay, it is only necessary to use a comparator at the output designed to switch at 3.5V.

The use of Darlington transistors and the inclusion of Tr2 are only needed when the discharge must take place in a few tens of microseconds.

My thanks to Richard Wessen of University Technology Sydney for simulating the circuit.

Martin Gosnell
Warriewood
New South Wales
Australia

**Ac/dc voltage detector/isolator**

When the ac or dc input to the circuit shown is over 50V and below about 250V, output is 5V; otherwise, it is zero.

Output is optically isolated from the input. Since the input goes to a bridge rectifier, it may be ac or dc. Bridge output is dropped across R1, smoothed and regulated to some extent by the 10V zener. Resistor R3 limits current to the isolator’s led.

When input rises to over 50V, current flows in R3 and the output reaches 5V. Input voltage limit is about 0V-250 V. The optoisolator shown, a 610A-2 x001, is a high-voltage type and is recommended.

Keith W Saxon
Deva Medical Electronics
Runcorn

Work out the wattage needed for R1 for maximum expected input – Ed.
a powerful combination

For the past ten years Weir Electronics has been part of the Lambda Group – the largest power supply company in the world. Weir are now trading under the name Weir Lambda, to show that Weir is part of this major worldwide force. Same company, same people, same trusted power supplies and power systems, but with all the benefits of being part of the largest power supply company in the world – a partnership that packs a powerful punch.

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Sub-woofer filter extends bass response

The response of a loudspeaker in a sealed enclosure is that of a second-order high-pass filter, the cut-off frequency and Q being easily determined for any enclosure size.

Such a speaker fed through a filter and power amplifier produces a flat response below its resonant frequency, forming a simple, but effective, sub-woofer.

A constant 12dB/octave bass boost, produced by the circuit shown, causes the low-pass system response to become a high-pass one, Q and cut-off frequency being the same as before.

Inputs to the circuit are from left and right speaker outlets, which are added at the virtual earth of the integrating op-amp. This gives a 6dB/octave drop and is followed by a second 6db/octave circuit, the pair providing the 12dB/octave output to feed a power amplifier.

Jeff Macaulay
Chichester
Sussex

Mains connection checker

If you have no meter handy and need to know whether a wire has mains on it, this little device will provide the answer.

An antenna consisting a few inches of stiff wire feeds any signal received to the input of a CD4060B 14-bit counter/oscillator, used as a divide-by-16 counter. If the antenna is held near a wire carrying mains voltage, the led flashes at about 3Hz; if the wire is not connected to the mains, the led is either lit or unlit, but does not flash regularly.

I mounted mine on the top plate removed from an old PP3 battery, one of its press connectors being mated to one of battery studs. To switch the device on, I simply rotate the circuit on its top plate until both connectors make contact. A drop of wax as shown prevents the circuit being switched on accidentally when not in use.

S Arnesen
Oslo
Norway

Irregular use of regulators in headphone amplifier

For a few pounds, you can build a headphone amplifier, which gives very good sound quality and which is quite capable of deafening you.

It is a single-ended, direct-coupled Class A amplifier whose main claim to fame is its use of 317T voltage regulators instead of fets or bjts in the output. Quiescent current for the output stage is set up by IC3; its adjust pin is taken to negative and the Vset pin is 1.25V more positive, current being defined by that voltage divided by R4.

The other 317T is the output amplifier, which is in the feedback loop of the op-amp. Its adjust pin produces a constant 50µA sourced by the op-amp to keep it in single-ended operation whatever the signal polarity. Feedback maintains the output voltage and input voltage at the same level, compensating for the 1.25V between IC3 adjust and output pins.

The regulator ics should be on a heat sink and electrically isolated.

Jeff Macaulay
Chichester
Sussex
Back issues of Electronics World are available, priced at £2.50 in the UK and £3.00 elsewhere, including postage. Please complete the coupon and send with correct payment to:
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Quick stop for induction motors

Although this method of rapidly stopping and locking induction motors is known, patents covering it have not described this simple variant.

Single-phase and three-phase motors will stop very quickly and be locked if a dc source is connected to the motor. In this arrangement, the dc comes from a bridge across the ac input, which is switched in and out, as is the motor supply.

The possibility of thermal cut-out means that the switch must be released fairly quickly and, if the motor stops too precipitately, a resistor in the dc circuit will slow the action down.

Scott Arnesen
Oslo
Norway

Simpler than usual circuit to stop dead and lock an induction motor by applying dc to it.

Differentially tuned LC oscillator

Tuning of the bridge oscillator shown is linearly proportional to the difference in value between the two variable capacitors $C_x$ and $C_y$, frequency increasing when $C_y$ increases and decreasing when $C_x$ increases.

If $R_2 = R_4$, frequency is determined by the equation,

$$f = \frac{1}{2\pi \sqrt{L_1 (C_x + C_y)}}$$

when $R_3 = R_1$. Resistor $R_1$ represents parallel losses in the LC circuit and $R_3$ the sum of $R_2$ and the equivalent resistance of the LT1228 op-amp in the amplitude stabiliser, which is controlled by rectified output voltage from the oscillator.

The graph shows a linear change in frequency for a change in either of the variable capacitors.

Lech Tomawski and Mariusz Slawiec
University of Silesia
Katowice
Poland
Direct-reading inductance meter

This inductance meter will measure down to 0.1µH and up to 100mH, displaying the result on a digital voltmeter. It uses much the same principle as capacitance meters in which the pulses from a monostable flip-flop are of a length proportional to the capacitance.

The monostable here is lila, which drives +5V

L2

220mH

R3

100R

L3

2

R1

1mH

2k2

100

74HC132

SW1a

T

21:C2

C1

10n I I 100n

(R1 1)

R2

10k

Tr1

BC239

04

10µ

+5V

Swib

Pulse -integrating inductance measure inductances from 0.1µH to 100mH. Components to be measured resistance to avoid affecting constant of the monostable.

x

SW1C

IC1b

IC1c

T2T1

R5

100k

+5V = 111

R4

220k

R9

150k

'VVV

102

Op amp

R8

-5V

100k

To voltmeter

Pulses at U1c output are integrated and applied to the output op-amp, which scales the voltage level and sets zero for the digital voltmeter. I use a 4.5-digit meter on the 2V range, the three inductance ranges being 1mH, 10mH and 100mH.

Flavio Fontanelli
Genova
Italy

1-180s timing in two steps

Analogue timer MC14536 provides intervals that vary by a factor of two for each position of a hexadecimal switch. Intermediate steps are not possible without the addition of a separate divider chain. This circuit interposes steps of two to give, with sixteen positions of the switch - a progression 1, 1.4, 2, 2.8... 180.

