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The nature of power

A Balkan war involving UK and European troops will undermine European and UK electronics industry interests: shooting wars are expensive, distracting and usually serve no economic purpose. A vicious little war on our doorstep might safeguard a few defence industry jobs in the short term but it is really no substitute for economic policy.

The stupidity of some politicians is breathtaking. They actually contemplate committing further armed forces to this and a number of other trouble spots around the world without seeming to consider the reasons for doing it. Politics is about power. Wars are a way of projecting raw power. A thinking politician would only commit armed forces where there are clear economic goals. The Gulf war was a sensible one in this respect. The Balkan tribes have been knocking eight varieties of God out of each other for several thousand years and there is every reason to decline the invitation to their party.

I recently attended a rather different gathering, a celebration to mark the opening of GEC Plessey’s 3in GaAs wafer fab line at its Caswell research facility. It represented to me the real war that we should be fighting. Routinely writing half micron lines on a brittle piece of III/V semiconductor with a dice yield of up to 80 per cent was not nearly as impressive as the fact that just 40 per cent of production will be sold into military applications. The plant’s management anticipates that the bulk of its output will find civil use, some of which you don’t yet know that you need.

For instance, the back of the average office computer is a mess of wires. Devices made at the new GP semi plant incorporate all the difficult bits of a 2GHz microwave modem as a single chip for a price which we will be happy to pay.

Another example would be low cost electro-optical devices for superhet data transmission systems. These devices will power the revolution that information technologists have been promising us for so long.

The sort of development typified by the new production facilities at Caswell ought to be regarded as a heavy weapon in an economic war that we should be willing to fight.

There is no constructive value in a seat on the UN Security Council; if there had been, it would have long been occupied by Japan or Germany. Both of these countries profess institutionalised pacifism. In reality, they have come to the sensible conclusion that there is more to be gained from economic rather than shooting wars. We should do the same.

Frank Ogden.
Government bans sale of unbuggable phones

The DTI is blocking exports of the GSM digital cellphone system because it believes the encryption used is too powerful. High level sources say this is because the security services in the UK and US fear they will no longer be able to monitor telephone calls. These sources want the matter, and the DTI’s handling of it, debated in public.

Industry chiefs say the DTI has woken to the problem five years too late, creating a muddle which is crippling trade.

In January the technology desk of the DTI’s press office said it was aware of the problem, and knew that the DTI stood accused of fouling the market. But internal politics obliged the technology desk to refer enquiries to the trade desk where a spokeswoman, who knew nothing of GSM, took note of questions on the DTI’s export block and the source of the encryption algorithm.

Though she was given the names of the DTI officials responsible for GSM, after 24 hours she still refused to arrange a conference call with anyone inside the DTI who could discuss the matter sensibly. Finally she passed the matter back to the technical desk where someone who did understand the question tried unsuccessfully to get BT to talk about its work.

This experience tallies with the pantomime that four of the firms trying to sell GSM equipment (Motorola, Ericsson, Nokia and AT&T) have described. They were all in Bahrain at the end of January for a conference on Arabic communications, and their frustration was running over.

GSM was developed in the mid 1980s by the Groupe Special Mobile, now part of the European Telecommunications Standards Institute, ETSI. European manufacturers and telecommunications authorities shared the work. The technology was officially blessed by European Commission Directives in 1987 and the standard has been agreed in a Memorandum of Understanding (MoU) signed by 27 operators in 18 European countries.

The EC’s plan and MoU promised a pan-European GSM service by 1991. This would allow business travellers to roam, using the same portable phone anywhere in Europe with calls billed back home. This is impossible with existing cellphone services because different countries use different analogue technology.

GSM has been slow to take off in Europe because the existing analogue services are too successful, so manufacturers have been looking to export. The name was changed in 1990, to Global System for Mobile Communications, to make this easier.

Virtual reality set to boldly go

Virtual reality is viewed either as the stuff science fiction is made of, or as a set of games devised by hippy programmers for those who want to go one step beyond Nintendo and become part of the action.

But VR tools are creeping into such diverse and arcane areas as molecular modelling, medical imaging, architectural design, and the offshore energy industry.

While some of the VR acolytes have claimed year after year that its rise to dominance is imminent, the next couple of years could see their wildest dreams come true.

By the middle of the 1990s, synthetic realities will be commonplace tools in the engineering and scientific fields. The rapid pace of processor and software developments, along with falling prices with regard to power, will lay the foundations for a completely different computing model. Moore’s Law for processors, which stipulates that transistor density on a device should double every 18 months, is one example of such momentum on the hardware front. This Law is being realised through increasingly powerful products, like the Alpha and Pentium chips, coming out at regular intervals.

It can cost little more than £10,000 today for a workstation that can run the best 3D visualisation and design packages. As the cost of systems able to generate VR falls to affordable levels, its use will become more widespread. This process will be aided by the parallel growth of multimedia and advanced 3D computer-aided design.

VR is a computer-generated environment in which the user is immersed and can interact directly with objects contained in the so-called virtual world. This would allow an architect to walk through a virtual building – based on actual plans – long before it is built. Similarly, chemists can
In November 1990 the Commission of the European Communities warned of the need "to work towards the lifting of any obstacles concerning the export of GSM technology". But it took until 1992 for the DTI to raise its objection on encryption. This followed a decision by the US government to reject GSM and use the D-Amps system, which has weaker encryption.

Whereas all existing cellular phone systems transmit speech as analogue waves, GSM (and PCN) converts speech into digital code running at a low data rate (13Kb/s). Although existing scanner radios, as used to eavesdrop on analogue cellular calls, cannot decode digital speech, no-one doubts that Far Eastern manufacturers will soon start selling scanners that can.

With this in mind the GSM designers, including BT, built encryption into the standard. The system is called A5, and is similar to the US government's Data Encryption Standard. The US government has always been very worried about the export of any system which relies on DES. When software company Norton included DES encryption of text in its Utilities package, the feature had to be removed for sale outside the US.

Either DES or A5 encryption would bar eavesdropping in real time. This is what alarmed the FBI, which wants to listen in to mobile phones. It also alarmed GCHQ in Cheltenham, which monitors all radio traffic round the world.

With its close involvement in Gulf politics and special relationship with the US, and BT's input on A5, the UK is spearheading the push to block exports of GSM technology without special licence. It has asked for revision of the GSM standard, either by watering down of A5 to A5X, or by omission of encryption altogether.

This means GSM equipment makers must re-design their microchips. But they cannot start until the A5X standard is set by ETSI; the earliest hope is for May. Any change will inevitably split the standard and this will rob GSM of its major selling point, freedom to roam between countries with the same phone. Manufacturing costs will rise too as mass production benefits are lost.

Although Middle Eastern states are hungry to buy GSM, manufacturers dare not sell without clearance from the DTI. They all have manufacturing ties with the UK and fear black listing by the US and UK governments.

Middle Eastern states all want to use Europe's GSM technology. As market research company EMC in the UK notes, all the countries have state-run telecommunications authorities and all are oil rich and can afford to buy what they like. Qatar and the UAE want to be first with GSM in the Gulf, with Bahrain next. But the firms making GSM equipment cannot sell it. This creates the risk that the market will be lost to rival digital systems from the US and Japan.

In Bahrain, Motorola, Nokia, AT&T and Ericsson all told the same story. They do not know which countries outside Europe they are allowed to sell A5 GSM to, and cannot get answers from the DTI.

Nokia, tendering for the Bahrain GSM contract, says, "There is no logic. We don't know what is happening or why. How can we sell a global system which is not global?"

Nokia says it "hears rumours" that sale to the UAE has been cleared, but cannot get confirmation.

Ericsson says it "thinks that Hong Kong and Singapore and Australia have been cleared too", and says there is "some indication" that NATO countries may be approved. "But it is always difficult to get anything in writing."

AT&T claims to have clearance from the DTI to sell in the UAE and expects to sign a contract "imminently".

Motorola, which has already won an order from Qatar, says it has to contend with a "double whammy" because it must get clearance from the US and UK governments.

Motorola says: "It's rumoursville. The whole industry is running on rumours. If you find out what is happening, we'd like to know. We tried to find out from the DTI. Gosh how we tried, without any success. Perhaps the oxygen of some unfavourable publicity will wake the DTI to the harsh reality of hard business."

Barry Fox

render their molecular compounds in ways which allow them to see complex mathematical relationships and even give the equations a tactile quality.

The means of interaction and the level of immersion are variable. They range from the full head-mounted display (HMD) and data glove as a pointing device combination, to using an electronic wand or other pointing device to manipulate images on a television or computer screen. The goggles usually contain two liquid crystal displays presenting an optimised perspective for each eye. The gloves have sensors which replicate hand movements as actions in the artificial reality seen by the user.

But here lies the present cost penalty incurred through using advanced virtual worlds. Getting the user totally immersed in a synthetic environment is expensive and graphics hardware and software will have to improve substantially in performance as well. It takes a vast amount of memory and processing power to convert gestures into real-time action in VR, apart from the demands of creating the environment in the first place. A good quality HMD can still cost up to £1 million, but again the price will fall.

VR is starting to appear as a hybrid technology, combining visualisation with a higher degree of interaction. At last year's Siggraph graphics event in the US, Sun Microsystems demonstrated a holographic workstation. Equipped with 3D goggles and a mouse, the Sun machine lets the user do things like machining on a virtual lathe. The mouse moves the cutter onto the edge of a cylinder and this action is accompanied by relevant sounds from the workstation's built-in audio hardware. This system is available on the market, with third party hardware. It is being used for geographical information systems, such as road walk-throughs to see how a planned development could be improved.

Such hybrid applications do not necessarily need high quality rendered graphics, but they are delivering workable VR in some form today. Rolls-Royce is using VR to improve servicing and Nuclear Electric is simulating robot work in a similar way. And the aircraft simulation business, which spawned the first VR systems, is now using the technology to create synthetic air traffic control situations, helicopter cockpits and the view through a tank periscope.

The computer industry will have to be convinced it can make a profit from VR before it is embraced wholesale. VR attracts the standing joke among industry executives that it is the world's fastest growing zero billion dollar business. Silicon Graphics, a workstation and leading graphics software vendor, is another company to believe in the value of non-immersive - and therefore cheaper - VR. The company attributes a fifth of its $2 billion turnover to desktop VR systems in one form or another. This reveals the value of the underlying VR business happening today.

One of the most significant boosts to the growth of VR should occur this year.

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40GHz GaAs line gives UK lead in Europe

The UK’s first high volume gallium arsenide chip making plant has come on stream with the opening of a new facility at GEC-Plessey’s Caswell research facility. The £20 million wafer line provides almost totally automated handling of GaAs 3in wafers resulting in a production cost of around $2 to $3 per mm² of chip area. The process includes direct electron beam writing on chip which can build 0.25μm gates, small enough to produce transistors operating up to 40GHz. This, together with robotic cassette handling of the fragile wafers, makes the Caswell facility the most advanced of its kind in Europe and on a par with the best US and Japanese plants.

GEC Plessey expects to process only 3in wafers, making the Caswell facility the most advanced of its kind in Europe and on a par with the best US and Japanese plants. The £20 million wafer line provides almost totally automated handling of GaAs 3in wafers resulting in a production cost of around $2 to $3 per mm² of chip area. The process includes direct electron beam writing on chip which can build 0.25μm gates, small enough to produce transistors operating up to 40GHz. This, together with robotic cassette handling of the fragile wafers, makes the Caswell facility the most advanced of its kind in Europe and on a par with the best US and Japanese plants.

The Caswell process includes a couple of features which make it particularly suited to monolithic microwave IC manufacture. Two layers of passivation are available. The first, silicon nitride, allows high quality dielectric capacitors to be created on chip and a layer of metal interconnect. A second polyamide plastic polymer layer can be metalised with a further layer of device feature interconnect.

Caswell can also drill micro holes right through the die which may then be metalised to provide a third layer of metal interconnect on the back of the wafer. Production device yields are said to be in the region from 20 to 80 per cent.

According to plant manager Dr Fred Myers, the bulk of Caswell’s production will be for civil and commercial applications such as its wireless lan transceiver chip which allows office computers to communicate by 2.4GHz radio link at up to 700kb/s.

Frank Ogden

DTI launches campaign to kill the waves that kill

Manufacturers of almost every product with electrical or electronic components will have to meet strict limits on levels of electrical interference by the end of 1995.

This follows an EC directive on electromagnetic compatibility which became UK law in October 1992.

In the meantime, companies must either meet the regulations for each country with which they trade, or adopt the new directive in full and be able to trade freely throughout the EC.

The DTI has launched an awareness campaign to help companies get to grips with the new standards.

Edward Leigh, trade and technology minister, said at the launch of the campaign: “EMC is an environmental issue. The airwaves are rapidly becoming polluted with the spurious electromagnetic output proliferating from various electrical and electronics devices. The aim of the directive is to reduce this electromagnetic smog to a level which is acceptable so that the various communications, broadcast, and electronic control systems can co-exist and thereby not interfere with each other’s legitimate operation.”

He gave examples of incidences where spurious signals had caused death, danger, and destruction.

In the UK interference caused a computer controlled crane to drop its load killing a worker. And in Japan interference caused robots to go out of control causing two deaths.

Other incidences include a portable radio causing a semi-submersible oil platform to move, mobile radios activating car locking systems, electric trains causing computers 5km away to malfunction, and cars travelling at 70mph having their anti-locking braking systems come on due to a radio transmitter five miles away.

The awareness campaign includes an EMC helpline (061-954 0954), special journals and reports, a workbook for seminars and tutorials, and EMC clubs.

Trio plans to halve MMIC costs

The University of Kent, Philips Microwave, and Barnard Systems have joined forces to develop computer-based design tools to halve development costs of very high frequency GaAs MMICs.

The DTI is putting £1,437,000 into the project which plans to create a package including accurate models of MMIC components and all the elements needed in the design process, all built into a single workstation.

Adam Jastrzebski, a senior lecturer at Kent University, said: “At present, the combination of process speed and insufficient accuracy of computer simulation often makes it necessary to repeat the design loop two or even three times before the chip is constructed. The partners in the project want to develop the software tools which would guarantee the correct design right first time.”
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VISA

(VISI NO. TO ON REPLY CARD)
A novel method for generating clean millimetre wave signals from optical sources, has been developed by a team from the University of Wales in Bangor and the Alcatel-SEL Research Center in Stuttgart (Electronics Letters, Vol 28, No 25). The advantage is that it becomes possible to send light long distances through glass fibres - something that can not be done directly with millimetre waves!

Traditionally, delivery of RF signals to remote locations has been by amplitude modulating a CW laser with the RF signal. Demodulation at the far end then reconstitutes the RF. Unfortunately, bandwidth of available modulators places an upper limit on the frequency that can be regenerated in this way.

An alternative method, in theory, would be to send two coherent CW laser beams along the same optic fibre and then mix them at the far end to obtain millimetre-wave RF by heterodyning: the necessary stability can certainly be achieved by frequency-locking the two lasers to the required frequency separation.

Unfortunately the phase noise can not be locked. So, even with the best narrow line-width lasers, the resulting RF signal has a line-width of several tens of kHz.

The ingenious solution developed by the Welsh and German group makes use of a single laser to generate the two optical components. A lithium niobate Mach-Zehnder modulator - carefully biased to suppress the basic laser frequency - is fed with an RF sine wave at 18GHz, generating two optical carriers, each offset by 18GHz. When these optical signals are recombined at a distant point in a pin diode detector they generate a clean RF signal at 36GHz.

Secret of success lies in the fact that the two optical components are generated by the same laser and so have correlated noise components. When the beams are mixed together, these components cancel out leaving an RF signal, the line-width of which is no greater than that of the original RF signal used to drive the modulator. Standard DFB lasers can therefore be used, even if their intrinsic line-widths extend over a several MHz.

Benefits of the technique do not stop there. Amplification of the optical signals in an erbium-doped fibre amplifier (EDFA) make it possible to regenerate a millimetre-wave RF signal at a relatively long distance from its source. The experimenters tried sending the optical components along 8km. of fibre with no noticeable degradation to the line-width of the reconstituted RF.

In their recent paper, the group says that the system should find ready application in future picocellular radio systems or anywhere where it is necessary to generate small amounts of millimetre wave RF in awkward places.

Electrical spectrum observed after the EDFA and the 8km of fibre.

---

**Apparatus used at Bangor for experimenting with optical generation of mm waves.** Left to right is the 18GHz signal source, a light-wave component analyser used for the DFB laser source, spectrum analyser use to monitor the output signal of the fast PIN diode. At the extreme right is the 8km drum of fibre, next to the erbium doped fibre amplifier.
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Wired up head gasket reveals combustion picture

An instrumented head gasket for testing combustion in production engines has been developed at Sandia National Laboratories' Combustion Research Facility in Livermore, California. Its purpose is to provide detailed information that could lead to ways of improving the combustion process, reducing engine knock (spontaneous pre-ignition) and lowering the emission of unburnt hydrocarbons.

Sandia's gasket consists of a multi-layer printed circuit board whose wiring pattern is a circular array of ionisation probes. The probes detect the instant the flame front — the area of burning gases in the engine — passes. Although ionisation probes have been routinely used in combustion research for many years, this is the first non-intrusive system that requires no access through specially machined ports — allowing it to be used on production engines.

So far it has been successfully tested at the General Motors V-6 Powertrain Division test facilities in Flint, Michigan and is now being manufactured privately on a small scale.

The board comprising the gasket is made of a temperature-resistant glass-reinforced polyimide material that remains rigid up to 270°C. Wiring is etched using standard printed circuit technology.

Combustion in an engine cylinder, beginning with the spark, evolves as a burning flame surface that propagates radially outwards. The high temperatures ionise the gases which, because of their electrical conductivity, clearly mark the edge of the advancing flame front.

Ionisation probes built into the instrumented gasket are merely exposed conductors carrying a voltage. As soon as a flame contacts the probe, a distinct signal is produced.

Information obtained is critical in optimising factors such as position of the spark plug, design of the intake port and shape of the combustion chamber. It can also determine the direction and magnitude of the "swirl" motion of the gases. The motion is important in enhancing combustion rates, but is hard to observe without special viewing ports in the cylinder.

The Sandia researchers say their new device will provide information on the latest boogey of the environmental movement — unburnt hydrocarbons. One of the major reasons why unburnt fuel is expelled from an engine is that it gets lodged in the crevice between the piston and the cylinder wall where it is shielded from the advancing flame. Experiments using multipoint ignition at the perimeter of the combustion chamber appear almost to have eliminated the problem of unburnt hydrocarbons.

Perhaps the most technically ingenious use of the new instrumented gasket is to locate the source of engine knock. Knock occurs when unburnt gases self-ignite ahead of the main flame-front, often triggered by some local hotspot in the combustion chamber. If designers can discover precisely where this takes place, they can set about improving cooling of the head in the relevant area. The new instrumented gasket, because it contains numerous sensors, can actually triangulate the source of engine knock by comparing the timing of the pressure oscillations accompanying knocking.

Making chips at home (all you need is an SEM)

Researchers have demonstrated a simple room-temperature procedure for making a phototransistor, obviating some of the normal fabrication processes that require high temperatures or high energies. The technique, developed at the Weizmann Institute in Rehovot, Israel, requires a powerful electric field to be applied across a homogeneous piece of semiconductor. The result is permanent distortion or change in the crystal lattice giving the affected parts conventional semiconductor properties. No doping or heat-curing stages are needed, avoiding the usual complexities and dangers of contamination or overheating the wafer.

So far the method has only been applied to copper indium selenide, and to the creation of a two-terminal phototransistor. But applications to silicon are planned though the technique has not yet been refined enough to be used to create micron-sized components. The team caution that even if it does prove practical for commercial devices, many years of research lie ahead.

Using electric fields to manipulate charges in semiconductors is not a new idea. But no-one has previously suspected that by applying such fields at room temperature, sufficiently localised, sharply defined and stable charge distributions demonstrating transistor action could be demonstrated. The detailed physical mechanisms underlying this unusual low-temperature charge reorganisation are currently being investigated.

Normally, semiconductors are made by using chemical doping to tailor the electronic properties of components. The dopants are added by a variety of high temperature processes involving diffusion, ion implantation or layer-by-layer vacuum deposition. Dr David Cahen and his colleagues in Israel have used electric fields to break down the uniform space distribution of semiconductor ions and hence avoid the need to introduce foreign dopant ions. Stable charge distributions formed in the material resemble those obtained with normal doping techniques in which unbalanced positive and negative charges exist in close proximity. The team believe the charge reorganisation effect may be due to a combination of ion migration and the formation of physical crystal defects resulting from the transient high voltages used to create the electric fields or localised heating. A P-type or N-type semiconductor is the result. The hope now is that the initial results will stimulate interdisciplinary investigations needed to clarify the effect of large electric fields on semiconductors.

Researchers at Weizmann Institute of Science, Department of Materials and Interfaces, have demonstrated a simple room-temperature procedure for making a phototransistor.
Every which way – but not loose

What, according to the University of Michigan: "moves forwards, backwards, sideways, spins on a dime, never gets lost, and looks like a coffee-table on wheels?" As you might guess it's a robotic vehicle, and inventor Johann Borenstein claims it is more manoeuvrable, more stable and more reliable than its competitors.

The robot has been designed to carry materials or conduct inspections in hostile environments, such as inside nuclear reactors, and can negotiate tight turns more easily than traditional mobile robots with limited degrees of freedom.

Multi-degree-of-freedom (mdof) vehicles are notoriously difficult to control – Borenstein describes it as trying to control a car with four steerable wheels. The moment the steering loses coordination, wheels begin to slip and control is lost.

Main problem is that because mechanical components can not react as fast as the computer commands, it is impossible to eliminate all errors in speed and direction. Any practical motor system will always undershoot or overshoot the desired movement.

But accurate motion is vital where a robot is exploring and mapping an unknown area. So far this has only been achievable if a human is watching the vehicle from a remote location and applying the necessary correction.

To solve the slippage problem, Borenstein has developed a technology which he calls "compliant linkage". It uses two movable chassis, each of which has a set of wheels and sensors linked to a central computer. Whenever the distance between the two chassis deviates from a preset value, the second chassis compensates immediately by sliding forwards or backwards on a track beneath the vehicle. Feedback sensors detect this movement and the computer then directs corrective action.

Compliant linkage makes it possible to design mdof robots with the levels of slippage found in conventional robots that have only limited manoeuvrability.

Borenstein believes that a development of this technique will make it possible to eliminate slippage completely for extended operation.

The dual chassis system also means that the robot should be able to rescue itself if it gets struck or if a component fails while manoeuvring in some hostile environment, such as a nuclear waste dump.

Sound reasons for short term memory

New York University scientists have announced that they have located the area in the brain that stores short-term memory of sounds. In a recent paper they describe how superconducting detectors have made it possible to monitor the extremely weak external magnetic fields that accompany the firing of nerve cells – neurons – in particular parts of the brain.

The team has also established, for the first time, a direct relationship between brain activity and the way these memories decay. In principle, the studies could help locate the source of memories for touch or sight, and the work has made it possible to measure, objectively, the lifetime of auditory memory. Initial indications are that the length of short-term auditory memory differs greatly from one person to another.

In a recent paper, (Science, Vol 258, 1993), Zhong-Lin Lu, Samuel J. Williamson and Lloyd Kaufman describe an experiment in which each subject's judgement of the loudness of a "memory" tone, heard in a series of tones, was found over the course of a few seconds to decay from its correct value to the average loudness of all the tones recently heard. Auditory memory for the loudness of a specific tone appears to decay exponentially from its true value to the average of other sounds heard in context.

What is particularly interesting about this latest finding is that the researchers have correlated their practical observations about auditory memory with direct physical measurements of the relevant brain activity. Their technique, called magnetoencephalography, allows the team to monitor brain activity as the subjects responded to the musical tones.

What these magnetic measurements have revealed is that specific auditory memory--and its decay time--are all contained within the primary cortex area of the brain, about five centimetres above each ear.

Psychologists have long known that the brain retains information on the physical characteristics of stimuli for only a few seconds. For hearing and sight these are known respectively as "echoic" and "iconic" memories. Such temporary buffer-type memories form the basis for subsequent processing and longer term storage. What is new is that the New York researchers have, for the first time, measured brain activity associated with very short-term memory.

Volunteers participated in experiments to determine how accurately they could recall the loudness of a tone by comparing their memory of it with a "probe" tone played a second or more later. The subjects pressed one computer key if the probe tone seemed louder than the memory tone and another key if it seemed softer. Many thousands of measurements were made in to obtain meaningful statistics.

The most remarkable finding of all was that echoic memory seems to vary from less than one second in some subjects to more than three seconds in the case of others. This is an extremely wide spread compared with most other physiological variables, such as, say, reaction time. It is not easy, though, to determine whether wide variation in echoic memory has any practical significance.

Sensory memories, in general, are lost very quickly, after which we draw on a longer-term global experience that recalls only average levels in relation to other ambient factors. That is one reason why we don’t generally notice the effects of an audio compressor that operates slowly and gently. Or why turning up the volume does not greatly change the character of a voice.

Research Notes is written by John Wilson of the BBC World Service

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Anyone who claims competence as an electronic designer must be able to apply the design basics of digital signal processing. As with all things the way to learn is to try it and see. Jean-Jacques Dauchot puts together an experimenter's kit for DSP

The dominant development in electronics during the 1970s was the microprocessor chip. It revolutionised microelectronics. It quickly replaced circuits which used hard wired digital logic and before long, low cost development systems became available.

As the analogue world was transformed into the digital domain, a new type of microprocessor was introduced, the digital signal processor. DSP chips are microprocessors which use special instruction sets specially geared to signal processing.

DSP chips have now invaded modems, radios, medical instruments, music instruments and many other products. DSP sales in 1991 exceeded two billion pounds yet DSP technology still seems a mystery to many engineers.

The greatest hurdle to many engineers is the perceived difficulty of using DSPs. One way of getting grips in the use of DSP is to obtain a DSP chip set and try to use it. Another way is to obtain a DSP development system and software tools from the intended manufacturer, but they are very expensive.

DSP applies complicated mathematical methods such as filter functions to achieve the processing of the signal. It requires high degree of mathematical knowledge to be able to convert a filter transfer function to a set of instructions for the DSP chip to process. DSP manufacturers are now trying to redress this situation by providing substantive software and hardware support to encourage engineers to develop systems using DSP.

Software packages are available which generate DSP assembler code from a given set of filter parameters. One such package is the Atlanta Signal Processors Inc. DFDP (Digital Filter Design Program). There are also numerous third party text books describing digital signal processing.

The developer

The developer is based on the TMS320C10 running at 20MHz. An 80C31 microcontroller is used as a master processor to the DSP, its serial port is used to connect to an IBM PC. A software package on the PC communicates to the DSP developer which allows DSP programs to be download to the DSP program memory.

The TMS320C10 is a first generation DSP chip which was introduced in 1983 by TI. It is an old design but for the purpose of this exercise, the device is perfectly adequate for a demonstration of what DSP can be used for, and it is easy to use.

The DSP developer is constructed on an extended single eurocard with all the C10 data, control and input/output signals connected to the 64-way DIN connector. The decision to construct the DSP developer on a eurocard system is to allow easy access to examine waveform and signals with the use of an oscilloscope.

No ADC/DAC circuitry is available on the main DSP board. This allows various types of analogue interface boards to be connected to the DSP developer via a backplane. The board also includes a three channel timer chip controlled by the 80C31 CPU which can be used for anti-aliasing, sampling rate, and reconstruction filtering. The three timer channel outputs are connected to the DIN connector. A backplane is available to connect the DSP board to analogue/digital interface cards complete with 64-way DIN connectors.

The main DSP developer board must be interfaced to an analogue board via a 64-way back-plane.
This figure shows the main simulator screen. The large window across the middle of the screen is the memory window. The light blue horizontal cursor shows the location of the program counter. The green cursor is used to set breakpoints. In this example, a tracepoint has forced the simulator to halt.

This figure above shows the simulator graph layout window popped up over the main screen. Up to four traces may be plotted, only one has been defined here. Main memory, data memory, input or output ports may be graphed with any start position and trace length. Number representation may be either signed or unsigned. Bar or line graphs may be plotted and the Y-axis may be either log or linear.

There are three timer channels, under the control of the 80C31 CPU, available on the developer to provide sampling rate, anti-aliasing and reconstruction filtering. DEVCOMMS provides the facility to set, start and stop any of the timers.

The block diagram of the developer shows that low cost DSP development system can be acheived with minimal chip count. The use of the 80C31 microcontroller as the master controller provides high intergation and interfacing to the PC via a serial link allows user flexibility to experiment without the risk of damage to the PC.
The 80C31 CPU is the master controller of the developer. Its functions are controlled by the DEVCOMMS package via the PC serial link. The 80C31 has a built-in serial port and the RS232 signals are supplied by the MAX232 IC.
Circuit description

The 320C10 and the address/control multiplexers (U17, U16, U15, and U10) are controlled by setting port P1.5 of the 80C31 high or low. When high, the reset line of the C10 and pin 1 of the four multiplexer ic's are held low via the inverter U06. In this state, the C10 is in its reset mode, its data lines are tristated and the DSP program memory address lines are placed under the control of the 80C31 processor.

The 80C31 has full access to the DSP program memory which can be loaded with a program or contents read. When P1.5 is set low, the TMS320C10 reset line goes high and the control of the address/control lines is returned to the DSP. The DSP then starts running its program from address 0.

Data is written/read to the DSP program memory in byte size via the two bi-directional buffers (U6 and U7). Port P1.7 and the inverter (IC4) controls which byte of the word is accessed. When P1.7 is set high, the MSB can be accessed, when set low the LSB can be accessed.

The TMS320C10 can address up to 4K words of external program memory. The circuit utilises 8Kx8 memory ic's. The 80C31 port P1.6 is connected to A12 pin of U8 and U9 which allows two DSP programs to be loaded at any one time. Toggling P1.6 allows switching between the 4K word pages determining which program the DSP should run.

The TMS320C10 can be replaced with a TMS320C15 DSP; P1.4 port is programmed to set pin 3 of the DSP chip to run either in microcontroller or microprocessor modes.

The 82C53 timer clock is derived from the 5MHz clock from the 320C10 CLKOUT pin. P1.0 to P1.2 output ports of the 80C31 controls the timer outputs.

Input/output

The 320C10 has two instructions which perform input and output operations and can read and write to eight port addresses. The IN instruction reads data from a peripheral port and places it in data memory and the OUT instruction transfers data from data memory to an external peripheral port.

Examining the timing diagrams for the OUT instruction and TBLW instruction, which writes data to program memory, the /WR line is active during data transfer while the /MEM line remains high. If i/o addressing is not fully decoded, any OUT instruction could write over the first eight program memory locations. Any TBLW instructions would also activate the /WR line on the i/o address ports.

The circuit comprising of U13, U14 and U10 fully decodes i/o addressing. It prevents TBLW instruction activating the i/o ports and /WR line while addressing program memory from location eight onwards. It also blocks the /WR line to the program memory while executing the OUT instruction. The only restriction here is that no TBLW instructions should be attempted which involve the first eight locations of program memory, it will overwrite the program instructions at those locations.

Normally, prom would be used as program memory so that the problem of writing to program memory while executing OUT instructions would not exist. Examples of TI's own circuits uses 2K of prom for the lower part of program memory and 2K sram for the upper part. However, full i/o decoding must still be used to prevent write access to the ports.

The output of U13, U14 and U10 fully decodes i/o addressing. It prevents TBLW instruction activating the i/o ports and /WR line while addressing program memory from location eight onwards. This signal must be used together with PAO-PA2 signals to decode i/o on any boards plugged onto the bus.

Note that no buffering is used on the DSP data and control lines' main board, therefore suitable buffering on the interface board must be employed to prevent any damage to the DSP developer.

The software

The developer comes with two software packages, the IBM PC DEVCOMMS developer interface and a real time protocol engine in an eprom plugged into the DSP developer board.