To obtain the intermediate steps, capacitor C2 is connected by the least significant bit pin of the hexadecimal switch, the capacitor having a value 2.2 times that of the main timing component C1. Other switch outputs go to the timer programming inputs A, B and C and the switch output c to 6V in response to the binary code corresponding to the sixteen switch positions. On even positions of the switch, C2 is floating, while for odd positions it is connected to 6V and the clock based on R1,2C1 slows down.

The extra capacitor must be larger than the original timing capacitor since the waveform at the bottom of the clock components is attenuated, besides being slowed by C2; R1 allows the use of a preferred value for the extra capacitor. Minimum frequency of the timer must be 512Hz to take account of the internal divider between the clock and the monostable output circuit; this falls to 362Hz when C2 is connected.

Stability is about 1% over a supply range of 6-9V. Pulse output from pin 13 will drive an earpiece or led, but ensure that pickup on C3 does not cause spurious oscillation.

C J D Catto
Cambridge
LETTERS

Letters to “Electronics World” Quadrant House, The Quadrant, Sutton, Surrey, SM2 5AS

Feedback

Having read Mr Ellis’ letter in the September issue with attention, I am of course immeasurably pleased at being invited to ‘re-optimise’ my amplifier design.

I agree entirely that the Miller capacitor looks as though it would do a lot more good when wrapped around the output stage as well, because this promises that the crossover distortion could be greatly reduced. Unfortunately, no amount of wishing will alter the fact that this connection is prone to instability because of the extra phase-shifts in the output stage. Mr Ellis admits this when he says “I did observe some oscillation...” I have no doubt he did. I would take this capacitor speculation more seriously if someone published some measurements and proved it really did work.

On a model amplifier – small-signal output devices, hence much quicker – I found the effect on crossover spikes disappointingly small.

I fear Mr Ellis has not completely understood the subtle nature of the generic amplifier circuit; the second stage does not and cannot load the input stage. I profoundly disagree with his labelling the second stage as the ‘driver’ when everyone else calls the devices directly before the output transistors the drivers. It seems to me there is quite enough confusion about, without getting The Naming of Parts wrong as well. The second stage provides all the voltage gain, so calling it the voltage-amplifier stage seems to me pretty uncontroversial.

The voltage-amplifier stage does not load the input stage, because the input pair is a transconductance amplifier – i.e. voltage-in, current-out. The voltage-amplifier stage on the other hand is a transimpedance stage – current-in, voltage-out – which would ideally have zero input impedance. This is the whole basis of the pole-splitting action that makes this beloved circuit so dependable.

If Mr Ellis had studied my Blameless design a little more closely, he would have seen that the input-pair local feedback measures he recommends with such enthusiasm have already been implemented. Indeed I thought I was one of the main advocates of this way of minimising the Miller capacitance required, though in the way I implemented it I called it ‘constant-gm’, degeneration.

If the input emitter resistors are increased significantly beyond 100Ω the noise performance is degraded; life is full of compromises. In selecting for praise the Lohazoh & Oota amplifier he has chosen a bad example. I know this circuit very well indeed, for, unlike as it may seem, in 1975 I was entrusted with the task of putting a version of this design into quantity production. While I have doubt made my share, and more, of mistakes during the development period, the eventual conclusion of us all was that the design was over-complex, inefficient, non-linear, and of very doubtful high-frequency stability. Ironically, in this design, loading of one stage by the next really is an issue.

There are other points I could make, but I fear to little purpose. One question remains; Mr Ellis must realise that his understanding of the subject is imperfect, as it is for everyone in varying degrees. Why is he so confident he is justified in lecturing the rest of us?

Douglas Self
London

Thomson’s electron

Your timely article by Tom Ivall in the August 1997 edition of Electronics World on J J Thompson’s 1897 discovery of the electron and its technical context was interesting and informative.

I am particularly interested in J J Thompson himself, his origins, family and further works (if any) he accomplished. Has any reader any suggestions of previous articles or other material to point the researcher in the right direction?

S Katzen
Magee College Belfast

Singing preamp

A long time, I built a version of John Linsley Hood’s simple Class A amplifier, (E, April ’89). The output transistors were 2N3055s mounted directly, i.e. without insulation, on two rather inadequate ‘Christmas-tree’ heatsinks.

For some reason, while I was driving the amplifier hard with a music signal, with the speakers disconnected – like you do – I found I could still hear a very faint, tinny and distorted version of the music, with a large peak in its spectrum at the frequency at which the heatsinks would ‘ring’ if flicked with a finger.

The sound loudest near the top surface of the output transistor cans and was not affected by moving the wiring around, nor by holding magnets near the transistors. The only factor that affected it was the load; connecting a 6.8Ω resistor across the output terminals made the sound louder.

I wonder if this could be a thermal effect. Does anyone know?

Chris Bulman
Bedford

Remote control enhancement

I have an enhancement to Alex Birkett’s circuit ‘Two-wire remote control’ in Electronics World’s June issue.

In the original circuit, if two buttons are pressed together, the output corresponds to an undefined voltage band and is not representative of either button. This voltage depends on the haphazard parallel connection of two resistors.

I have inserted in the parallel resistors a series resistor chain with buttons to short the ‘cold’ end of each resistor to ground, as in the diagram, accidently pressing two buttons together results in an output corresponding to one of the buttons pressed – specifically the one nearer the ‘hot’ end of the resistor chain. This is arguably less confusing to the user.

Enhancement to the two-wire remote control circuit to make accidentally pressing two keys give more predictable results.

Singing preamp

I was glad to see Nick Wheeler’s sensible and timely article on skin effect in relation to loudspeaker cables, but it may be worth mentioning a small additional point.

In rf applications, where skin effect causes currents to be almost entirely near the surface of a conductor, I believe that the use of silver plating is almost never justified simply because silver is a slightly better conductor than copper. The real advantage is that oxidised silver, though unwelcome in appearance, is a far better conductor than oxidised copper.

Bob Pearson

Silverskin

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Bob Pearson
NEW PRODUCTS

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ACTIVE

Linear integrated circuits
Regulator ics. Allegro offers the A8188S Series of low drop-out voltage regulators for use in battery-powered equipment. Output is 2.5-3.3V at 250mA peak, the pmos series element allowing a drop-out voltage of 90mV at 60mA. Quiescent current is 55µA and does not increase much near drop-out; sleep current is under 1µA. A bandgap reference provides stability to within 10mV and regulation to between 25mV and 700mV at 60mA. Quiescent current is 3.3V at 250mA peak, the pmos series powering equipment. Output is 2.5-3.3V.

Microprocessors and controllers
New PICs. From Microchip, the new PIC16C77 range of 5mips eeprom-based, low-power microcontrollers provided with 8K by 14-word of program memory, 376byte of data ram and an 8-channel, 8-bit a-to-d converter with sample-and-hold, 11bit accuracy and a 16µs acquisition time. Other facilities include a real-time clock, one 8-bit and two 16-bit counter/timers. There are two channels of 160ns resolution for input, compare and pwm. Communications facilities, with 33 pins for i/o, include an 8-bit asynchronous slave port, a synchronous serial port for SPI and I²C, a 6.25Mbs uart with baud-rate generator, two 8-bit pwm outputs and direct led drive and tric interface. Arrow-Jermyn. Tel., 01234 270027; fax, 01234 214674/791501.