The PC Developer interface software is a set of windows and pop down menus which allows the user to control the developer via the serial interface using simple commands. The command set allows TMS320C10 programs to be downloaded to the developer, contents of DSP program memory to be uploaded to the PC to be examined, setting timer frequencies, etc.

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The run time kernel monitors the activity of the DSP developer and update a status window showing various states of the developer.

**Software development tools**

The software package used to develop the programs for the TMS32C10 is the Hippo Solutions TMS3201X development tools. Together with the Developer, it offers a complete and low cost development system for the processor chip.

The software system consists of a compiler, assembler, linker and simulator. Broadly speaking, the compiler translates C language subset source into assembly language, these sources are combined with other assembly language sources, assembled and linked into a single executable image which can be either split into high and low byte images with the provided utility UIX or loaded into the simulator or downloaded into the DSP Developer to be run.

The compiler allows code to be developed in a high-level-language, an integer only subset of the C language, which takes a lot of the pain out of DSP code development. Special provision is made for particular number representations with the availability of Q15 and Q14 multiply operations from high level source. In addition to this, the compiler allows a very close machine interface with the provision of NAND, NOR and XNOR operators.

The output of the compiler and associated optimiser is assembly source. The input to the assembler, whether from the compiler, user supplied assembly source or from one of the applications examples, must adhere to the format in the manual which regrettably makes it incompatible with the TI convention in some ways. Hopefully however, these deviations should not prove prohibitive to useful application of the provided tools.

The compiler and the assembler produce object files which are combined with the supplied linker to produce a single executable

**The output current of the DAC-800 is converted to a voltage by U2. The voltage is fed into a MAX292 8th order low pass switched-capacitor filter IC. The internal op amp of the MAX292 is used to provide a 2nd order low pass filter to remove the clocking frequency used to clock the filter.**
ties, the user is recommended to make full use of the simulator's debug potential. Using the simulator it is possible to set breakpoints, set global trace points, examine memory in a variety of formats, mark regions of memory as being out of bounds, assign files to processor input and output streams, simulate activity on processor interrupt and BIO pins and many other features too numerous to mention.

The simulator has a generalised user interface which means that anywhere numeric input is required, the user can enter complex C language type expressions except struct/union and post/pre increment/decrement operators. One of the most powerful facilities of the simulator is its graphic capability. It is possible to display up to four traces as graphic information where a trace may be either a port input stream, a port output stream, an area of program memory or an area of data memory. In all cases, graphs have a start address, a length, they may be signed or unsigned, they may be presented as either straight line graphs or bar graphs and they may be in either linear or log representation. Clearly, since signal processing tasks are concerned with the relationship between input and output in the majority of cases, the simulator's graphic capabilities are a formidable step in the direction towards simplifying DSP development.

**Experimenting with the developer**

On its own, the DSP developer board can do little. A suitable I/O board must be connected to the developer via the 64-way back plane. An eight bit ADC/DAC circuit is shown below. It uses the popular DAC8000 dac chip and the ADC7575 ADC chip.

Anti-aliasing filtering is accomplished with a 2nd order low pass filter and an 8th order clock tunable low pass filter using a MAX291 chip from Maxim. The clocking frequency for the filter is provided by the 82C53 timer channel 0 under the control of the 80C31 controller. The clocking frequency must be a hundred times that of the highest frequency being sampled. Sampling is initiated by reading the ADC which loads the contents of the previous sample and starts the next sampling process. The device has a built in sample and hold circuit. The sampling period can be determined by the use of the 320CIO's BIO pin connected to the output of the monostable. The 82C53 timer 3 is programmed to trigger the monostable.

The output of the dac is connected to another MAX291 filter circuit which reconstructs the signal and filters out the sampling frequency. The conventional 2nd order filter circuit removes any residue clocking signal. The 82C53 timer 0 provides the clock for the switching capacitor filter.

A simple test of a complete system can be done by loading the following program to the TMS320E15. Finally, the code and symbol files can be loaded into the simulator to give symbolic debug information.

The output of the linker consists mainly of a COD file which can be used in one of four ways. Firstly it can be used to directly program a TMS320E15. Secondly it can be used with the supplied utility U1X to program two banks of eproms with the supplied utility U1X to program two banks of eproms. Thirdly it can be used to directly program the DDRAM of the DSP developer. Finally, the code and symbol files can be loaded into the simulator for program timing and verification.

The simulator is the flagship of the available software support. Because the DSP Developer provides no run time execution debug facilities, the user is recommended to make full use of the simulator's debug potential. Using the simulator it is possible to set breakpoints, set global trace points, examine memory in a variety of formats, mark regions of memory as being out of bounds, assign files to processor input and output streams, simulate activity on processor interrupt and BIO pins and many other features too numerous to mention.

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A simple test of a complete system can be done by loading the following program to the DSP developer.
It waits for the BIO pin to go low, then reads the ADC and delivers it to the DAC. The timer outputs must be programmed to deliver suitable square waves to drive the tunable filter and to provide the sampling frequency. Use the menu option to enter the counter divider values. Timer 1 and 2 should be set to 2 (2.5MHz) and timer 3 to 113 (approximately 44kHz). Connect a signal generator to the input of the analogue board and an oscilloscope to the output. Set the generator to deliver a sinewave of less than 20kHz, 1V p-p. You should see a reconstructed sinewave on the oscilloscope. By changing the timer values, serious cases of aliasing can be demonstrated.

TMS32010 digital filtering

One of the most common applications of DSP microprocessors is that of digital filtering. There are a number of important advantages which digital filters have over their more familiar analogue counter parts. For instance, the filter response is independent of component tolerances, temperature instabilities and ageing, ensuring the filter response is exactly reproducible from one system to another. It can be guaranteed that a filter has a phase response which is linear. This means that the filter output is not distorted due to unequal phase delays for the different frequency components which make up the input signal. Also the filter passband flatness and roll-off characteristics can be far superior than any realisable analogue filter.

The FIR filter

There are a number of algorithms which can be used to implement filter functions. Of these the finite impulse response filter guarantees linear phase, is generally the easiest to design, and is the most robust with regard to numerical stability.

The FIR filter has the form shown below. It comprises a digital delay line through which successive samples from the input source (e.g., ADC) are shifted. At each delay stage a multiplier weights the delayed sample by some multiplier weight. The output from a differentiator is the gradient of the input signal. In the sampled data domain this gradient can be calculated from the given sample rate the gradient can be approximated as the gradient of a straight line connecting successive samples. For this simple differentiator the coefficients would be 

\[ \{1, -1\} \]

Using 16-bit fractional two’s complement integer representation, the coefficients would be \[ \{-32767, 32767\} \] i.e. \[ 2^{15} \]. The scaling by \( f_i \) would be accomplished by amplifying the dac output appropriately if required. Better differentiators can be designed using more sophisticated techniques but this will suffice for this example.

Using these coefficients and 16-bit arithmetic the filter output could overflow causing distortion of the output signal as the accumulated results wrap around the full scale number representation –32768 and 32767 in this case. There are a number of approaches which can be taken to this problem:

1. Assume that the input sequence which could cause overflow is improbable and make no allowance for its occurrence.
2. Ensure that overflow will never occur by scaling all the coefficients by a fixed factor. For 16-bit arithmetic this factor S can be determined from \[ S = 32767/\text{sum from } i=0 \text{ to } \text{NAbs(a(i))} \]
3. Use the overflow mode of the DSP processor (SOVM in the TMS32010) such that, if an overflow does occur, instead of wrapping around the arithmetic will saturate to its max-
FIR Implementation

The kernel code for the implementation of a TMS3210 FIR filter is shown.

The data memory is organised so that the delayed sample array appears as the first element, so that auxiliary register AR0 can perform two roles, a decrementing counter and to access data samples from memory. Auxiliary register AR1 is used to reference the coefficient array memory locations.

Both registers are initialised to start at the end of their respective arrays at the oldest filter sample.

The accumulator is cleared ready for all the partial products from each of the multiplies to be summed. The last partial product is calculated first and its result left in the product register for its corresponding filter coefficient and the delay line by shifting everything by one sample.

The sample delay line value is multiplied by the next sample delay line value; this current delay line value is copied into the next highest data memory location. (see also the DMOV instruction) This effectively clocks the delay line by shifting everything by one sample as the filter calculation proceeds.

The sample delay line value is multiplied by its corresponding filter coefficient and the result left pending in the product register for the next execution of the loop and ltd instruction.

The banz instruction branches if the current auxiliary register is not zero and decrements its value by one. This is a common method of implementing a counting loop in the TMS3210.

Note the final apac instruction adds the last remaining pending product from the P-register, since when the loop breaks, the ltd instruction will not be executed again to perform this task. The final filter output value now resides in the accumulator and can be output to a dac, etc.

Direct digital synthesis on the TMS3210

Traditionally sinewaves are generated by analogue techniques. It is now possible to generate low distortion sinewaves digitally with frequency stability and accuracy derived from a crystal oscillator. This technique is known as direct digital synthesis. There are available, a number of special purpose DDS chips and modules. However, they are expensive and are still confined to professional and military equipment.

The system described here is a low cost route to DDS. The software will run on the TMS320 Developer with the corresponding ADC/DAC card for analogue output. DDS in software, even on a DSP micro, will never compete with dedicated hardware in terms of speed. The system described below will generate sinewaves from 5mHz to 50kHz in steps of 5mHz.

The direct table lookup method is the simple and quick. Readers wanting a fuller exposition of the theory are directed to Tierney, 1971. The method requires that one stores a number of uniformly spaced samples from one cycle of a sinewave in a lookup table. If we now step sequentially through this lookup table reading out successive samples to a DAC at a constant output rate then we get one cycle of a sinewave. At the end of the table it folds back to the start and continues to generate a sinewave indefinitely. One can calculate the maximum frequency from the generator if the time taken to go round the software loop once is known. Assuming you have a 20MHz crystal in your TMS3210 board then, using the code supplied with the kit, the maximum frequency is 50kHz. The corresponding minimum frequency is 5mHz. Such low frequencies are difficult to generate by analogue means.

Direct Digital Synthesis of the Sine Program

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I would like to thank Dr James Dripps and Dr Keith Manning for contributing to the article.

March 1993 ELECTRONICS WORLD + WIRELESS WORLD 195
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PILGRIM WONKS (Dept. WW), Stairbridge Lane, Pilgrim, Worcestershire WR17 6PA.
Signal processing software package DADiSP works like a spreadsheet, except that the rows and columns of numbers are replaced by an array of graphs which can be interdependent. A change made to one graph (or data set) causes a corresponding ripple through the whole set of graphs.

Application areas span all of engineering, and since the package has been designed for the needs of data processing and representation, the range of options within the package is immense – appreciating its full capability is quite daunting.

Version 3.0 adds some major extensions to the package. But it is not all good news. The inclusion of a parallel port dongle is a real pain, there because of massive copyright infringements. In one Mediterranean country the software became one of the most popular signal processing packages, yet only a handful of legitimate copies of Version 2 were sold there.

The dongle is not the only sticking point. Users accustomed to running their PC with Windows 3.1 will find a conflict with the memory managers. In fact the user manual devotes an entire appendix to likely problem areas – a not very reassuring move for the third version of a well known product.

More facilities
The range of functions has been greatly expanded over earlier versions and more control is now possible on each graphical display. Axis labelling, scaling and grid features are all improved, and 3-D plots are supported and generated. Once a 3-D plot has been created a contour plot of it can also be generated in another window. Another attractive feature is the ability to create X-Y (or polar) plots. Data can be displayed in a table format and edited at will, and can be imported from a traditional spreadsheet and then subjected to the impressive range of statistical tools available. As in previous releases, Version 3.0 MACRO helps automate repetitive operations (whether standard or user defined) by compiling them into an executable array (a macro). In this way a user can customise DADiSP to perform a sequence of specific tasks.

Matrix operations options, for processing arrays of data are included with Version 3.0, some accessed through the pull-down menus. Matrices can be displayed in a number of ways, for example, as density plots where the matrix ele-
ments are given different colours: or represented by a 3-D or contour display.

Two macros supplied have lots of additional matrix operations sometimes encountered in matrix maths (column and row statistics, extracting triangular areas). Although many may never be used, they do illustrate the comprehensive design of DADiSP.

Data acquisition
In addition to data and signal processing, a package such as DADiSP could not be complete without data acquisition. Strangely, the IEEE-488 control and expansion board control are options, and since neither was supplied for review, I can only guess at their capabilities.
DADiSP-488 option allows control of IEEE-488 based instruments through the commands contained in the IEEE standard. The data acquisition option apparently allows DADiSP to read data directed through recognised, commercial expansion cards into a graphic display for further processing. So DADiSP can take on the role of an oscilloscope with several channels.

A-version
Documentation can be described in a single word - poor. The reference manual is written like a dictionary, and though it might be useful to old hands, it will leave the first time user none the wiser. The user manual is slightly more illuminating - but not much. It serves as a useful introduction to the package, but not a lot more.
No mention is made of data acquisition or the IEEE-488 options which is rather strange since these features will be of great importance to many users. Overall, the earlier versions of DADiSP were excellent products, and the enhancements found in Version 3.0 do not seem to add greatly to its functionality.
That said, DADiSP still remains one of the most flexible and comprehensive signal data processing packages on the market.

References
1. DSP Design with DADiSP, EW + WW, Dec 1989, p1151

Supplier details update
A new version has been released to tackle some of the package's previous shortcomings, particularly the Windows incompatibility problem. DADiSP/PRO-32 conforms to the VCPI and DPMI standards and now operates under Windows 3.0 and 3.1.
Up to 64MByte of extended memory can be addressed. Hardcopy has been improved, says Adept, and now includes higher resolution and better quality. On screen the speed of image display is said to be much faster.
Rotation of 3D plots is now mouse driven.

DADiSP/32 is £1295 ex VAT. Adept Scientific, 6 Business Centre, West Avenue One, Letchworth, Hertfordshire.
Tel: 0462-480055

Imaginary image processing
Image processing is mentioned as an application for the software. But I find it difficult to believe that DADiSP Version 3.0 could possibly serve as a suitable platform for performing image processing tasks. It does not recognise standard image file formats and has no well-known image processing operations within it array of menus. Until these two deficiencies are addressed DADiSP is not a viable image processing tool.

3-D, contour and X-Y plots can be generated.
Graphical demonstration of scientific power

Extensive software is available to help tap into the power of a PC's mathematical functions - the ubiquitous spreadsheet comes readily to mind. But packages are largely commercially oriented. Programs designed for the needs of scientists are not only more difficult to find but often expensive.

So it is pleasing to discover the powerful CC4 package, targeted specifically at scientists and engineers, which transforms the PC into an excellent scientific calculator, with good graphing options and comprehensive programming facilities. Files can be output to disk or graphic screens to a printer, and all functions are entered as they are written. With other custom procedures created in a Pascal-like environment. The considerable power derives mainly from the enormous range of commands that can be used to program not only the calculator but embellish its graphical output.

The calculator presents
A basic three window environment is presented to the user, for information, data entry and manipulation, and output. At the top of the screen is a summary of function key assignments, and the screen is vertically divided by the input and output windows. Expressions are entered into the input screen, as are additional commands needed for text enhancement of graphic control, and any amount of extra explanatory text. The beginning of each text line is marked with '/' to indicate that the calculator should not attempt to evaluate the remainder of the line, though an expression on the same line, placed before the marker, can be processed.

The output window is used to report the computed result of expressions submitted to the calculator. If syntax of an entered expression is incorrect, CC4 puts up an explanation box, removable with the esc key, calling for another attempt. The program's default insert mode can be switched to overwrite, and there are facilities for editing programming and explanatory text by marking, inserting lines and so on. At the bottom of the screen a rolling tally shows current memory available as well as the last result and the last graph held in memory.

Editing facilities include cut and paste for moving or duplicating entries, using F3 to mark the text and F4 to add text from the buffer to the cursor position - repeatedly if necessary. The entire entry at the cursor can be selected so that pressing enter or del either selects, or cuts and selects, the entry.

Ten buffers may be used for saving text, with the latest selection recalled by Alt-1. Each time a new selection is made, the previous selections are promoted through the
buffers and can be recalled using ALT-1 to ALT-9. It’s a nice touch – provided you can remember the order. Memory consumed by the buffers can be regained using SHIFT-F4.

Amongst the default assumptions, radians are used rather than degrees – unless specified otherwise with the DEGREE command – and the current mode is indicated as a reminder in the top window. Various line styles, thicknesses, and colours can be summoned as alternatives to the default settings, and the location of the help file and also the last graph file and printer type can be set in a configuration file.

Where more than one printer is on a system, it is best to let the program throw up a prompt each time for the printer type in use. Simply enter E for an Epson compatible, I for an IBM printer, H for HP Laser, or N for “None of These”. The configuration file can save time if hardware is fixed.

**Entering data**

Numeric values are entered into the calculator’s input window in the usual way. Scientific notation can be used, and functions for SQRT, ABS, RANDOM, INT, FRAC, ROOT, LN, GAMMA, BINOM all have the usual meanings. It can be entered as either PI or ALT-P, and trigonometrical functions require the argument to be enclosed in brackets. Brackets can also be used to determine calculation precedence.

Extensive notes on entering the logarithmic base, integers and non-integers (the former can be combined bit-wise), hyperbolic functions, hex or binary notation, with inter-conversion functions can be found in the help file. Complex numbers can be used and any of the program’s functions which are mathematically defined for complex values, such as sin(x) or exp(x), will accept complex arguments. Huge numbers are indicated by the prefix “&”. To store an exact action, two exact parts can be input; for example &2/3 is stored as the fraction and not 0.666666667. Again, detailed help notes indicate the way these numbers, and matrices involving them, are handled.

**Editing and controlling input**

A carriage return will force CC4 to compute an expression while the cursor is on any part of the appropriate line, or an entry may be left for later activation by using the up/down cursor keys.

Formulas already evaluated are displayed as normal text, while unevaluated expressions, including the one currently being entered, are displayed in bold or in a chosen colour. Complex expressions are not limited by line length as the entry can just flow to subsequent lines.

Any existing expression can be run repeatedly, so the effect of modifications can readily be seen, without the need for re-entry. In addition, the calculator is not like a spreadsheet where the spatial order is significant. Output depends on the order expressions are used and not their entry order. CTRL/ENTER opens up as many blank lines as required for additional entries at the cursor position, so compiling a program of entries is quite easy. To control precision of a calculation, PRECISION(N) can be used, where (n) is a number between 1 and 9. Using 0 causes the result to be displayed in scientific notation.

Results of expressions can be stored in case-insensitive variable names up to twelve characters long, and variables are all conveniently listed in the output window which, when full, can be scrolled with function keys F5 and F6.

A defined variable can be reused in another expression, and variables can be redefined at any point. Useful for recovering the memory assigned to a variable is FORGET(X) which destroys the variable X in memory. Greek letters can be used as variables by making appropriate alt/letter entries in the usual way.

Custom procedures and subroutine functions can use all standard control structures with value and function name parameters. Subroutines can call one another – whatever the order of input, and recursion is fully supported.

**Shareware strategy**

*Calculus Calculator CC4*, is the latest in a line of development versions of a programmable calculator from David Meredith of the Department of Mathematics at San Francisco State University. Its predecessor, *CC3*, is a standard commercial prodac, now published by Prentice Hall and called “CC - The Calculus Calculator”. The version reviewed here is, for the time being, offered as shareware – a try-before-buy category of software, available from specialist libraries for practically nothing. In this case registering your copy of *CC4* for support and updates is also completely free, though the commercial program, probably more extensively debugged and costing $30, comes complete with a 200 page manual.

Additions to previous versions, such as matrices and infinite precision arithmetic, are all summarised in an UPDATE.CC file on the shareware distribution disk as well as in the excellent on-line help file that comes with the program. This effectively replaces a user manual.

**Extensive facilities**

*CC4* can deal with basic arithmetic, logarithmic, trigonometric and exponential functions, real and complex number handling, polynomials, Boolean expressions, hexadecimal and binary numbers, exact fractions and huge integers, string handling, calculus, matrices, vectors, differentiation and integration, and statistical operations.

Eigenvalues and eigenvectors are also offered as well as equation solving, polar and parametric curve fitting, 2D graphing, and 3D surface plotting, and a huge range of options for p of presentation and enhancement.
On-line help

On-line help is comprehensive, freely interposing pseudo-hypertexted references to subordinate information and explanations, avoiding the tedium of too much detailed help.

Under broad headings such as "Capabilities of CC", "Help for individual commands", "Function keys and other special keys", "Printing", and "Disk and memory management", the help file fully describes the program. Its 84k is well-used disk space, but it can be omitted to operate from the much slower and not really recommended floppy disk. The EXE file itself is 210k but an overlay file of 242k is used to reduce the PC memory overhead to 512k.

Supplier details

CC4 shareware can be obtained for a disk copying fee of £4.65 from The Public Domain Software Library, Winscombe House, Beacon Road, Crowborough, Sussex TN6 1UL. Tel:08926 63298. No registration fee is requested by the author for this version.

Impressive graphics

Graphical modes are impressive. Two-dimensional graphing of expressions can be made in a window defined by WINDOW(a,b,c,d) where a, b, c and d are real numbers. Optional axes can be injected into the plot, and all the figures in a window can be deleted without rescaling the window, using the ERASE command. The four extremities of a window may be returned by specific commands – which is useful for programming.

The ability to include multiple graphs in a plot, controlling line colour and thickness, graphing selected points, varying the orientation of added text labels, size and type, point marking and unmarking, box and line drawing, selected fill with shading, variable axis colour all add to better presentation of plots. Point graphing commands include DOTGRAPH, QUICKG, and SKETCH for increasing the speed of screen drawing by selecting spaced points for inclusion.

Good control over the extent of a plot includes coordinate specification, as well as zoom and unzoom to particular locations. Automatic colour differentiation between the plots is assisted by KEEPLINE and varyline commands.

One other option worth a mention, is the crosshairs facility in 2D graphing. The crosshairs can be moved under cursor control, with a choice of two speeds and line thickness of placement, and the current coordinates are shown in the window. They are useful for locating the intersection of curves or points where plots cross axes.

RETURNING the last X and Y coordinates of the crosshairs, so for example the function y=f(x) has been graphed, equation f(x)=0 can be solved by using the crosshairs to find the point where the plot crosses the X axis. Returning to the input window and entering SOLVE(f(x)=0, x=RETURNING) will give a good estimate of the desired root.

Crosshairs can also be used to specify a zoom range. Any part of a plot can be outlined and subjected to a full screen redraw. CRTL+U will undo a zoom, and plots can be magnified or reduced in stages. Screen redraw is quite rapid, even on hardware of modest power.

3D graphing

Three dimensional graphing brings its own extensive range of commands. Broadly they are as in 2D graphing but without the crosshairs and zoom facilities. A 3D plot can be rotated on the screen about any of the three axes to view it from different angles. Entering the three axis letters x, y, and z rotates the image in one direction – repeated keystrokes will produce multiple jumps – and the corresponding capital letters reverse the direction of rotation. The image movement is not fast enough to be described as real time. But it is rapid, and can be made it significantly faster. Four elements of shading, which determine the effective resolution of the plot and therefore the time to redraw on screen, can be summoned with keys 1 to 4. 1 is the default setting for a high res image, while 4 is the simplest plot and the fastest to redraw. 4 also produces transparency.

Returning to the package in general, CC4 has a whole host of other facilities and functions that are far too many to mention here. Of course it may not be perfect, but any deficiencies can be forgiven in view of the author's generous marketing stance. The current shareware version is said to be a beta copy, and the author expects to publish "next year" so it might not be available for too much longer.

Buy while the going is good, and if you don't like it, it will have cost you only the disk copying fee of £4.65.
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Automatic generators for VGA sync

Low-priced colour computer monitors, often seen on the surplus market accept negative-going line sync, which makes them incompatible with the VGA standard used by many PCs. This circuit produces negative-going sync, whatever the polarity of the input.

Dividing the 5V line in the \( R_2, R_3 \) potential divider to 2.5V eliminates the need to amplify the input signal. A CR network \( R_C, C_1 \) on the input of an exclusive-Nor holds pin 2 high when a negative sync is applied, the pulses being too narrow for \( C_1 \) to discharge below the gate threshold: the syncs are therefore not inverted. If the input is positive-going, pin 2 is not allowed to rise and the input is inverted.

A buffer, an inverter and a BC337 current amplifier prepare the signal for combining with the vertical syncs.

Tom Sheppard
Swaythling
Southampton

Programmable timer

Three seconds to nearly six hours is the range of this crystal-controlled timer, at a resolution of 1s. In comparison with D Ibrahim's design in September 1990 Circuit Ideas, the time range is increased by a factor of 200, accuracy is better and fewer ICs are needed.

Four thumb-wheel switches set the time interval. Initially, the counter \( IC_2 \) is inhibited and preset, inputs to \( K_a,b,c \) being low. A start signal from the switch sets the D-type flip-flop \( IC_1 \) and the counter begins to count, the \( Q \) output of \( IC_2 \) resetting the flip-flop at time-out. During the count cycle, the LED illuminates and the relay is on.

The CD4013 generates 1Hz pulses at crystal accuracy, but any other type of pulse generator would be suitable.

V B Oleinik
Kaliningrad Moscow Region Russia

LC square-wave generator

As an alternative to the usual RC arrangement, an LC circuit offers better frequency stability. This Colpitts oscillator circuit provides a square wave, using an op-amp as the active element.

Any of the standard op-amps will work reasonably well at audio frequencies, trace (a) being the output of a 741 at 6kHz; trace (b) is an LM6361 at the same frequency. The third trace is an LM6361 at 1.3MHz. Output frequency is \( 1/(2\pi\sqrt{LC}) \); the two capacitors being effectively in series.

Michael A Covington
Athens
Georgia
USA

Oscilloscope traces of square-wave oscillator using an LC circuit for stability. Top two are at 6kHz, using a 741 (top) and an LM6361. Bottom is an LM6361 at 1.3MHz.
Fast, small-signal rectifier

For the rectification of signals of less than 100mV at frequencies above 100kHz, the common op-amp and two diode circuit suffers from distortion and low bandwidth. This circuit overcomes the problem. Here, the comparator detects signal polarity and connects Vin(t) in either normal or inverted form, via the Maxim 14516 switch, to the output. With a TL072 as the op-amp, and an LM360 as the comparator, bandwidth is zero to 500kHz with signals of 50mV-5V. A high-speed inverter would increase the operating frequency to 1MHz, since rise and fall times are less than 100ns.

B Jacobs, G Seehausen
Warner MM, Europe
Alsdorf
Germany

Adjustable, accurate stepper controller

In addition to providing precision, variable drive for a small stepping motor by means of a digital input, this circuit allows direction reversal. In Fig. 1, the input from a microcontroller or other digital source is taken to the 4059 divide-by-n counter driven by a crystal clock, which forms the pulse sequence. The 14516 up/down counter also takes its input from the digital source and feeds a binary output in the forward or reverse sequence dictated by IC2 and the direction input to the counter to the 4028 binary-to-decimal decoder, which drives the output transistors. Eight 2N3706's handle motors needing up to 200mA and the circuit in Fig. 1 will control motors with eight phase windings; for those with four, use the circuit in Fig. 2.

V B Oleinik
Kaliningrad
Russia

Fig. 1. Digital input controls stepping motors needing up to 200mA, accurately and reversibly.

Fig. 2. If the motor is an 4-phase type, modify the drive circuit like this.
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Traditional GPS radio architectures require access to ASIC technology for the hardware processing channels, but also suffer from long acquisition times. Philip Mattos describes a high performance radio system based on direct conversion techniques, and also outlines Plessey's single chip GPS implementation.

In 1988 when my original GPS designs were published in an IEE paper there was not, at that time, a need for a new radio architecture as the market price for GPS receivers was astronomical. However between 1990 and 1992 the market price fell from £2500 to £600 for the leisure marine GPS set, and with the manufacturer only getting a small fraction of this retail price, and with the radio section now the dominant cost element, something had to be done. The large GPS companies invested heavily in radio-frequency ASICs or mmics. Having done so, they were unlikely to release their chips onto the open market. The merchant market for radio chips did not respond for several years, probably waiting until GPS developed a mass market. Thus, after seeing the problems met by several new start-up companies in GPS, I decided to design my own radio from discrete components. Part of this article is dedicated to that design. However during 1992, both Plessey and Avantek released information on single chip products.

Quad-helix antennas come close to the ideal gain of 3dBiC over the entire hemisphere. The patch is cheaper and its +7dBiC overhead makes it useful on land where low angle satellites can be blocked by trees.

The Avantek ROC receiver on a chip is not yet available, just vague details in sales presentations, but I have tested the Plessey GP3010 single chip GPS radio so this will be discussed here along with my own design.

The hardware differences between my design and the conventional fall into two areas. In the radio, a direct down conversion approach is used, from L-band to baseband. Besides requiring only one local oscillator signal, and one mixer, it additionally means there are no image problems to overcome, and thus no spurious responses. Local oscillator filtering can be done with standard GPS frequency filters available off the shelf.

Additionally, the radio was simplified by omitting the synthesiser entirely; for a fixed frequency there are cheaper ways of generat-

The GPS radio is about 50mm square. Production version (left) has its own screening can, while the same circuit in a die-cast box (centre) is released to licensees to enable them to access the signal path while debugging their own versions. (Top right) radio without screening; (bottom right) low-noise amp normally built into the base of the antenna.
ing the local oscillator signal even at 1.5 GHz. In the hardware processing, simplicity in the extreme is achieved by processing everything in software.

Conventional hardware is replaced by a shift register, counter and link adapter that pack the samples from the radio and send them up the transputer's serial link where a DMA engine puts them into memory without CPU involvement. This four chip /o solution can be replaced with two pafs if required, much cheaper but using more power. In the computer side of the hardware, ultimate simplicity is achieved by using a Transputer.

It needs just two address latches to interface with eprom and ram. I/O for keyboard and screen can be done on a two wire serial link as was the radio data. This provides a computer of up to 30mips, 256kBytes, with on-chip 64-bit floating point unit in just 11 chips. If the FPU is not required, a version of the transputer is available in high-volume for under 20 dollars, and is being used in several volume GPS equipments world-wide.

As explained in an earlier article, a dual down conversion receiver is difficult because of problems with the image filtering: there is no correct first IF. Despite this, it has been the industry standard. A single down conversion usually has problems with the RF image, and a triple has excessive complexity and cost. My design tackles the problem by using a direct conversion to baseband. Requiring discrete components, complexity has to be avoided at all costs. The Plessey design goes the opposite direction since, being all on silicon, complexity comes free. It therefore uses triple down-conversion to 4.3MHz, not baseband, which means there is a fourth conversion performed later. Both radio modules are about 50mm square on single sided board with ground-plane. Both consume about 700mW.

Discrete design

Component selection was aimed at volume products intended for satellite TV and the cellular telephone market. The main supplier for my radio design is Avantek, now part of Hewlett-Packard. It offers gain blocks, essentially self-biased transistors, with 50Ω input and output impedance and unconditionally stable.

Running from a 5V supply, these provide 15dB gain at 15mA with a 4-5dB noise figure. (MSA06 family). Another version offers 2-3dB noise figure, and 25dB gain, and still takes 15mA (INA031 family). This would appear wonderful, but its layout demands conflict with the requirement for a low-cost board, so it is only used in the LNA.

The signal path needs some 120dB of gain, in order to bring the -114dBm thermal noise up to 0dBm (equivalent to about 0.6V p-p across 50Ω), allowing for losses in filters etc. To maximise stability, this should be evenly split between RF and baseband, but as baseband gain is less expensive, and consumes less power, there may be some bias in that direction. The LNA provides about 40dB gross, which after filter and cable losses yields about 3dB measured. The mixer provides about 10dB of conversion gain, thus there remains 120-43, or 77dB to provide, and this is split with 60dB at baseband, and 34dB gross at RF. The block diagram Fig. 1 shows the gain block line-up.