Motors and drivers
Small motor clutches. Denso's Wrapping clutches provide torque transmission up to 8.5Nm in small packages, their low inertia allowing greater acceleration than that obtained using rare-earth servomotors at lower cost. Unidirectional or bidirectional clutches in normally-engaged and normally-disengaged versions are available. In multi-axis systems, the clutches allow several axes to be started or stopped quickly at a speed synchronised to that of a single drive motor and the ability to engage and disengage under electronic control allows drive to be taken from other sources such as handwheels or petrol engines. Denso Corporation. Tel., 01707 278692.

Memory chips
3.3V, 64Mb dynamic memories. Smart Modular Technologies has announced new 3.3V, 64Mb, 72-pin dram-based modules operating on 5V designed for use in multichip workstations needing 1Gbyte or more. Two problems are thereby solved: voltage translation and a step-down converter built in these 3.3V simm devices to work in existing 5V systems; and four times as much memory, compared with existing designs using 16Mb dram, become available without the need to redesign the motherboard. Smart Modular Technologies. Tel., 01908 234030; fax, 01908 234191.

Discrete active devices
Efficient GaAs fets. MGF900 Series n-channel, Schottky gate GaAs fets from Mitsubishi are intended for uhf high-power use. Three new types, 0909A, 0910A and 0911A are 2.3GHz devices, the 0909A providing 45% high-power added efficiency when compared to other similar devices with a power gain of 11dB at 200W power input at 2.3GHz. Power output is typically 30dBm. The 0910A is similar but for Class A operation, while the 0911A, also for Class A, has added efficiency of 40%, at 2.3GHz, output power is 41dBm and power gain 11dB. Mitsubishi Electric UK Ltd. Tel., 01707 276100; fax, 01707 278692.

Optical devices
White leds. Light emitting diodes from Sloan AG emit white light with no mixing of primary colours, giving a typical brightness of 400mcd at 20mA and 3.6V. Colour temperature is 8000K and viewing angle 60°. They are rated at 100mA. Roush Electronics Ltd. Tel., 01724 281770; fax, 01724 281650.

Microwave components
700100; fax, 01959 700300.

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PASSIVE

Cameras
Small ccd camera. Sony offers the Mini Unit CCD family of ccd cameras, which measure 18.3 by 7.1mm, complete with timing generator, sample and hold ic, digital signal processing and lens, together
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with the ccd. Output is conventional Y/C analogue and there is auto white balance and autoexposure. Lenses are 2.9mm or 4mm fixed focus and resolution is 220 tv lines in the centre of the picture. Colour, exposure, white balance and minimum gain are all programmable by external eeprom. Sony Semiconductor Europe. Tel., 01256 478771; fax, 01256 816194.

Passive components

Toroidal transformers. Drake Transformers can offer 1kVA toroidal transformers with dual secondaries of 41, 45, 50, 55 and 120V and dual primaries for supplies of 120 or 240V, 50/60Hz. Outer insulation can be specified as polyester tape wrap or woven tape with a varnish finish. The transformers are designed for use in cold-cathode fluorescent lamp power supplies for display back light in lce televisions, camcorders, laptops, etc. Output voltage is typically over 1kV and the range of power available is 0.7W to 6W, with a choice of input and output voltage. Sizes range from 12 by 11 by 5mm for the 0.7W type to 21 by 23 by 6.5mm for a 6W version. Dielectric strength of the independent primary and secondary is over 2.5kV. Drake Transformers Ltd. Tel., 01256 560040; fax, 01256 560261.

Backlighting transformers. Toko’s BLC Series of miniature inverter transformers are designed for use in cold-cathode fluorescent lamp power supplies for display back lighting in lce televisions, camcorders, laptops, etc. Output voltage is typically over 1kV and the range of power available is 0.7W to 6W, with a choice of input and output voltage. Sizes range from 12 by 11 by 5mm for the 0.7W type to 21 by 23 by 6.5mm for a 6W version. Dielectric strength of the independent primary and secondary is over 2.5kV. Criket Distribution Ltd. Tel., 01992 444111; fax, 01992 464457.

Connectors and cabling

Ultra-SCSI connector. Honda has a slim, high-density cable-to-board connector to meet the Ultra-SCSI specification, having two rows of phosphor bronze contacts at 0.8mm spacing. The contacts are underplated with nickel and then plated with palladium and gold. IDC termination is present and the connectors use jackscrews for retention and to ensure shielding. Honda Connectors. Tel., 01793 523388; fax, 01793 521777.

S-m signal/power board connector. Molex Microfit surface-mounted board connectors are believed to be the first true S-m types to handle both signal and power lines, the 3mm-pitch pins withstandting 1.5kV ac and a current of 250V ac at 5A per contact. A solderable hold-down tab avoids the need to drill holes in the board. Connectors are available with 2-24 contacts in even numbers and there is polarisation and positive latching. Flint Distribution. Tel., 01530 510333; fax, 01530 510275.

High-density board connectors. Robinson Nugent’s PXS-5 low-profile, high-density, SMT board-to-board connectors have, as an option, a floating contact in the receptacle to cope with potential torsional effects and up to 0.3mm of discrepancies caused by tolerance build-up. They are available in sizes from 20-100 positions and mating stacking height of 5mm. Contact pitch is 0.5mm. Also available are bellows contacts to allow stacking of 3-mm to resist mating stress and contact resistance of 80mΩ maximum. Robinson Nugent (Burndy) Ltd. Tel., 01256 842626; fax, 01256 842673.

Mixed-signal ICs

Pwm drivers. M S Kennedy MSK4220/1 board-mounted motor/driver/amplifier modules contain all the drive and control circuitry in the one hybrid chip. Each has a complete H bridge and switching circuitry and the modules are suited to the drive of brush motor speed control and Class D amplification. In the case of the 4220, one analogue signal is needed to control motor speed and direction or an audio signal for amplification. Both devices are on a ceramic metal-coated substrate for direct mounting on heat sink. Ashwell Electronics Ltd. Tel., 01438 364194; fax, 01438 313461.

Camerawideccd camera. PEL466 by Premier gives 291000-pixel picture from a 1/3-in interline ccd sensor. Either a 3.6mm lens or a pinhole 3.7mm fl 4.5 is provided as standard and there is a range of other lenses to choose from. The camera works in light levels down to 0.5lux, the electronic iris working from 1/4sec to 1/100000sec. running on internal or external synchronisation. Video output is 1Vp-peak at 75Ω and a 12V, 130mA supply is needed. Premier Electronics. Ltd. Tel., 01922 634652; fax, 01922 634616.