The 60dB of gain is maintained in the baseband circuit, but the 34dB at RF loses 3dB in the L-band filter, and 10dB in dielectric losses in the circuit board, due to the low-cost board used. Besides intellectual challenge, the other reason for keeping the physical size small is to minimise these losses.

Designing with the Avantek MSA06 family is simple. They offer 50Ω connections so all that is required is to select the bias resistor to suit the supply voltage in use, and select input and output coupling capacitors. The bias resistor is chosen to operate the device at the correct current, usually around 15mA but bear in mind that a temperature dependent 3.5V is developed across the device. At 8V and 300Ω resistive collector load, the design checks out correctly for both MSA and INA devices. However at 5V, the MSA is satisfactory, but the INA could be dangerously overrun if it were at the low end of the device voltage, so an alternative approach is required.

Instead of the resistor, a constant current feed is used via a PNP transistor, itself decoupled at RF, and isolated by an inductor. As the bias resistor is effectively in parallel with the output for signal purposes, it reduces the gain considerably at this low value. This is solved by adding an inductor in series to keep the RF out of the bias path.

At 1.5 GHz both discrete and stripline construction are possible; the latter may be too large, but discrete wired components do not behave normally. A wound inductor exhibits a noise figure than a LNA can precede the filter, and shows a better noise figure than a complex GaAsfet circuit placed after the filter.

Front end

The LNA consists of one INA031, a filter, and one MSA06, with circuitry to allow power feeding up the coax, and in the 5V version, to stabilise the feed to the INAO31. The power feed consists of an inductor to isolate 1.5GHz, decoupling, then an inductor to isolate lower frequencies that may have been picked up on the coax, and more decoupling. The general form is shown in Fig. 2.

The INA031 is placed before the filter to yield a more stable amplifier with a noise figure not degraded by filter losses. This saves 3dB from the noise equation. Disadvantage is that powerful local transmissions may saturate the front end device, or even burn it out... the latter only likely with badly sited aerials for radar or satcomms on the same boat.

Also, two strong transmissions with a sum or difference frequency of 1.5GHz could mix to cause interference, which would not have occurred if the filter were first. Indeed, over-powering intermediate products at 1575MHz can originate in a broadband front end from transmitter pairs operating at frequencies far below the GPS slot. Products follow the general formula fsum=|f1+f2|, where f1 and f2 are

![Fig. 1. New radio architecture: direct down conversion allows a very simple radio architecture, and makes the local oscillator signal frequency where low-cost filters are available.](image1)

![Fig. 2. A simple, robust LNA can precede the filter, and shows a better noise figure than a complex GaAsfet circuit placed after the filter.](image2)

![Fig. 3. Feeding the LNA up the coax is common practice, but not easy at these frequencies. A cascaded pair of inductors does the job, but to be safe against short circuit, should really be fed from a current limiting transistor. A resistor is not adequate, due to variation in current between different LNAs and lack of voltage headroom.](image3)
are interfering transmitters and $m$ and $n$ are integers generally in the range 1 to 5.

In this design, I retain the low-noise version, and depend on the selectivity of the patch antenna to rejet such interference, both in frequency and in response angle since it possesses a high degree of out-of-band rejection.

A quad helix antenna would be better suited with the filter first.

The main RF gain sections are a further two MS06 stages, preceded by another ceramic filter, both to give further stop band rejection, and to reject any interference picked up on the coax. Two MS06 stages are used rather than one IN031, despite the 15mA penalty, both because of the cost (2 at £1 versus one at £4), and because it is very difficult to keep the IN031 stable on the thick low-cost board. Avantek recommend multiple vias under each ground pin on ultra-thin board, and although feasible in production, this was not feasible for prototype boards.

The reason for the criticality is that the IN031 is a two stage device, and if there is any inductance in the ground lead, it causes unwanted coupling back from the output stage to the input, causing oscillation. The only other complication is the power feed up the coax to the LNA with signal and DC isolated by inductors.

Mixer

I could write an entire article on the mixer, so long did it take to get it right. The original plan was to use the Avantek MS86 two port mixer in self oscillating mode, locked to a low level injected signal. (Fig. 4).

A prototype was built using this system, but was found to be easily pulled off frequency by signals from the signal path; being a two port mixer, RF and LO at 0dBm was present after the mixer, necessitating serious filtering. To feed a low frequency amplifier for the baseband filter covered later, so another approach was tried.

During development, a low cost plastic three port mixer became available, Fig. 8. This vastly simplified the circuit, as being a three-port, balanced mixer with buffered IF and LO ports, almost no external circuitry was required. The resistive T previously used to couple the LO into the signal path could go, resulting in 6dB more signal, and 6dB more LO, with three less components.

The input and output matching circuits described above could also be removed, and the baseband filter no longer had high power LO and RF breakthrough to remove, as the mixer is balanced. So much better was the performance with this mixer that one stage of baseband gain had to be removed.

The baseband filter

The baseband filter has two functions... to remove RF and LO signals from the path, and to set the desired bandwidth. With the MS86 mixer, RF and LO at 0dBm was present after the mixer, necessitating serious filtering.

As shown in Fig. 9, a small inductor blocks 1.5GHz from reaching the second mixer, as that one could not operate at such high frequencies. The second inductor/capacitor pair then operate from fifty ohms into several kilo-ohms, to set the desired bandwidth. With the MS86 mixer, RF and LO at 0dBm was present after the mixer, necessitating serious filtering.
The mixer also includes an output buffer, so it is tolerant to mismatch on the output. Thus the baseband filter becomes a simple L-C pair. The other problem in the four production stages is the oscillator chain, as by feeding the LO into the signal path not at the mixer, but one stage earlier; the LO signal could be boosted in the RF path. However the move from a two port to a three port mixer reduced this possibility, and some of the capacitor values were not suitable for production. For example there was a phase shift network with two 0.5pF capacitors.

Also, the separation of the 525 MHz harmonic without either pulling the 105MHz oscillator out of lock, or saturating the next stage with bleed through of 105MHz, meant very delicate coupling of the two stages. Fig. 12 shows the full circuit. The final version extracts the 15th harmonic of the 105MHz signal directly using a GPS frequency ceramic resonator. This is not so much filtering out the existing 15th harmonic, rather hitting the high-Q filter hard and allowing it to ring for five cycles. This presents a high impedance to the fundamental, so does not disturb the oscillator. It provides very good coupling at 1575 MHz. As a result it was possible to generate −3dBm of LO with just two stages, meeting the requirement of no manual tuning or select-on-test components. In fact the only frequency determining components other than the 105MHz crystal are the feedback phase-shift network that selects the correct overtone.

The smallest capacitor used is now 5pF, and there is about 20% tolerance in the circuit. Conveniently, the phase shift network can never quite achieve 180° to compensate for the 180° shift in the amplifier stage, so the crystal must provide the rest.

The oscillator frequency is always about 3kHz low, resulting in a LO signal some 80kHz below nominal which is exactly what is required. Note that the reference frequency does not need to be precise.

While the software does not like a rapid change in frequency, the absolute value is immaterial. Thus calibration of the GPS receiver consists of placing the antenna near a known signal source and running the self-test software.

The nominal centre frequency is then stored in non-volatile memory, rather than the reverse operation where the centre frequency is moved. For the professional surveying GPS receiver, where the sampling clock and the LO are known signal sources, the software does not like a rapid change in frequency, the absolute value is immaterial. Thus calibration of the GPS receiver consists of placing the antenna near a known signal source and running the self-test software.

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Fig. 11. Discrete version baseband section. Four transistor stages – three for voltage gain, the last for driving the 50Ω coax when the radio is remote from the CPU, replaced the op-amp design.

Input
Loop
175 MHz
filter
Front end mixer
Stage 2 amplifier
mixer
VCO
Loop
Filter
-10
10
Reference
Detector
Lock indicator
Dividers
AGC
AGC out
latches
magsign
sample clock
AGC ap
latches
I I
magsign
sample clock
not adding to the problems of the system.

The noise in the system is, and should be,
defined by the first stage of the LNA at the
antenna. The other element that does add to
the loss is using a single-bit A/D converter.
This costs about 2dB versus an ideal system,
about 1dB versus other realisable schemes, but
in return allows the software spread-spectrum
processing, and removes the need for an AGC
in the radio.

The added flexibility of software acquisition
processing more than recovers this 2dB loss,
as it is acquisition that defines the minimum
received power threshold. The new radio
design has no image frequency to reject, and
as the GPS military band is 20MHz wide, and
the commercial 2MHz at its centre, the com-
bination of tuned patch antenna and ceramic
resonator filters means that there are few prob-
lems at RF. The interference power equation
runs like this. The radio has some 50dB of RF
gain, with a maximum permissible power of
+10dBm in the amplifier, so -40dBm is the
maximum tolerated from the antenna. The two
RF filters provide 60dB of attenuation to the
nearest interferers, and the patch antenna
another 20dB, so an interferer of up to
+40dBm (10W) incident (not transmitted)
power, would be needed to upset the radio.
Such power could only possibly occur as
pulse power from a radar, and if applied con-
tinuously would certainly destroy the LNA.

If this radio has a weak point, it is break-
through of unwanted signals through coax
cables, power supply, and chassis shielding.
Because there are only two frequencies, RF
and Baseband, the gain, and hence sensitivity
of each has to be high.

The RF circuitry is very well protected, pro-
vided good quality coax is used between LNA
and radio. However the baseband circuitry has
some 60dB of gain at 0-1MHz, and it
is extremely hard to prevent long and medium
wave broadcast stations drowning the set if
installed in a car driving beside the transmitter.
The development prototypes have avoided this
problem by construction in diecast aluminium
boxes, and this is fine for professional appli-
cations.

However with the target price for car use
being so low, a combination of power line fil-
ters and plasticised metal casing is needed.

Single chip radio version
The GEC-Plessey GP1010 provides all the
active circuitry required for the GPS radio
except the LNA. This is ideal, partly because
the LNA may need to be remote, and partly
because it thus allows the choice of silicon or
GaAsFet LNA, with the bi-polar radio chip
offering the lowest cost and highest density.
Its architecture is a triple downconversion to
4.3MHz, with the synthesiser for the three local oscillators provided on chip.

The system designer must provide the LNA as discussed, a frequency reference at 10MHz, the sampling clock of his choice and all but the last IF filter. A block diagram is shown in Fig. 13.

Signal path

The IFs are 175MHz, 35MHz AND 4.3 MHz. The 175MHz filter is built from discrete components and is not critical in design, as its main function is to reject the image frequency at 165MHz. While care must be taken to shield the board from local VHF transmitters, especially by decoupling the power rails, there should not be any hostile signals here. If there were, they would have been at 1505MHz, nasty terrestrial microwave links, but these can be taken out by the ceramic 1575MHz resonator, or two if a wideband helical antenna is used.

The main band shaping is done at 35MHz in a saw filter. This allows very sharp edges to the passband, with minimal ripple and phase distortion. While a dedicated filter is available, it is very similar to those used in the IF channels of colour televisions, so is not expensive.

At this frequency, it can be provided on chip... a feature that angered me greatly initially, as it locked me into GEC-Plessey's frequency plan, and 4.3MHz is too high for my own radio described earlier. 2.0MHz is 300kHz, ideal for the processing developed for a GPS-specific transputer prototype during 1991.

It used a 20MHz reference, the transputer clock, and divides this into two non-overlapping clocks A and B at 10MHz. In general, a clock, and divides this into two non-overlapping clocks A and B at 10MHz. In general, stream A is used, but if the phase of the tracker has to be delayed, one pulse is deleted from the A stream.

If the tracker has to be advanced, one pulse from the B stream is gated into the A stream between its own pulses. If one pulse is inserted for every 87 pulses of the 20MHz reference, the modified A stream becomes 10.229885 MHz or 11ppm below 10.23 MHz. This is perfect, as even in synchronous operation a drift rate is required in order to give vernier style measurement.

The software implementation of this is discussed in a later article. The division of the 20MHz reference to give 10 and 2.046 MHz can be done in three TTL chips, or in two 16L GALs, but even this complexity can be avoided with some sneaky software that clocks the simulated code-generator asynchronously... so my implementation of the GP1010 radio uses the on-chip oscillator at 10MHz with an external 74HC390 dual decade counter that generates 2MHz for the sampling clock, 5MHz for the processor clock, and 256kHz for the byte I/O clock.

The on-chip synthesiser.

The clever part of this chip is a 1400MHz VCO. Once this is included on chip, all other frequencies can be derived from it, while it itself is locked back to the 10MHz reference. Another clever feature is the choice of 1400MHz. This means that the same design can be used for the military GPS L3 frequency, at 1226MHz. However it also means that

Aliasing tricks

However there are tricks that can be played by deliberate aliasing with the sampling clock. If the sampling clock were put at 2.0MHz, the signal in the computer would appear centred at 3000kHz, ideal for the processing developed for my own radio described earlier. 2.0MHz is easily derived from the 10MHz reference. If

Board screening ring is 50 x 47mm, with the power supply, RF and output connectors outside. Two dual op-amps used for baseband gain (left of board) and L-band filters for the GPS frequency (right) are shown. The low cost gain-blocks - self-biased transistors - are used in the RF path, the mixer and local oscillator generation.

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there can be no transmissions on the image frequency, as it is reserved for GPS down
transmission. The GP1010 provides a gain stage to maintain oscillation, but a TCXO is
required to maintain oscillation, but a TCXO is recommended. A TCXO is far too expensive
for automotive use, but on the other hand, the software can absorb carrier shift easily, it is
less tolerant of sampler rate change. This is the main reason for the change from 2.14MHz
synchronous operation to 2.00MHz asynchronous, as the 2.14MHz at the required 10ppm tolerance requires temp
erature compensation, while the asynchronous operation can tolerate 25ppm
rule without problem, up or down. The 25ppm limit relates to requiring an integer
number of bytes in samples in each millisecond,

- The 1400 MHz is divided by ten to give a 140MHz as the second LO. It is also
divided by 140 to look to the reference, and by

35 to generate a 40MHz clock, used by a future partner chip.

Frequency reference
The frequency reference itself merits some discussion. The GP1010 provides a gain stage to maintain oscillation, but a TCXO is recommended. A TCXO is far too expensive for automotive use, but on the other hand, the software can absorb carrier shift easily, it is less tolerant of sampler rate change. This is the main reason for the change from 2.14MHz synchronous operation to 2.00MHz asynchronous, as the 2.14MHz at the required 10ppm tolerance requires temperature compensation, while the asynchronous operation can tolerate 25ppm tolerance without problem, up or down. The 25ppm limit relates to requiring an integer number of bytes in samples in each millisecond, as it is handled in bytes.

Acknowledgement
I am grateful for the assistance of GEC-Plessey Semiconductors at Swindon for access to their early GP1010 application board, and for the information contained in their Application Note AN-139.

Philip Morris is a consultant engineer working for users.

Next month: DSP software to extract the GPS signal out of the noise.
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Transistors are so powerful and versatile that just a handful are needed to build almost any kind of high-performance circuit: a fast op amp, a video buffer, or a unique logic circuit. They are also uniquely adept at causing trouble. For example, a simple amplifier will probably not survive if the input is shorted to the power supplies or the output to ground. Fortunately, most op amps include forgiving features, so that they can survive these conditions. When the μA741 and LM101 op amps were designed, they included extra transistors to ensure that their inputs and outputs would survive such abuse. But an individual transistor is vulnerable to damage by excessive forward or reverse current at its input, and almost every transistor is capable of melting. So transistor circuits should be designed so that the transistors do not blow up, and circuits must be analysed if they do.

A simple problem is to install the transistor correctly: three terminals mean the possibility of a wrong connection is considerably greater than with a mere diode. Small-signal transistors are often installed so close to a printed circuit board that it is not clear whether or not the leads are crossed or shorted to a transistor's can or to a PC trace.

Next to proper installation comes correct design, and unless they are completely protected from the rest of the world, transistors require input protection. Most transistors can withstand dozens of milliamperes of forward base current but will die with "only a few volts" of forward bias. Military standard MIL-HDBK-217F states that a circuit's reliability decreases when components are added. Yet when resistors or transistors are added to protect an amplifier's input or output, the circuit's reliability actually improves.

Similarly, pumping current out of the base of a transistor will cause the base-emitter junction to break down or "zener." This reverse current - even if it is as low as nanoamperes or very brief in duration - tends to degrade the low-current beta of the transistor, at least on a temporary basis. So in cases where accuracy is important, find a way to avoid reverse-biasing the inputs.

Transistors are also susceptible to ESD - electrostatic discharge. Charge yourself up to a few thousand volts by walking across a rug on a dry day, then touch your finger to an NPN's base. It will probably survive because a forward-biased junction can survive a pulse of a few amperes for a small part of a microsecond. But pulling up the emitter of a grounded-base NPN stage, or the base of a PNP, risks reverse-biasing the base-emitter junction. The reverse bias can cause significant damage to the base-emitter junction and might even destroy a small transistor.

When designing an IC, sensible designers add clamp diodes, so that any pin can survive a minimum of + and −2000V of ESD. Many IC pins can typically survive two to three times this amount. These ESD-survival design goals are based on the "human-body" model.

Ballast resistors, also known as sharing resistors, are often connected to the emitters of a number of paralleled transistors (a) to help the transistors share current and power. In an integrated circuit (b), the ballast resistors are often integrated with adjacent emitters.
in which the impedance equals about 100Ω in series with 1500Ω. With discrete transistors, whose junctions are considerably larger than the small geometries found in ICs, ESD damage may not be as severe. But in some cases, damage can still happen.

Delicate RF transistors such as 2N918s, 2N4275s, and 2N2369s sometimes seem to blow up when they are little more than are just looked at because their junctions are so small.

Other transistor-related problems arise when engineers make design assumptions.

Every beginner learns that the $V_{BE}$ of a transistor decreases by about $2mV/°C$ and increases by about $60mV$ decade of current, and this should not be forgotten or misapplied in extreme temperatures.

Sloppy assumptions about $V_{BE}$ should not be made either. For instance, it is not fair to ask a pair of transistors to have well-matched $V_{BE}$ if they are located more than 0.1in apart and there are heat sources, power sources, cold drafts, or hot breezes in the neighbourhood.

March 1993 ELECTRONICS WORLD + WIRELESS WORLD

Matched pairs of transistors should be glued together to improve results. For best results, monolithic dual transistors like the LM394 give the closest match.

It is fair to assume that two matched transistors with the same $V_{BE}$ at the same small current will have about the same temperature coefficient of $V_{BE}$. But make no rash assumptions if the two transistors are from different manufacturers or from the same manufacturer at different times.

Similarly, transistors from different manufacturers will have different characteristics when going into and coming out of saturation, especially when being driven at high speeds. A components engineer is a very valuable person to have around and can save a lot of grief by preventing unqualified components from confusing circuit performance.

Another assumption engineers make concerns a transistor’s failure mode. It is often said that a transistor, like a diode, fails as a short circuit or in a low-impedance mode.

For the design of hybrids, make sure the substrate of a chip is connected to the correct DC level. The bottom of a fet chip is usually tied to the gate, but the connection may be a large and unspecified impedance.

The substrate of a discrete bipolar transistor is quite predictable. At room temperature, $g_m = \frac{38.6 \times I_s}{V_{BE}}$ - much more consistent than the forward conductance of diodes. Since voltage gain is defined as $A_v = \frac{g_m}{Z_{in}}$, computing it is often a trivial task. This simple equation may have to be adjusted in certain cases: for instance if an emitter-degeneration resistor $R_e$ is included, the effective transconductance falls to $1/(R_e + g_m)$. $A_v$ is also influenced by temperature changes, bias shifts in the emitter current, hidden impedances in parallel with the load, and the finite output impedance of the transistor.

Higher beta devices can have much worse output impedance than normal.

Also, although the transconductance of a well-biased bipolar transistor is quite predictable, beta usually has a wide range and is not nearly as predictable. Adverse shifts in performance can result if the beta gets too low or too high and causes shifts in operating points and biases.

Another way to increase effective “beta” is to use the Darlington connection, though this may degrade voltage gain and noise, make the response a little flaky, and decrease the base current only slightly. I keep learning more and more reasons not to use Darlingtoners or cascaded followers, and for many years, it’s been more important (in most circuits) to have matched transistors than to have sky-high betas. You can match betas yourself, or can buy monolithic dual matched transistors like the LM393. Or you can buy four or five matched transistors on one monolithic substrate, such as an LM3045 or LM2086 monolithic transistor array.

One of the attractions of bipolar transistors is that their transconductance, $g_m$, is quite predictable. At room temperature, $g_m = \frac{38.6 \times I_s}{V_{BE}}$ - much more consistent than the forward conductance of diodes. Since voltage gain is defined as

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Field effect transistors

For a given operating current, field-effect transistors normally have much poorer $g_m$ than bipolar transistors - measure devices to see how much lower. Additionally, the $V_{th}$ of fet can cover a very wide range, thus making them harder to bias than bipolars.

Jfets (junction field-effect transistors) became popular 20 years ago because they could be used to make analogue switches with resistances of 30Ω and lower. Jfets also help make good op amps with lower input currents than bipolar devices, at least at moderate or cool temperatures.

The National Semiconductor bifeet process made it feasible to make jets along with bipolars on a monolithic circuit. It is true that characteristic of the best Bifet inputs are still slightly inferior to the best bipolar ones in terms of $V_{BE}$ temperature coefficient, long-term stability, and voltage noise.

But these Bifet characteristics keep improving because of improved processing and innovative circuit design. As a result, Bifets are better than bipolar transistors in terms of voltage accuracy, and offer the advantage of low input currents, at room temperature.

Jfets can have a larger gate current when current flows through the source than when no current flows - called $I_{gs}$ This looks to be caused by impact ionisation, or "hot carriers." Either way, the gate current has a tendency to increase as a linear function of source current, with an exponential dependence on high drain-source voltages.

For the design of hybrids, make sure the substrate of a chip is connected to the correct DC level. The bottom of a fet chip is usually tied to the gate, but the connection may be a large and unspecified impedance.

The substrate of a discrete bipolar transistor is the collector. Most linear and digital IC substrates are tied to the negative supply. Exceptions include the LM117 and similar adjustable positive regulators - their substrates are tied to $V_{out}$. The LM196 voltage regulator's substrate is tied to the positive supply voltage, $+V_{cc}$, as are the substrates of the MM74H00 family of chips, the NSC LM741 or LM308 and the LMC601 family, and most of the dielectrically isolated op amps from Harris.

So, be aware of an IC's substrate connection. If an LM101 op amp's metal can should not fail, but develop a fault at some time in the future producing obscure effects.

Mosfets, widely used in digital ICs, are also very useful in analogue circuits such as analogue switches. Quad switches - eg CD4011B and CD4066 - are popular because of their.

Beta better?
The $h$ parameter, $h_{ij}$, is equal to $\Delta V_{BE}/\Delta V_{CE}$ with the base grounded. Many engineers have learned that as beta rises, so does $h_{ij}$. As beta rises and $h_{ij}$ rises, the transistor's output impedance decreases; its Early voltage falls; its voltage gain decreases; and its common-emitter breakdown voltage, $BV_{CEO}$, may also decrease. Early voltage of a transistor is the amount of $V_{CE}$ that causes the collector current to increase to about double its low-voltage value, assuming a constant base drive. $V_{FSat}$, is approximately equal to $26mV \times (I_{ds})$. So, in many circuits there is a point where higher beta simply makes the gain lower, not higher.
Most mosfet-input linear ICs do have protection diodes and may be able to withstand 600V, though they can not usually survive 2000V. When working with unprotected mosfets, such as the 2N160, keep the pins securely shorted until the device is soldered into its PC board in which the protection diodes are already installed. I do all of that and wash the transistor package with both an organic solvent and soap and water. I also keep the sensitive gate circuits entirely off the PC board by pulling the gate pin up in the air and using point-to-point wiring. Air, which is a superior dielectric, is also a good insulator. So far, I haven't had any blown inputs or bad leakages - at least nothing as bad as 10A.

On the other hand, when using cmos digital ICs, always plug them into live sockets, never use conductive foam, and never wear a ground strap on the wrist. Beware of any devices that manufacturers claim are safe from ESD.

In some cases cmos ICs with ESD may fail instantly, but may become unreliable and fail at a later time. So beware of latent unreliability problems. If you must troubleshooting cmos ICs while not grounded, or plug them in while the power buses are hot, remember that the result could be long-lasting harm to an occasional IC.

**Power transistors may hog current**

The temptation when building bipolar transistors bigger and bigger is to go to extremes and make a huge power transistor. But there are practical limitations. Soon, the circuit capacitances cause oppressive drive requirements, and removing the heat is difficult. But the most serious limitation is secondary breakdown, when a transistor is driven outside its "safe operating area." At very high currents and low voltages, the distributed emitter resistance of the device - which includes the resistance of the emitter metal and the inherent emitter resistivity - can cause enough I x R drop to force the entire emitter and its periphery to share the current.

But half the current and double the voltage: dissipation is the same, but the I x R drop is cut in half. Now continue to halve the current and double the voltage to a point where the ballasting will not be sufficient, and a hot spot will develop at a high-power point along the emitter. The inherent decrease of V BE will cause an increase of current in one small area. Unless current is turned off promptly, it will continue to increase unchecked, "current hogging" is a failure mechanism, and may cause the area to melt or crater - secondary breakdown, exceeding the secondary breakdown of the device. Designers of linear ICs use ballasting, cellular layouts, and thermal-limiting techniques, all of which can prevent harm in these cases.

Some discrete transistors are beginning to include these features. Fortunately, many manufacturer data sheets include permitted safe-area curves at various voltages and for various effective pulse-widths. So, it is possible to design reliable power circuits with ordinary power transistors. Probability of an unreliable design increases as the power level increases; as the voltage rises; as the adequacy of the heat sink decreases, and as the safety margins shrink. For example, if the bolts on a heat sink are not tightened enough, the thermal path degrades and the part can run excessively hot.

High temperature per se does not cause a power transistor to fail. But, if the drive circuitry was designed to turn a transistor on and only a base-emitter resistor is available to turn it off, then at a very high temperature, the transistor will turn itself on and there will be no adequate way to turn it off. Then it may go into secondary breakdown and overheat and fail.

Overheating does not by itself cause failure. But good practice is to stop your power transistors heating up, and to have a base drive that can pull the base off when they do.

Problems may also result if the screws on the heat sink are too tight, or if the heat sink under the device is warped; or if it has bumps or burrs or foreign matter on it. Tightening the bolt too much will overheat and warp the transistor and die attach, and may cause the die to pop right off the tab. The insulating washer under the power transistor can crack due to overstress or may fail after days or weeks - or months. Even without an insulating washer, overtorqueing the bolts of plastic-packaged power transistors is one of the few ways a user can mistreat and kill these devices.

**Apply the 5s rule**

A finger is a pretty good heat detector - just be careful not to burn it with high voltages or very hot devices. A good "rule of thumb" is the 5s guide: if you can hold your finger on a hot device for 5s, the heat sink is about right, and the case temperature is about 85 °C. For components hotter than that - too hot to touch - dot the finger with saliva and apply it to the hot object for just a fraction of a second. If the moisture dries quickly, the case is probably around 100 °C; if it sizzles instantaneously, the case may be as hot as 140 °C. Alternatively, buy an infrared imaging detector for a price of several thousand dollars. You won't burn your fingers and will get beautiful
mosfets avoid secondary breakdown

When it comes to power transistors, mosfets have certain advantages.

Mosfet switching is faster than bipolar transistors, with smaller drive requirements, and mosfet devices are inherently stable against secondary breakdown and current hogging because the temperature coefficient of $I_{DS}$ vs $V_{GS}$ is inherently stable at high current densities. If one area of the power device overheats, it tends to carry less current and thus has an inherent mechanism to avoid running away – a self-ballasting characteristic that is a major reason for the popularity of mosfets over bipolar transistors. But recent criticism points out that running a mosfet at high-enough voltages and low current means the current density gets very small, the temperature coefficient of $I_{DS}$ vs $V_{GS}$ reverses, and the device’s inherent freedom from current hogging may be lost. So at high voltages and low current densities, watch out for this possibility.

At high enough $V_{GS}$, mosfets can exhibit current hogging and “secondary breakdown” similar to that of bipolars, though newer power mosfets are considerably more reliable and less expensive than older devices. Even though a lot of transistors may be necessary to turn the gate on or off quickly, unlike with a bipolar transistor, a lot of amps are not required to hold it on. The newer devices can be turned off quicker, too, if enough transient gate drive current is available.

But mosfets are not without their problem areas. Too many watts into a mosfet will melt it just as with a bipolar device. If not overheating, the easiest way to cause a problem is to forget to insert a few dozen or hundred ohms of resistance (or a ferrite bead) right at the gate lead of the device. Otherwise, these devices have such high bandwidths that they can oscillate at much higher frequencies than bipolar transistors.

As with bipolar transistors, mosfets are very reliable if their voltage, current, and temperature ratings are not exceeded. Dissatisfaction with a device’s reliability or performance usually stems from the drivers or the related circuitry. Most mosfets have a maximum $V_{GS}$ rating of just 20 or 25V. They may temporarily survive operation with 30 or 50V on the gate, but it is not safe to run it up there forever. Applying excessive gate voltage may produce gradual gain and threshold degradation. Also, power mosfets are not quite as rugged as bipolars when it comes to surviving ESD transients. A common precaution is to add a little decoupling, clamping, or current-limiting circuitry, so that terminals accessible to the outside world can withstand ESD.

Dmos fets are so easy to apply that it is easy to forget about the parasitic bipolar transistor lurking in parallel with them. If $dV/dt$ is too large at the drain, or the drain junction is avalanched at too high a current and voltage, or the transistor overheats, the bipolar device turns on and dies an instant death due to current hogging or an excursion from its safe operating area.

I’m accustomed to linear ICs, which have protection transistors built right in, so the user rarely has a problem. (But most of the transistor troubles are left to the IC designer). Discrete designs are appropriate and cost-effective for many applications, but the availability of linear ICs – especially op amps – can simplify a design considerably, while improving reliability.

References

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P8237B 0-20V-1mA — Voltage & Current Controls - Twin Meters
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MULTIMEDIA - the perfect marriage?

Is it a television? Is it a computer? Julia King reports how the advent of multimedia is blurring the distinction between the two as the prospect of an all-in-one entertainment and information system moves inexorably towards a reality.

Multimedia unites all the current screen-based technologies, harnesses them to a computer storage mechanism such as a CD-rom and adds in superior sound quality. So is it the ultimate realisation of all consumer trends: or is it just marketing hype? The answer may well pivot on whether users can cope with an even greater barrage of information than they are being fed at the moment.

The ultimate multimedia aim is for a workstation or PC to form the central component of a system, providing a source of information and a means of entertainment which can be tailored to meet a particular field of interest.

Users will be able to view and/or listen selectively, instead of being limited to watching whatever is being broadcast by television or radio stations or loading a pre-recorded video that happens to be on the shelves. The emphasis is on greater choice for entertainment and education.

Perhaps the best way to understand the structure of a multimedia system is to think of it as a layered database - at any point, the user can interrogate the system to delve deeper into a particular layer or subject. A subject might be represented as a tree, with the trunk forming the main structure and different areas represented by branches which then sub-divide again and again until the leaves are reached. The whole structure is not static but is growing continually, updated remotely - perhaps via a satellite link or over a network.

"Pasting" moving images

The concept still has a long way to go. But one company that has advanced to quite a degree with its thinking is Digithurst. The Royston-based company began in the image analysis business around ten years ago and now produces PC-compatible image capture, processing and compression cards. It also markets the Windows 3-based multimedia authoring package, PictureBook, designed to enable electronic books to be created by combining objects that contain text, graphics and images (either still or moving) into pages. Text and graphics can be imported from other applications because the system is Windows-based, and artificial intelligence is used to form a linking mechanism where the software looks for similar combinations of words and matches them. The package is typical of those being produced by software houses, with products and operating systems encouraging the use of multimedia.