Displays

Bitmap lcd driver. The NJU6583 driver by NewJapan Radio drives 33 by 86 dot graphics displays, containing 3696 bits of display data ram, microprocessor interface circuitry, an instruction decoder and 32 common and one icon common by 96-section lcc drivers. The data is transferred to display data ram by serial or 8-bit parallel interface. Standby current is 0.05μA and operating voltage 2.4-6V. Young-ECC Electronics. Tel., 01628 810727; fax, 01628 810807.

Fitters

900MHz saw filters. Designed for the GSM and AMPS mobile communications market, Fujitsu’s FSCE-920 is a high performance, compact filter allowing 3mm square in a surface-mounted LCC package. The filters use a double-mode surface acoustic wave design to provide a stopband attenuation of about 60dB with anpass band ripple of under 1dB. Temperature drift is 33ppm/°C. Fujitsu Microelectronics Ltd. Tel., 01628 76100; fax, 01628 781494.

Chip filters. Mitsubishi’s LT Series of surface-mounted chip filters are made using a new technique to be produce a steep insertion-loss curve and an insertion loss of –30dB to over 1GHz. They do not need to mounted with any specific orientation and are made in a composite dielectric and ferrite material to allow a number of filter types such as T-section and pi-section versions to be produced. First available are T filters cutting off at 10, 22, 47 or 100MHz, rated at ±3dB to over 1GHz running on internal or external synchronisation. They are 1Vp-peak at 75Ω and a 12V, 130mA supply is needed. Premier Electronics Ltd. Tel., 01922 634652; fax, 01922 634616.

Hardware

Rackmount keyboard drawer. Taking only 1U of rack space, the TR-KBD keyboard drawer made by Tri-Map is claimed to be simple to fit, since mounting a lce all at the front. The drawer has ball-bearing slide rails and there is a space for a wrist rest, one version of which has a trackball compatible with PS/2 or serial mouse ports. Standard Cherry
Please quote “Electronics World” when seeking further information

Radio communications products

Data communications. Free from Licence requirements, the TX2/RX2 transmitter and receiver board-mounted modules from RF Solutions provide data comms at 490kb/s over a distance of 300m on open ground or 75m in buildings. Full screening has minimised radiation and the need for external filtering is thereby reduced. Both are in sit-mounting modules, operating on the European 433.92MHz or on the UK 418MHz. The transmitter is a saw-controlled fm type, while the receiver is based on a double-conversion fm superhet, taking 14mA when active. RF Solutions Ltd. Tel., 01273 488880; fax, 01273 480661.

Test and measurement

250Msamples/s dso. LeCroy’s 9344 digital storage oscilloscope provides 500MHz of analogue bandwidth at 1Gsamples/s and 250K points of memory. Four 250Msamples/s, 50Kpoint channels may be combined to form two 500Msamples, 100Kpoint or the one 1Gsamples configuration. When running at under 200Msamples/s, there is 5ns peak detect for fast transients and ‘smart’ trigger functions to capture intermittent or complex conditions characterised by pattern, interval, dropout, etc. The instrument will carry out over 40 parametric measurements, showing average, low and peak values and standard deviation and a pass/fail function tests up to five parameters against selected thresholds, waveform limit test is made against masks defined in the instrument. There is a 9in raster display and a full range of peripheral facilities. LeCroy Ltd. Tel., 01169 344882; fax, 01169 349000.

Oscilloscope calibration.

Oscilloscope calibration workstations based on Wavetek’s Model 9500 Calibrator are introduced in versions with bandwidths of 400MHz, 600MHz and 1.1GHz, upgrading from low to high being provided. All have a new version of the company’s pc software, which includes an up-datable library of procedures for commonly available analogue and digital oscilloscopes. All are provided with an Active Head, which applies signals directly to an instrument’s BNC input to avoid the distortion and other errors caused by long cables. Active heads will not turn into inactive heads on being dropped, since they have already been dropped and survived the test. The software renders calibration automatic, so that even the most wondrous instruments can be done in ten minutes, reports and certificates being generated. Wavetek Ltd. Tel., 01603 404824; fax, 01603 483670.

Emi test system. Two instruments which have commonly been used together in emission testing are now sold as a package at greatly reduced cost. ES-Plus consists of the Rohde & Schwarz ESPC 15kHz-1GHz (expandable to 9kHz-2.5GHz) first receiver and the R4131C 10kHz-3.5GHz spectrum analyser by Advantest, the two being accompanied by the R&S ESPC-K1 Windows software. Emc accessories by R&S can be used with both instruments. Rohde & Schwarz UK Ltd. Tel., 01252 611377; fax, 01252 811447.

Clamp-on power analyser. Hick’s 3166 power Hiester provides clamp-on measurement and analysis of single or three-phase power lines. An lcd gives instantaneous values of current and voltage, averages and frequency, max/min display and first and second integration displays, which show reactive/apparent power consumption. Options are a floppy disk drive and harmonic analysis. An RS232 interface is standard. Telonic Instruments Ltd. Tel., 01734 786911; fax, 01734 792338.

Pulse generator. TGP110 from TTI is a general-purpose pulse generator working at frequencies between 0.1kHz and 10MHz, providing independent control of pulse width, repetition rate and variable pulse delay. Pulses of a fixed width may be generated at any frequency down to a duty cycle of 1 in 105, pulse width range being 50ns-5s and period range from 100ns to 10s. Delay is independently adjustable over the same range as the width. As well as continuous pulses, single or multiple pulses may be generated by trigger or gating signals and there is an inverting switch. Square wave operation is also provided at a frequency set by the period control to provide variable-period edges. Thurlby Thandar Instruments Ltd. Tel., 01480 412451; fax, 01480 450409.

Literature

Numerical analysis software. National Instruments’ HiQ interactive problem-solving software is the subject of a new brochure, in which are described functions such as data fitting, numerical integration, data visualisation and report generation. The package will produce 2D and 3D graphs from data and functions and accepts data from Excel and LabView. HiQ is for Windows NT/95 and Mac. You can obtain an evaluation copy on www.natinst.com/hiq. National Instruments UK. Tel., 01635 572400; fax, 01635 524395.

Power conversion. Switch on to smart power conversion is a Celab publication describing the company’s power supplies for military and communications use that provide battery backup and charging, management and diagnostics and variable output measures (for ‘smart’, read ‘intelligent’). Celab Ltd. Tel., 01420 477011; fax, 01420 472034.

Emi and thermal interfaces. Product selection software from Chomerics gives guidance on choice of emi shielding and grounding and thermal interface materials. mPhorm uses...
multiple-choice questions to allow the user to make an application profile, the program then recommending some products, obtained from all Chomerics books, bulletins and brochures, to fill the bill. It faxes or prints data on all the products and the selections are savable and can be added to forms for quotations. InPhorm comes on or off on twelve 3.5m fopplels, being compatible with all embodiments of Windows. Parker Hannfil plc, Chomerics Division, Tel., 01628 486938; fax, 01628 476938.

Sensors. Lucas Control Systems can supply a new brochure on its Schweis sensing products, which include devices to sense and measure pressure, flow, temperature, inertial sensors for slope and tilt and instrumentation. Newly introduced are non-contacting laser sensors and new pneumatic switches for anti-corrosive and dirty environments. Lucas Control Systems Products. Tel., 01753 537622; fax, 01753 835253.

Dsp hardware and software. Innovative Integration's catalogue of products for digital signal processing and development tools is now available. For 1997, the catalogue describes a wide range of products, including the PC32 low-cost, 32-bit supercontroller with a 32-bit and d-to-a conversion, the ADD64 64-channel, 32-bit, PCibus data converter board, ViSiMind DSP software/hardware for the design of control systems and DASYLab data acquisition software. Adept Scientific Micro Systems Ltd. Tel., 01462 480055; fax, 01462 480213.

Industrial products. Farnell has a new catalogue of equipment for industrial electronics – separate from the electronic components catalogue. Monstran would like to point out that, among the 32000 products, its accelerometers and linear variable displacement transducers with instrumentation are on offer. Farnell Components Ltd., Tel., 0113 2205700; fax, 0113 2891163.

Signal conditioning. JTech's publication, The Signal Conditioning and PC-based Data Acquisition Handbook, is both a primer on the subject and a quick reference guide for experienced engineers. There is an introduction to the use of transducers with pcv by way of data acquisition systems which describes the common interfaces. Discussions on device control requirements, multiplexing, amplification, noise reduction and digital signal conditioning. Scientech Scientific &Engineering Systems Ltd. Tel., 01296 397676; fax, 01296 397678.

Materials
Safety matting. Plastic Extruders offers Vynastat, which is a vinyl matting designed to protect electronic equipment and employees from the effects of electrostatic discharge. It is made from semi-conductive pvc, is slip-resistant in the form of an open grid and possesses permanent static dispersal properties. Vynastat comes in 10m rolls, 60cm or 91cm wide. Plastic Extruders Ltd. Tel., 01883 623329; fax, 01883 625506.

Power supplies
Lithium-ion battery protector. New from Unibro is the UCSn911 two-cell lithium-ion battery-pack protector, which has an on-chip fet switch for increased reliability; it copes with overcharge, overdischarge and short circuits, overcharge and overdischarge. A band-gap voltage reference detects overcharge or overdischarge by comparing it against a fixed voltage, which is in a feedback loop that controls the fet and allows for pack recovery. In the overcharged state, the loop only allows discharge current to pass the fet and in overcharge, only charging current passes. In the overdischarge state, the chip sleeps until charging current is detected. Battery power drain is 20µA and 3.5µA when shut down. Unibro (UK) Ltd. Tel., 0181-318 1431; fax, 0181-318 2549.

Low-cost psus. XP's NANN55 Series of universal-input power supplies are supplies in single, dual and triple versions, delivering 55W with forced air cooling, 45W without, from a standard 5 by 3 by 1.2in module. They fully comply with all kinds of safety standard and are protected against overvoltage, shorts and conducted noise to EN55022 level B. Outputs available are 5V, +5V, -12V or 5V and ±12V, both available in ratings of 500W, 750W and 1kW supplies. Coutant-Lambda's RP Series of universal-input power supplies are available in peak power and regulated versions. XP plc. Tel., 01734 845515; fax, 01734 843423.

1kW supplies. Coutant-Lambda's RP Series of power supplies are multiple-output, ac-fed types for large computer installations such as workstations, file servers and industrial controllers. They are available in ratings of 500W, 750W and 1000W, are CE-marked and may have one, two or four outputs, the main, high-power output being adjustable from 3V to 6V. Other outputs may be 12V, -12V, 3.3V, 5.5V, 24V or 48V. Inrush current for universal and power factor correction is included, and all output protection is provided by an output switch. XP plc. Tel., 01734 842211; fax, 01734 843423.

Touch screens. Data entry touch screens from Rafi, using a five-wire technique, are versatile enough to be tailored for individual needs. Operating speed is high, resolution is 1096 by 4096 points, sensitivity 2mm and contact pressure under 100g. The screens are sealed and protected against moisture and dirt and, since they are designed for point-of-sale use, vinegar and ammonium products and cleaners. Fingers, gloved or ungloved and other objects can be used on the screen, which has a light transmission of 75%. All necessary electronics are supplied, including driver software for dos, Windows, Apple Mac, OS/2 and Unix. Rafi (GB) Ltd. Tel., 01737 778660; fax, 01737 778772.

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Switches and relays
Pcb relays. Meisei board-mounted relays come in the popular switching forms and ratings, are enclosed and have standard 2.54mm pin spacing. Series relays are four-pole, double-throw types switching up to 2A at 100V ac/dc, in an 11.5mm high housing. For dual-in-line mounting, the M1, M3 and M4 types switch 30W dc or 60VA ac and are meant for telecommunications: M1 is a dpdt type with 5-48V coils. M4 relays provide high sensitivity, taking 6.3mA at 48V dc and 29.9mA at 5V dc, while the M3 versions and higher, which is the MFC1SR, is 1.6 by 2in. An aluminium substrate helps with heat transfer. Power Control Ltd. Tel., 00353 61 474133; fax, 00353 61 474141.

Telemetry receiver. Wood and Douglas has replaced its TR450 with the STR450 synthesised telemetry receiver, which is physically and electrically compatible with the earlier model. Improvements include better large-signal handling, an fsk data extractor and either a 1200baud modem or controller options. STR450 draws less than 11mA from 5.5-14V and meets ETS 300 220 and MPT 1329. Frequency range is 450-470MHz or 350-490MHz to order, the operating frequency being programmed by exchanging a disposable PIC microcontroller or by an internal 2400baud Rs232 serial port from a pc. A -80dB local oscillator notch filter is available as an option. Wood and Douglas Ltd. Tel., 0118 9811444; fax, 0118 9811557.

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Transducers and sensors
Disc thermistors. New miniature disc thermistors from ES are for general use and come in a wide range of values and two sizes. The 3mm type have values between 500Ω and 30kΩ, the 5mm version ranging from 5Ω to 1MΩ. The screens are sealed and protected against moisture and dirt and, since they are designed for point-of-sale use, vinegar and ammonium products and cleaners. Fingers, gloved or ungloved and other objects can be used on the screen, which has a light transmission of 75%. All necessary electronics are supplied, including driver software for dos, Windows, Apple Mac, OS/2 and Unix. Rafi (GB) Ltd. Tel., 01737 778660; fax, 01737 778772.
output proportional to the applied conditions and produces a rail-to-rail industrial, automotive and other trying and rotary position sensing in things. The device is suited to linear by varying temperatures, among other circuit to cancel offset voltage caused linear Hall sensor is uses a chopper Linear Hall sensor. Allegro's A3516 217704; fax, 01234 217083.