Microsoft's Gillian Kent sees another use of multimedia in "kiosk-type" applications. For instance, a visitor to a show may be seeking information on a particular type of product. The system installed in a booth or kiosk can be interrogated about the product's location in the show and graphics or still or moving images can be brought up to help give the visitor the information required.
DSP becomes the slave

The move towards multimedia has created a shake up in the digital signal processing industry. In PCs most signal processing has been carried out by the host CPU without a separate DSP chip. But as the requirements of multimedia increase, these host CPUs are finding it harder to cope and a need has arisen for dedicated CPU chips. One firm that has already gone down this road is Olivetti which is using an Analog Devices DSP to compress digitised audio signals on one of its machines.

Julian Hayes from AD said: "This is a new problem in positioning audio signal processing than using the host CPU.”

Until now, most applications for DSP circuits have seen the DSP as the master of a logic system. Put the DSP in a computer and there already is a master – the CPU. This has forced the DSP designers to develop special operating systems and system managers for their DSP chips, such as Texas Instruments and IBM's Mwave.

Jay Reimer from TI said: "In a multimedia PC there is already a master CPU. The DSP cannot be the master. It has to cooperate with the existing CPU. The operating system needs to comprehend that environment.”

He added: "It also has to be an operating system that's tailored to the real-time scheduling needs of DSP tasks. It has to support dynamic welding and scheduling of tasks. This is new to the DSP world.”

Also, the nature of multimedia means there is a need for multichannel I/O of different types, and of multifunctional channels. This I/O needs to be integrated onto the chip in an upgradeable way.

Reimer said: "I want to provide different types of functions in different ways. I want an appropriate number of serial ports on the same chip as the DSP CPU. We have already seen this, and we will see more of this.”

One problem has been the conservative nature of large computer companies in adding anything new to their machines. It is one thing for a user to buy a DSP board to plug in the back. It is another for it to be included as standard in a cost conscious market.

Hayes said: "When you put the boards in as a standard fitting every penny counts. You have to reduce the hardware costs and provide the right software support. Compression and recognition algorithms are very complex. The challenge for us is to make the algorithms available bundled with the DSP.”

Price is moving in the right direction. In 1986 AD introduced a 16bit fixed point DSP for more than £100. A similar chip, the ADSP2105, released this year costs less than $10 in volume.

Fixed point is suitable for most multimedia applications such as audio and data communications, but there is a possibility that video may go to floating point and some algorithms for these have already been developed. The problem is that most video applications are being done quite well in software and it is not yet clear whether a DSP will be needed.

Steve Rogerson

Over at Digithurst Peter Kruger is looking forward to the newspaper of the future, when the written word will no longer be sufficient and newspapers will be reader-complied, with the user selecting only stories of interest.

The concept is an interesting one, conjuring up images of commuters with their notebook PCs busily reading the day’s news from their screens. But perhaps more usefully, an electronic newspaper is being launched for the blind in March by a company called Etna, Royal National Institute for the Blind, The Guardian. Aptech and Intelligent Research have between them produced a system that enables the blind reader to select: stories. The newspaper is transmitted overnight to the reader’s PC, via the channels normally used to transmit teletext, and can be caye-load via a Braise reader, voice synthesiser, or a special large print screen.

Initially we certainly will be limited to The Guardian, though it is hoped that other newspapers and magazines will be available in the future.

The system is not yet able to transmit pictures, but it is notable as marching the harnessing of the written word with computer storage and database-like access techniques.

City applications

Kruger is keen to exploit teletext as an information source. Digithurst's TV1 card can capture text while its infra red remote controller could be used to switch between different TV channels or to control a video recorder. The TV2 card replaces the remote controller with a tuner module allowing it to decode TV signals.

Teletext cards are primarily being used in City applications, according to Kruger: "Because they are cheaper than the Reuters feeds. People are starting to look at teletext for the dissemination of information".

The received teletext image is stored in a database released on user request – and the system can be programmed by the user to search for particular areas. Those pages will be updated automatically as the teletext screens are changed, saving the normal wait to access a teletext page.

"Teletext is a massive media tool that is already set up", says Justin Howard, one of Digithurst's engineers. "You want a teletext receiver that is capable of special software control.”

Apple’s Pamela Schure agrees that multimedia is more suitable for business than dome-tic applications – at least for the moment. "At home you have to ask, what do people need, and how do you explain it to them? You also have to think how can we get the cost down?"

Even in business, she is not convinced that there is always a need for video: "It has to be appropriate”, she says, with training or demonstrating products being two suitable uses.

Multimedia architecture

Back in October, Apple launched a new version of its multimedia architecture for the Mac. QuickTime 1.5 – now bundled with upgrade products for the System 7.1, operating system – supports Kodak’s Photo CD technology and allows photos digitised onto a CD rom to be imported into applications and altered as required. It can also handle full-screen full video motion at frame rates of up to 30 frames/s.

Schure stresses the differences between the Mac’s architecture and that of dos-based platforms, which have "no real architecture for handling timebase data". The Mac operating system does not require special drivers to be written. For example, a video card broadcasts its potential which is recognised by the operating system.

Mac uses ADPCM (adaptive pulse code modulation) techniques for interleaving sound and video. The run time of any video is maintained: if a video is intended to last ten seconds, it will last ten seconds. Sound is interleaved automatically and if there is any problem, video frames are dropped, not sound.

Apple has developed its processes to the stage at which users can cut, copy and paste movies. QuickTime offers support for Word and WordPerfect, allowing movies to be opened from either. "We are changing the dynamics of the broadcast industry”, says Schure.

A ready Apple is viewing CD rom as the distribution mechanism for multimedia, a development Schure describes as "vital”.

Apple has saved itself thousands of pounds on postage by disseminating information internally on CD roms. Now, apart from the most basic model, almost every Mac sold has a CD rom facility built into the CPU.

A though the cost of CD rom is dropping (bought separately, an Apple CD rom drive will cost £275), Schure believes it is still not low enough to encourage home use of multimedia. A lot more work and thought needs to put in before multimedia is truly workable as a concept. The evolutionary process is underway, but the market will take a long time to be educated.

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<td>Schematic Capture and PCB CAD</td>
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Amateur television is no longer just for the dedicated enthusiast. John Cronk explains how widespread repeaters combined with off-air beacon signals and receiver designs – such as his own – make amateur TV accessible to a much wider audience.

Picking up a clearer picture of amateur TV

Interest in amateur television has increased steadily following the licensing of the first 1.3GHz repeater in 1984. There are now 18 repeaters around the country and hundreds of amateur stations equipped to send and receive amateur television.

In the past, the popularity of amateur television was inhibited by the fact that the chance of finding a transmission at random was low. How likely was it to be aiming in the right direction on the right frequency at the right time? As a result, ATV contacts usually had to be arranged beforehand.

But development of repeater stations with their channelised frequency, omni-directional antennae and good locations means there are now much better opportunities for enthusiasts to view some ATV.

When not actively repeating, most repeaters transmit a beacon signal to aid receiver adjustment, reminiscent of G9AED. There is also a recommended simplex frequency of 1255MHz.

A range of about 30km can be expected from 24cm TV repeaters, but contacts in excess of 150km have been reported. Receiving antennas need to be as high as possible, as propagation is generally line of sight at this frequency, but freak conditions are frequent and signals from the continent are possible from good locations.

Due to low atmospheric noise at this frequency, there is scope for using low-noise mast head amplifiers and high gain antennas, but a modest wide-band 10dB gain Yagi fed
Fig. 2. Front end module uses an inexpensive gasfet with 0.9dB noise at 2GHz at 20mA. The 0685 is unconditionally stable, has 5052 and at 16mA offers 17dB gain at 1GHz.

Fig. 3. Tuner layout shows compartments conveniently screened using PCB. Microstrip is lengths of PCB of set width positioned using adhesive.
with high-quality low-loss coaxial cable can perform well within the service area of most repeaters. Video standards conform to the UK broadcast standard, CCIR system L, and comprise 625 line pal colour with 6MHz intercarrier sound. Repeaters however use frequency modulation. Usually, peak deviation is 3.5MHz where pre-emphasis is used, requiring a bandwidth of 18MHz. All stations use horizontally polarised antennas. At a first glance satellite tuners may seem ideal for receiving ATV since most operate between 750 and 1500MHz. They do work but because they are intended to be used with much higher deviation, they perform poorly. Their bandwidth is excessive and the demodulated video output is low.

Being simply tuners, satellite TV receivers need a preamplifier to provide useful sensitivity. Most will not tune the 6MHz intercarrier sound. Their wide bandwidth also makes them susceptible to radar interference, which is widespread on 1.3GHz from air traffic control systems. The design that follows is a receiver that tunes to ATV between 1248 and 1308MHz and is easy to construct at low cost. As a bonus it can be made portable, operating from 12V, and feed a 75Ω monitor with 1V composite video.

Microwave equipment is difficult to create, but the 50Ω devices used here simplify the design and implementation. The outcome is a practical design that is typical of current amateur 24cm ATV receivers. Fig. 1, shows the initial layout. I used automatic gain control but the 6MHz intercarrier sound. Their wide bandwidth also makes them susceptible to radar interference, which is widespread on 1.3GHz from air traffic control systems. The design that follows is a receiver that tunes to ATV between 1248 and 1308MHz and is easy to construct at low cost. As a bonus it can be made portable, operating from 12V, and feed a 75Ω monitor with 1V composite video.

The design is basically an IF amplifier/bandpass filter with adjustable gain. It is tuned to 1298MHz, and is simple to create. The transmitted signal is filtered to pass frequencies up to 40MHz, and is amplified by 20dB. The current of 20mA. According to the fet data sheet a noise figure of 0.9dB at 2GHz is obtainable with a drain

GaAsfet tuner module

At the heart of the IF amplifier front end is an inexpensive Avantek 20135 GaAsfet, Fig. 2. According to the data sheet a noise figure of 0.9dB at 2GHz is obtainable with a drain current of 20mA. To keep losses low, the input tuned circuit consists of an air-spaced adjustable line. This circuit is heavily loaded and as a result quite broadly tuned, but in addition to matching the

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### Amateur TV repeaters and frequencies

<table>
<thead>
<tr>
<th>Callsign</th>
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<th>Location</th>
<th>Status, April '92</th>
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<tr>
<td>GB3CT</td>
<td>RT2</td>
<td>Crawley, East Sussex</td>
<td>Operational</td>
</tr>
<tr>
<td>GB3ET</td>
<td>RT2</td>
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<td>GB3GT</td>
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<td>Bellahouston, Glasgow</td>
<td>Temp. U/S</td>
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<td>Temp. Low Power</td>
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<td>RT3/2</td>
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<td>Linked to GB3TV</td>
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<td>FM Only</td>
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</table>

Repeaters are also under consideration for: East Sussex, East Kent, Birmingham, Goole, Anglesey, Dorset and Northampton.

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![Fig. 1. IF amp/bandpass filter](image)

![Fig. 4. In the intermediate-frequency amplifier, an easy to align Butterworth bandpass filter with adjustable gain centres on 40MHz. Values in parenthesis are theoretical.](image)
IF Module construction

A small choke and resistor form a non-resonant load that couples the fet to the MSA 0685 Mosfet. This is an unconditionally stable gain block with 50Ω input and output impedance that requires 16mA for about 17dB gain at 1GHz. When the supply current is correct, the monolithic microwave IC will have 3.5V across it.

For this circuit the off-the-shelf double-balanced mixer shown is perfect, but it is expensive. Costs are minimised by using the 1GHz version. The double-balanced mixer used is an HC2 by W&G Instruments but according to reports the well known SBLI-X, also rated at 1GHz, has proved satisfactory at 1.3GHz.

For the first IF amplifier another MMIC feeds through a 50Ω low-pass π filter broadly tuned to 40MHz, providing a low impedance to ground for the signal and local-oscillator frequencies.

Oscillator power of at least 7dBm is required for the balanced mixer so local oscillator circuitry must be well screened from the rest of the circuit. The circuit shown provides even more output, allowing an attenuator to be added. This improves the matching needed to take full advantage of the mixer characteristics. All the stages are connected through 50Ω microstrip.

A 10V supply is obtained using a readily available 5V regulator but a 10V type would reduce component count. Battery operation necessitates protection but if the differential between battery voltage and the necessary 10V is low, the series diode shown can be replaced with a parallel fuse-shorting diode.

Fig. 5. Layout details of the IF amplifier with bandpass filter. Again, PCB is used for screening but here, one side of a double-sided screen conveniently carries components.

Layout details

The module was built in a tinned metal enclosure with a ground plane of double sided PC soldered near the centre to form two screened compartments. Figure 3 provides mechanical details.

The 50Ω microstrip is made from 1.6mm thick glass-fibre single sided PC (G10, 0.062in thick, dielectric constant 4.5) cut into 2.54mm wide strips. Cuts are made in the copper for the coupling capacitors and wider slots are cut for the MMICs. Apart from the oscillator line, L3, which together with the other components determines the local oscillator frequency, the lengths have no electrical significance.

The only holes through the ground plane are for a feed through capacitor, the oscillator output, the pins of the balanced mixer and a mounting screw for the regulator.

The balanced mixer is in the oscillator compartment, its case soldered to the ground plane. All micro strip lines are glued to the ground plane with cyanoacrylate based glue.

Clearance holes for the microstrip should be as small as possible and the two screens should be soldered all round.

These screens form two unwanted high Q cavities containing the input and output circuits of the RF amplifier. Unless their Q is reduced either by not using a lid or lining the lid with an RF loss-causing material, instability can result. Special microwave lossy materials are available but the conductive foam used to pack static sensitive semiconductors may suffice.

Undesignated RF chokes are made by winding about three turns of the tail of the associated resistor with a 1.6mm inside diameter. Capacitors in the local oscillator must have especially short leads. Commissioning the local oscillator is made easier if a frequency
counter is available. Small adjustments may be needed to make the oscillator tune the range 1278 to 1208 MHz.

Since microwave transistors are delicate, static sensitive devices, their installation is best left until last. A useful tip is to tin the points where the device is to sit, and temporarily link them to ground with fine wire. Earth the soldering iron to the enclosure too. Keeping one hand in contact with the module, unpack the device and position it, using one finger to hold it in place. Next disconnect the soldering iron, even if it is a low voltage one, and using its residual heat, solder the transistor into place and remove the grounding wires.

Check the current through the gasfet is about 18mA by measuring the voltage across the 47Ω source resistor. Final tuning adjustment of the input matching can only be made when the other modules are completed.

IF module

As Fig. 4 shows, this module has a Butterworth 50Ω band-pass filter with adjustable gain centred on 40MHz. Its theoretical values are shown in brackets. As the filter is easy to align by adjusting the coils, the more common capacitor values shown were used in the prototype. The 47pF and 100Ω components correct the filter’s source and termination independence.

At 40MHz the Modamps have high gain and as their power is current sourced, their input impedance can be used to control their gain very smoothly. The selectivity curve has the typical FM shape factor, Fig. 6. Strong signals, due to the amplification, can be used to control the gain very smoothly. The selectivity curve has the typical FM shape factor, Fig. 6. Strong signals, due to the amplification, can be used to control the gain very smoothly.

Demodulator

In the demodulator of Fig. 7, the first stage is a common base connected BFY90 that accepts low-impedance output from the IF module. It has a broad-band collector load that further attenuates out-of-band noise. Still containing amplifier information, the signal feeds a rectifier and the phase-locked-loop demodulator. Resulting rectified DC can be used to feed a signal-strength meter or for alignment.

As shown, the Signetics NE564 PLL demodulator circuit is a well tried configuration and particularly recommended. Output feeds an emitter follower whose low-impedance output feeds a passive 75Ω de-emphasis network that corrects video for the CCIR 405-1 characteristic.