Small load cells. In a range of sizes from compression types (micro) of 12mm to compression and tension versions (UTC) of 25-4mm in diameter, new load cells from Control Transducers will handle up to 5000kg in 14 ranges. The measuring elements of bonded strain gauges form a 3500 four-arm Wheatstone bridge taking 10V ac or dc excitation to produce 20mV maximum. Temperature range is -10°C to 65°C. Diameter of the smallest is 1in.

Linear Hall sensor. Allegro's A3516 linear Hall sensor uses a chopper circuit to cancel offset voltage caused by varying temperatures, among other things. The device is suited to linear and rotary position sensing in industrial, automotive and other trying conditions and produces a rail-to-rail output proportional to the applied magnetic field at a sensitivity of 2.5mV/gauss (gauss are still common in the USA). It is virtually self-contained, possessing a high-gain amplifier, voltage regulator, filter and the offset cancellation circuitry. Allegro MicroSystems Inc. Tel., 01932 253355; fax, 01932 246622.

Optical shaft encoder. HD Series Dipped by Control Transducers are to an open design to keep the cost down, but have the characteristics of enclosed types. Output of this non-contacting design is digital, resolution being 96-2042 lines per rev. Dual, double-sealed ball bearings ensure very low torque and allow speeds of up to 10000rev/min. Current taken is 40mA from 5V and the output is two-channel ttl-compatible for pulse multiplication and direction sensing. Diameter of the smallest is 1in.

PORTABLE SYSTEM

Portable system. AstroDAQ 2, introduced by Astro-Med Inc., is a data acquisition system for conditioning, analysing and networking data and is contained in a portable case about the size of a notebook computer. It uses Windows-based software for control and analysis or real-time waveform monitoring and display on a pc. The instrument will monitor data remotely while sending results via a modem to a central pc; twenty 5kHz channels can be recorded, regardless of the number acquired, data being stored on an internal 1Gbyte hard disk. All signal conditioner modules are dbsp-based, modules for temperature, bridge, voltage, strain and motion being available. Up to ten units may be controlled by one pc. Astro-Med Inc. Tel., 01628 668836; fax, 01628 664994.

Card reader. Omron's SGR family of latching card readers handle magnetic-stripe and/or smart-card reading processes over more than half-a-million passes. The readers use a solenoid-controlled mechanism and red/green leds to show the mode in use, all in the family being configurable to read when the card is pushed or pulled. They support ISO and CP8 chip contacts and are compatible with ISO1, 2 and 3 magnetic stripe standards. A 5V supply is used. Kestronics Ltd. Tel., 01727 812222; fax, 01727 811920.

Development and evaluation

TMP93CM41F starter kit. Toshiba announces the Topas 960 starter kit for the development and evaluation of the company's 16-bit TMP93CM41F microcontroller. The kit contains an evaluation board, Asl C compiler, assembler and Windows-based debugger, with documentation, software examples and data sheets on cd. With the board comes a serial cable and a power supply and there are connectors for access to mcu signals. Code downloaded to sram on the board can be tested using the source-level debugger, which is a subset of that used with the company's real-time ice. Arrow-Jermyn. Tel., 01234 270027; fax, 01234 214674791501.

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Energy efficiency is today's watchword and, since much equipment is not in continuous use, small, standby power supplies are in demand.

Television receivers, Internet equipment, fax machines and other equipment spending a significant time switched on but somnolent, often use a standby power supply instead of the normal one when in contemplative mode, to reduce power needs and operating cost. Power Integrations Inc. makes the TOP209 low-cost, switched-mode power supply for this purpose which, when compared with discrete switchers, is half the size and weight, uses many fewer components and needs little power itself. An application note on the subject is available.

Figure 1 is the TOP209 functional block diagram and Fig. 2 shows a typical arrangement of a dc-input flyback circuit using the device. Referring to Fig. 1, Drain is the output and also provides internal bias current on starting, all pins do several things; Control is the input to the error amplifier, carries feedback current for the pwm circuit and connects to the shunt regulator to give internal bias cur-

---

**Fig. 1.** Functional block diagram of the TOP209 standby power supply chip for output power to 4W.
rent in normal working; the Source pin is common and is actually two pins, normally connected together, one being for the high-voltage return.

Control voltage
Voltage \( V_c \) on the control pin supplies all controller and driver circuits and, via an external capacitor shown between control and source in Fig. 2, ensures that gate drive for the output fet is available.

Regulation of this voltage comes in two forms: for start-up and during an overload condition, the regulation is hysteretic in form, as in Fig. 3; in normal operation, the shunt regulator amplifier takes over. On start-up, current to supply the internal circuitry and also to charge the external capacitor, \( C_5 \) in Fig. 2, is supplied by the high-voltage current source between drain and control pins, shown symbolically in Fig. 1.

At switch-on, \( V_c \) reaches 5.7V and the current source turns off, having charged \( C_5 \) and the pwm modulator and output stage are turned on. In the circuit shown, the feedback derived from the extra secondary transformer winding keeps \( C_5 \) charged and supplies \( V_c \). Normally, the shunt regulator maintains \( V_c \) at 5.7V by shunting excess control current through \( R_E \) which determines the error amplifier gain.

In a fault condition in which \( C_5 \) discharges to the lower threshold of 4.7V, the current source turns on and recharges \( C_5 \); the mosfet output stage turns off, the control circuit being now in a standby condition. A 1V hysteresis in the auto-restart comparator results in the waveform seen in Fig. 3, where the current source is turned on and off, a counter preventing the mosfet coming on again until eight cycles have taken place; the resulting 5% duty cycle limits power dissipation. This whole cycle of events repeats until the fault condition is removed.

Controlled by the voltage across \( R_E \), the pwm modulator drives the mosfet in a duty cycle inversely proportional to the current coming into the control pin, the \( R_E \) voltage being compared with the sawtooth from the oscillator to produce the duty cycle ratio, a clock from the oscillator turning on the mosfet via a latch and the modulator turning it off again.

Protection
If an overvoltage occurs, the high-current pulse into the control pin activates a latch, which turns off the output. Either removing and restoring input power or pulling the control pin below the lower threshold resets this latch and normal operation is resumed. As regards a too high temperature, there is an analogue circuit to act in much the same way as the overvoltage circuit, reset happening in the same way.

Practical circuit
The inexpensive 5V, 2W standby supply in Fig. 2 takes its input from a high-voltage dc source provided by rectified and smoothed 85-265V ac, its output being regulated, even lower cost being obtained by using a transistor regulator or even a zener shunt, albeit at lower output power.

In all cases, the TOP209 drives the primary with \( C, R \) and diode providing voltage-spike protection and reducing ringing at the drain pin. The other secondary winding provides the feedback to the control pin, charging \( C_5 \) to provide bias for the TOP209. Capacitor \( C_5 \) also filters the voltage to the control pin and sets the auto-restart frequency.

Fig. 2. 2W supply using the TOP209. Using this device, costs are lower and component count, size and weight greatly reduced.

Fig. 3. Waveforms in TOP209 operation. The sawtooth is the auto-restart section, which repeats this operation until a fault condition is removed, power dissipation being thereby reduced.

Sequoia Technology Ltd., Tekelec House, Back Lane, Spencers Wood, Reading RG7 1PW. Tel., 0118 9258000; fax, 0118 9258020.
RF DESIGN

Fig. 11. Simulated performance of combined filter and two-stage distributed amplifier. Grey curve S21 is forward transfer gain while the black one is S11, the input reflection. Note pronounced ripple.

continued from p. 46

to optimise. The variable block in Fig. 9 shows the values obtained after optimisation, around 0.2mm for both. This is almost impossible to implement in hybrid form.

Figure 10 shows the theoretical response of such a design. Note how the bandwidth has spread compared with the resonant match amplifier. This is particularly evident from the trace for S11.

Finally, Fig. 11 demonstrates the theoretical effect of putting the filter at the inputs and outputs. Now if I could tidy up that ripple and increase the gain above 20dB, I could sell such an amplifier for about $600 a throw.

Then I wouldn't have to scrape a living as a computer nerd any more. Ah well, one can only dream.

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text/reference examines the complementary nature of electronics and optics and emphasizes high-speed technology in which the two fields are less differentiated. Beginning with an overview that develops a perspective and appreciation of analog high-speed technology in general, the book goes on to cover devices and circuits used at microwave and millimeter-wave frequencies, optical components, and optoelectronic integrated circuits and subsystems. Particular attention is paid to applications in the area of high levels of interest in this area and because many of the concepts are applicable in other fields. The book concludes with important coverage of the often-overlooked area of measurement and characterization of high-speed devices. Fully referenced and supplemented with hundreds of helpful illustrations, Introduction to High-Speed Electronics and Optoelectronics is equally useful as a professional reference or a textbook for senior undergraduate and first-year graduate courses.

RISC Systems and Applications

Professor Daniel Tabak has completely revised and updated his two previous books on Reduced Instruction Set Computer architecture to produce this new book, RISC Systems and Applications. This highly practical intensive introduction enables electrical engineers, applied physicists, and students to develop and identify tools for understanding, analysis, design, and characterization of high-speed components. Broad in scope, this unique introduction examines the complementary nature of electronics and optics and emphasizes high-speed technology in which the two fields are less differentiated. Beginning with an overview that develops a perspective and appreciation of analog high-speed technology in general, the book goes on to cover devices and circuits used at microwave and millimeter-wave frequencies, optical components, and optoelectronic integrated circuits and subsystems. Particular attention is paid to applications in the area of high levels of interest in this area and because many of the concepts are applicable in other fields. The book concludes with important coverage of the often-overlooked area of measurement and characterization of high-speed devices. Fully referenced and supplemented with hundreds of helpful illustrations, Introduction to High-Speed Electronics and Optoelectronics is equally useful as a professional reference or a textbook for senior undergraduate and first-year graduate courses.
As industrial and embedded control systems become ever more complex, control-system designers increasingly adopt a distributed architecture for system layout. Rather than having one, large processing centre with sensor inputs and actuator outputs, distributed control systems provide a means of breaking down the complexity of the tasks involved in system control. These individual control systems each concentrate on their own task, but as the need for higher levels of sophistication gains pace, so does the need to pass information and data between these centres, or nodes.

One versatile distributed architecture communications system that can be used on a macro or micro scale is controller area network, or CAN. A serial bus system, CAN is useful as an embedded communication system for microcontrollers, and as an open communication system for intelligent devices. This makes it a very attractive bus. Being a real-time serial bus system with multi-master capabilities - allowing several CAN nodes to request the bus simultaneously - this bus system is well suited for networking 'intelligent' devices as well as actuators with sensors within a machine or plant. It is suitable for difficult and harsh electrical environments, where a high degree of real-time operation and ease of use are needed - and at low cost.

Not only in automotive areas
Originally, CAN was developed for use in vehicles. Today, it is increasingly used in industrial, building automation and other applications. It is already used as a standard in some areas, in addition to the automotive industry, although mainly in Europe. Mercedes-Benz is using CAN in its S Class series of vehicles for fast transmissions of up to 1Mbit/s.

And the more widely used CAN is, the more likely it is to appear in unexpected areas such as agricultural, nautical, instrumentation, medical, textile machines.

In safety-critical applications, integrity of information exchange is paramount. In the field of medical engineering, CAN is typically selected since it meets the stringent safety requirements. Similar problems are faced by manufacturers of other equipment with very high safety or reliability requirements. Examples of such equipment are robots, lifts and transportation systems.

The relative simplicity of the CAN protocol means that very little cost and effort is needed on training and the CAN chips interfaces make applications programming relatively simple. Low-cost CAN controller chips, allowing simple connections to microcontrollers, have been available since the last decade.

To spring 1996, over 10 million CAN nodes had been installed. Today there are more than 25 CAN protocol controller chips from over ten manufacturers. And as the CAN protocol steadily gains in popularity the CAN chip's availability is guaranteed.

Reliability issues
One of CAN's features is high transmission reliability. In CAN networks, instead of addressing subscribers or nodes, the transmitted messages are prioritised. A transmitter sends a message to all CAN nodes and each node then responds to the message through the bus. CAN uses a non-destructive arbitration protocol, meaning that each node's message is checked for priority before being transmitted. This ensures that all nodes can send messages simultaneously without interference.

As several nodes can require to send messages on the CAN bus simultaneously, conflicts can arise. These can be resolved by non-destructive bitwise arbitration: each node's identifier can be in one of two states: dominant - logic 0 - or recessive - logic 1. During the arbitration, the dominant state overwrites the recessive state. So the 'losing' nodes become message receivers and do not attempt re-transmission until the bus is available again.

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decides on whether to process the message or not, based on the identifier received. The identifier also determines the priority of the message as it competes for bus access.

Such a content-oriented addressing scheme allows a high degree of system and configuration flexibility. As the data transmission protocol does not require physical destination addresses for individual components, it is particularly suitable for modular electronics and permits multiple reception – i.e. broadcasting and multicasting.

This addressing scheme also makes CAN suitable for synchronising distributed processes. In such applications, measurements needed as information by several controllers can be transmitted via the network in such a way that it is unnecessary for each controller to have its own sensor. Suitably, nodes can always be added to an existing CAN network without making any hardware or software changes to the others.

Each CAN message may have from 0 to 8 bytes of user information. Longer data information can also be transmitted by using segmentation. The maximum transmission rate specified so far on a CAN network is 1Mbit/s over lengths of up to 40m. For longer distances the data rate is reduced. For example, for distances of up to 500m, the specified data speed is up to 125kbit/s and for transmissions of up to 1km the data rate is around 50kbit/s.

When data is transmitted by CAN, no nodes are addressed. Instead, the content of the message, which can be a parameter such as revolutions-per-minute or engine temperature, is designated by a unique identifier. The identifier specifies the content and the priority of the message. This is important for bus allocation when several nodes are competing for bus access.

If the CPU of a given node wishes to send a message to one or more nodes, it passes the data to be transmitted and its identifier to the assigned CAN chip. Known as ‘making ready’, this is all the CPU needs to do to initiate data exchange.

The message is formed and transmitted by the CAN chip. As soon as the CAN chip receives the bus allocation – a condition called ‘send message’ – all other nodes on the CAN network become receivers of this message ‘receive Message’.

Having received the message correctly, each node in the CAN network performs an acceptance test to determine whether the data received is relevant to that node, an operation called ‘select’. If the data is of significance for the node concerned it is then processed – called ‘accept’ – otherwise it is ignored.

**Error handling**

When an error is detected, the CAN controller registers it, as well as which node it is coming from. It then evaluates it statistically in order to take appropriate measures. This could lead to the disconnecting of the CAN node that has produced the error.

The CAN protocol supports two message frame formats – standard or Version 2.0A – and extended CAN or Version 2.0B, Fig. 1. The only difference between the two is in the length of the identifier.

In the standard format the length of the identifier is 11 bits and in the extended format the length is 29 bits. The extended format CAN messages with 29-bit identifier was introduced after the American-based Society of Automotive Engineers, SAE, ‘Truck and Bus’ subcommittee decided to standardise signals and messages in addition to the data transmission protocols for a variety of data rates. Standardisation of this kind turned out to be easier to implement when a longer identification field is available.

For a standard message, the transmission message frame consists of seven main fields. A message in the standard format begins with the ‘start of frame’ bit. It is followed by the ‘arbitration field’, which contains the identifier and the ‘remote-transmission request’ bit that indicates whether it is a data frame or a request frame.

The ‘control field’ contains the identifier extension bit that indicates whether the format is standard or extended, a bit reserved for...
CONTROL ELECTRONICS

The ISO 11898 implementation of the physical CAN connection is just one of a number of options.

future extensions and a count of the data bytes in the data field.

Length of the 'data field' can vary from 0 to 8 bytes. It is followed by the cyclic redundancy check 'CRC field', which is used as a frame security check for detecting bit errors. The acknowledgement field, or 'ACK field', comprises of a one-bit ACK slot and one recessive bit ACK delimiter.

The bit in the ACK slot is sent as a recessive bit and is overwritten as a dominant bit by those receivers which have at this time received the data correctly, typically dubbed positive acknowledgement. Correct messages are acknowledged by the receivers regardless of the result of the acceptance test.

The end of the message is indicated by 'end of frame'. 'Intermission' is the minimum number of bit periods separating consecutive messages. If no other nodes demand to access the bus then the bus remains idle.

Extended format messages

In extended format transmission messages the 29-bit identifier consists of the existing 11-bit identifier (base ID) and an 18-bit extension, also known as the ID extension.

Distinction between standard format and extended format is made using the IDE bit, the so-called identifier extension bit. This is transmitted as dominant in the case of a frame in standard format. For frames in extended format it is recessive.

The RTR bit is transmitted dominant or recessive depending on whether data is being transmitted or whether a specific message is being requested from a node. In place of the RTR bit in standard format the substitute remote request, or SRR, bit is transmitted for frames with extended identification.

The SRR bit is always transmitted as recessive, to ensure that in the case of arbitration the standard frame always has priority bus allocation over an extended frame when both messages have the same base identifier.

Unlike the standard format, in the extended format the IDE bit is followed by the 18-bit identification extension, the RTR bit and a reserved bit (r1). All following fields are identical with the fields in standard format. Conformity between the two formats is ensured by the fact that the CAN controllers which support the extended format can also communicate in standard format.

The two formats can easily coexist on one bus. However their relevant messages are prioritised differently. The standard CAN message version always has priority over the message in extended format. This way, collisions of messages during requesting bus access are avoided.

Extended format CAN controllers can also send and receive messages in standard format. On the other hand, CAN controllers which only cover the standard format, or Version 2.0A, can transmit only standard format messages on that network. Otherwise, the extended format messages will be misunderstood.

There are CAN controllers however which can recognise extended messages even though they only support standard format. In these cases the extended format will ignore those messages. As a result, it has been marked as the Version 2.0B passive format.

Error signalling

The CAN protocol signals errors that occur on the network rather than use acknowledgement messages as in other bus systems. For error detection, the CAN protocol has three mechanisms at the 'message' level and two mechanisms at the 'bit' level.

At the message level, there are the cyclic-redundancy check, frame check and ACK errors. At the 'bit' level, errors mechanisms can be one of two types: 'monitoring' or 'bit stuffing'.

Cyclic redundancy checking safeguards the information in the frame by adding redundant check bits at the transmission end. At the receiver end, these bits are re-computed and tested against the received bits. If they do not agree there has been a CRC error.

Frame checking verifies the structure of the transmitted frame by checking the bit fields against the fixed format and the frame size. Errors detected by frame checks are designated 'format errors'.

Each CAN node can transmit or observe bus signals, Fig. 2. That way they can easily detect the difference between the bits sent and bits received, which indicates an error. This is 'monitoring'. With 'bit stuffing' the sender inserts into the bit stream a bit which can be removed by the receiver. This permits reliable detection of all global errors and errors local to the transmitter.

When errors are discovered by at least one node by means of the above mechanisms, the current transmission is aborted by sending an 'error flag'. This prevents other nodes from accepting the message which ensures the consistency of data throughout the network.

After transmission of an erroneous message has been aborted, the sender automatically re-attempts transmission – an action entitled automatic repeat request. There may again be competition for bus allocation. As a rule, re-transmission will begin within 23 bit periods after error detection and in special cases the system recovery time is 31 bit periods.

Lock-ups

Although this method of identifying and isolating a defective node proves effective and efficient, it may occasionally lead to all messages – including correct ones – to be aborted. This can block the bus system if no measures for self-monitoring are taken. The CAN protocol therefore provides a mechanism for distinguishing sporadic errors from permanent errors and localising node failures.

This is done by statistically assessing the node error situations. The aim of this is to still be capable of recognising a node's defects and if need be enter an operating mode where the rest of the CAN network is not negatively affected. This may go as far as the node switching itself off to prevent the abortion of correct messages that have been misinterpreted as incorrect.

On the other hand, errors can go undetected. However the statistical average of that happening on a CAN network operating at a data rate of 1Mbit/s with an average bus capacity of 50% is one undetected error in every thousand years.
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