The Signetics NE592 variable-gain video amplifier can be adjusted to 2V of video for feeding an emitter follower output stage. Video polarity can be made selectable. Although transmission is usually positive, i.e. peak white is a higher carrier frequency, the modulation sense of the video can be inverted in the receiver if the local oscillator is on the other side of the IF circuit.

Video gain control does not need to be controlled via the front panel as the level of demodulated video only changes if the transmitter deviation is altered.

The output is intended to feed a colour or monochrome monitor with 1V composite video when terminated with 75Ω.

Layout details

For this section, a double sided PC with its top side forming a ground plane is the best technique. Most of the resistors are mounted vertically, and the ICs are soldered directly, without bases to the board.

As the board needs only be about 50 by 90mm, the components are close together so that the other side acts as the shielding.

The Modamps are mounted on 50Ω microstrip as in the tuner. Being self supporting, the inductors can be adjusted by stretching the turns. An actual response curve is shown in Fig. 6.

Demodulator

In the demodulator of Fig. 7, the first stage is a common base connected BFY90 that accepts low-impedance output from the IF module. It has a broad-band collector load that further attenuates out-of-band noise. Still containing amplifier information, the signal feeds a rectifier and the phase-locked-loop demodulator. Resulting rectified DC can be used to feed a signal-strength meter or for alignment.

As shown, the Signetics NE564 PLL demodulator circuit is a well tried configuration and particularly recommended. Output feeds an emitter follower whose low-impedance output feeds a passive 75Ω de-emphasis network that corrects video for the CCIR 405-1 characteristic.

The Signetics NE592 variable-gain video amplifier can be adjusted to 2V of video for feeding an emitter follower output stage. Video polarity can be made selectable. Although transmission is usually positive, i.e. peak white is a higher carrier frequency, the modulation sense of the video can be inverted in the receiver if the local oscillator is on the other side of the IF circuit.

Video gain control does not need to be controlled via the front panel as the level of demodulated video only changes if the transmitter deviation is altered.

The output is intended to feed a colour or monochrome monitor with 1V composite video when terminated with 75Ω.

Layout details

For this section, a double sided PC with its top side forming a ground plane is the best technique. Most of the resistors are mounted vertically, and the ICs are soldered directly, without bases to the board.

As the board needs only be about 50 by 90mm, the components are close together so
RF DESIGN

Fig. 8. In the sound section, a single IC converts 6MHz FM from the demodulator into 1.5W of audio power.

Physically small capacitor types should be chosen. All parts are mounted on the ground plane side of the board except for the 1nF capacitor between pins 3 and 11 of the NE564 which is mounted on the track side.

Sound stages

For the audio, an SGS TDA11902 forms a complete TV sound channel capable of delivering 1.5W into an 8Ω load, Fig. 8.

Pin 12 forms both a low-level audio output and an input to the AF amplifier. With capacitive decoupling to output from the signal-strength meter, it could also be used as an aid to antenna alignment via noise level monitoring. Power to this section need not be stabilised.

If it is possible to fit this unit on a circuit board of similar size to that of the video circuit, it is the ideal tool for the IF circuits, but a signal generator and a multimeter on the signal paths on 1.3GHz are limited to line of sight, this should not be read too literally. A 160km contact could be used to adjust the antenna input circuit. The series capacitor, its tapping point, and the effective length of the input line are adjusted for the best signal-to-noise ratio.

Commissioning

How the receiver is aligned depends on the equipment available. Obviously, a wobbulator is the ideal tool for the IF circuits, but a signal generator and a multimeter on the signal-strength meter connection can be used to set up the IF filter. The coils can be squeezed or stretched, and the trimmer adjusted to achieve a symmetrical response.

Applying a 6MHz signal to the output of the NE564 detector allows the sound carrier to be checked via an oscilloscope or video monitor. The trimmer between pins 12 and 13 of the NE564 is best set when receiving a weak FM television signal.

Gain of the NE592 should be set for 1V of video at the output socket when receiving a correctly modulated signal and terminated in 75Ω. The coil on pin 4-5 of the TDA11902 sound IC is simply adjusted for best quality sound while receiving a suitable signal.

Finally, a weak signal or a modulated noise generator could be used to adjust the antenna input circuit. The series capacitor, its tapping point, and the effective length of the input line are adjusted for the best signal-to-noise ratio.

Propagation

Microwave propagation is an area where experience complements the textbook. Radio amateurs often refer to “getting the feel of the band.”

Most UHF and microwave propagating modes require visibility. Ionospheric reflection, common on the short wave bands, is unlikely, but reflection from buildings, gas holders and even trees is possible.

Refraction – bending due to the troposphere – is frequent. Waves can be bent and lost or refracted to follow the curvature of the earth. Temperature, pressure and moisture content affect the refractive index of air, so weather conditions can have a significant effect on signals. Widespread warm, dry, high pressure conditions during the daytime can result in layers of air with different densities forming at evening times. Signals can be ducted between these layers with low losses over remarkable distances.

As the input circuit is loaded by the antenna it is not at all sharply tuned.

References

1. ARRL Handbook 1987, Ch. 32, p. 28, UHF and Microwave Equipment

Further information

For more information on amateur television contact British Amateur Television Club’s membership secretary, Mr D Lawton, at “Greenhurst,” Pinewood Road, High Wycombe, Buckinghamshire, HP12 4DD or ring 0494 28899.

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Letters

Filter tips
In his article "Building bricks into brick wall filters", EW + WW, June pp.461-464 Bashir Al-Hashimi lists four steps in the design procedure of the FDNR filter.

On the first step, forming the LC prototype filter, data for this is available as indicated in the article. But the books are expensive - around £100 for Saal's Handbook (ref 2) and not found in many libraries - and these may well put off the novice designer faced with hundreds of pages of tables. Low cost computer programs are available to carry out this part of the design and require only the minimum of information to give the same results as in Fig 2. You reviewed one of my BBC computer programs in August 1989 which did exactly this.

Step two, transforming a filter to its dual circuit, is extremely easy. Values are the same as the minimum inductor prototype. All that has to be done is allocate them to different components, for example shunt C becomes series L. Most books of filter tables have this included.

The new design method for scaling that Al-Hashimi puts forward is virtually identical to that used by Williams in his "Electronic filter design handbook". I realise this is always the problem when the word new is used. Using FDNR implementation of steep cut-off low-pass filters has some practical problems not mentioned in the article. Particularly with elliptic filters, the Q factors of the zeros in the transfer function are fairly high. This causes a significant reduction in signal headroom.

For the example in the article I have plotted frequency response at the op-amp outputs in each stage. Amplifiers A3/4 have the highest peak at 12dB. This means that with a typical upper limit of ±10V output swing (±15V supplies) the input must be restricted to 1.7V RMS sinewave, or 1.9V peak square wave. A good trap for the novice.

As the introduction to the article gave a typical use as anti-alias filtering you might be interested in a circuit I have used to correct for the Sin(x)/x sampling loss. It corrects for this loss to within 0.1dB up to 4kHz F/2 for 8kHz sampling and is easily scaled.

David Markie
Ascot
Berkshire

Speed kills theory
I was interested in D Di Mario's article "Gravity and electric force link up in black hole?" (EW + WW, February 1993).

He creates an ingenious theory to explain the near equality in numerical value between Planck's time of 2.395 x 10^-43s, and the ratio of gravitational force to electric force of 2.2 x 10^-43.

However, there is one point that he has failed to consider - if the speed of rotation of the earth had been different from its present value by, say, 10%, the length of the second would be different, and there would be no coincidence to explain. Hence any theory to explain the coincidence must be able to predict the rate of rotation of the earth. This D Di Mario has failed to do.

JS Linfoot
Oxford

Where is the low region?
I was anxious to read Anthony Hopwood's article "Natural radiation focused by power lines" (EW + WW, November 1992) on a new link for power line cancer deaths.

I have been peripherally involved in the power line health effects issue for some while. As part of my work I have led a group that has made a number of fibre optic measurement systems over the last few years, particularly for the isolated measurement of electric and magnetic fields, both AC and DC. Aware of the small magnitude of the fields at or near ground level near power lines, we have been following the reports of the various attempts to link power lines to health effects.

The article suggests an interesting possibility but leaves me with a few questions.
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One is the idea that the power lines concentrate "the biologically most destructive lower energy particles". Presumably the term lower is relative to what particles normally of interest to cosmic ray researchers, but still high enough to be biologically damaging. Some numbers would be useful, indicating the range of energies normally studied, the range now being measured, and the range known to be biologically hazardous.

The author proposes the idea that the alternating fields associated with the power lines are sonic and therefore averaged because the particles travel so fast they respond only to the instantaneous fields. The average value of the alternating field is zero. The same holds for the magnetic field. The average focusing effect of either the electric or magnetic fields of the power line must therefore also be zero. Are we to understand that the focusing effect is due to the energy flow of the line (the average value of which is clearly not zero) or must we assume a nonlinear interaction of some sort?

Some measured results are presented (Fig. 2) and apparently show that there is an increase of up to a factor of two in the particle count to one side of the power line. The graph seems to be labelled incorrectly. The minimum count is shown as 100% above normal. I assume that the figure is meant to show half a symmetrical distribution but perhaps not.

Where is this increased count represent additional particles, rather than additional energy per particle? Where do these additional particles come from? I visualise the proposed effect as a concentration of particles descending as a more or less uniform flux from the high altitude radiation belt. The motion of these particles I imagine to be affected only within a few metres of the power line, a scale that is in keeping with Fig. 2. This amounts to a conservation of particles law, so should there not be a region of decreased count somewhere?

If there is a region of decreased count on the other side of the line the side not shown in Fig. 2, a symmetry argument might support the idea that the concentration effect is due to power rather than the electric or magnetic fields. Yet the article points to the line current as the significant parameter. And while it is suggested that, from the sky radiation point of view, the best place to be is right under the wires, the graph shows the count to be normal, but not decreased, there.

It is known that solar activity can affect the operation of power systems. The fields and ions associated with flare activity can induce slowly varying currents in long lines, particularly in the near-polar regions where auroras are seen. These currents may result in asymmetrical transformer saturation, either in power transformers or measurement transformers. So 1.}

### Under the brolly

To a retired agriculturist now a wireless historian, looking objectively at crossed field antennas - "CFA - no tricks" (EW + WW, Letters, December 1992), and previous letters and articles - it is an enigma.

Consider an electron disturbance propagating along a linear conductor represented by Fig. 1. The magnetic field lines are concentric, and the electric field lines parallel to the conductor, that is the fields are crossed. Conventional theory is that acceleration of the electrons causes energy to be radiated as em waves.

The pioneers reasoned that for maximum radiation, the negative and positive- going electron disturbances propagating along the antenna, which to all intents are half a wavelength, must be able to extend linearly to their full length.

But early transoconic signalling was by VLF, and the largest practical antenna structures were only a fraction of wavelength long. So, in order to cause RF current to flow in a short open ended conductor, the antenna structure was arranged as a huge LC tuned circuit, often as an umbrella supported by a mast typically 0.02X tall. This is represented by Fig. 2.

The umbrella and earth-screen formed a capacitor brought into resonance by a variometer, which in some installations was placed under the earth screen to minimise interaction with the magnetic field around the mast. Electric field lines connected the umbrella and earth screen, while the magnetic field lines were concentric with the mast, that is crossed fields.

Greater capacity allowed greater current in the mast. But available data indicates that this in itself did not increase radiation, which was very poor, presumably because the electron disturbances were confined largely within the variometer but the umbrella antenna did work, and by definition was a CFA.

Consider now a scaled down version of an umbrella antenna with the linear element replaced by a solenoid (Fig. 3); the philosophy being to reduce the physical size of the radiator. The plates create the electric field and the solenoid the magnetic field, but this is doughnut shaped. If em waves are radiated, can it be reconciled with established theory that concentric magnetic field lines are vital, or is there a new esoteric theory unknown to the pioneers?

The evolution of linear antennas from inductors and air capacitors (elevated plates or spheres) was the result of an enormous number of experimental experiences. Yet, for VLF communication, the umbrella prevailed and indeed with modifications still prevails. Moreover, if the arrangement shown in Fig. 3 worked, surely it would indeed with modifications still prevail. Moreover, if the arrangement shown in Fig. 3 worked, surely it would have been used for ULF signalling instead of Sanguine type antennas with linear elements 160km long.

George Pickworth

### Fig. 1. Electric and magnetic fields surrounding an electron disturbance propagating along a travelling wave antenna

### Fig. 2. Representation of umbrella type antenna indicating concentric magnetic field lines.

### Fig. 3. Effect of substituting a solenoid in place of the linear conductor.

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**Some mac comeback**

In your comment "Unacceptable standards" (January EW + WW) you ask why anyone should insist on the fitment of decoders (D2mac) for which there are presently no programmes and virtually no prospect of there ever being any.

You seem to less than well informed. Presently there are the following satellites/channels transmitting in D2mac:

- Eutelsat H-F3
- Telit-X
- Intelsat 11

**Happiness is an engineer**

Against all odds I find myself in full agreement with your editorial "Pay for the arts" (EW + WW, November 1992).

With perhaps the exception of modern art, that has been sadly commercialised, I like art in all its forms but at best it is a relaxation and does not contribute to the general well being of mankind. Our only hope for a high standard of living for all people lies in the hands of engineers who must be given all the assistance they need to produce the devices and systems to feed, clothe, shelter, educate, keep healthy and happy the world's population.

The technical societies, such as the IEE, bemoan the fact that engineers are not accorded the same level of social standing as people in other professions, and this is no doubt true. Such societies are too busy with a worship for academic qualifications, which deservedly do not impress the public at large, instead of reminding the public that the standard of their whole future is dependent on the skills of engineers and associated scientists.

Much as I hate to say it, it must be obvious to intelligent people that much of the money spent on art would be better directed to the wide field of engineering and used for the development of systems for the general good of all people.

FG Clifford

Wetton

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would not rule out the possibility of other interactions between power lines and solar activity. On the other hand, I am not convinced by the Hopwood article.

Much of what Hopwood suggests should be capable of rather straightforward empirical study. Some of the line research facilities of the power industry are set up to measure line profiles of, say, audible noise. It might not be too difficult or expensive to add Geiger counters to the parameters being measured, and obtain data from test lines in which the current and voltage can be separately controlled.

I therefore support your sentiment (EW + WW, November 1992) that the suspicion of a link between power lines and health effects will not go away until this is done.

Harold Kirkham
Sunland
California
USA

Particle poppycock

With regards to "Natural radiation focused by power lines: New evidence" (EW + WW, November 1992), I am considerably dismayed that an article containing so many scientific misconceptions should be published in what has always been considered a responsible and technically accurate publication.

For 25 years I have been doing engineering work for a group studying ionising radiation with origins external to the earth, by making direct measurements at locations from the surface out to the Jupiter radiation belts.

It is unfortunate that you should give prominence to an article containing so persuasive a message for the general reader, as there are persons only too willing to join the anti-powerline campaign.

I agree that there does seem to be evidence of association of power line corridors with childhood leukemia; it very doubtful indeed that this could be due to enhancement of naturally occurring ionising radiation by electric or magnetic fields caused by power transmission lines. These fields are orders of magnitude too weak to provide local focusing or enhancement background radiation, and they also vary in magnitude and sign.

The particle most easily deflected, the electron, for energies which would allow it to be focused, has a range in air of only a few centimetres. I do not doubt Anthony Hopwood's diligence but his lack of knowledge of charged particle physics is evident to anyone who works in this field.

Whether he has been observing, it is not charged particle focusing.

The problem needs further investigation, as he says. There may be some reason to suspect (in North America) the former use of herbicides and defoliants for control of scrub growth along the powerline corridors as a cause of increased childhood leukemia.

John Firth
Ottawa

Rationalising peace

The Scientists for Global Responsibility (SGR) organisation not only wants about the irresponsible uses of science and technology, but also supports their constructive uses. The successful Science for the Earth forum, organised jointly with Scientists for the Earth, in Cambridge in October 1992 is an example.

SGR was formed early last year and incorporated bodies include Scientists Against Nuclear Arms, Psychologists for Peace, and Electronics & Computing for Peace (ECP).

ECP intends to concentrate on two projects during this year. One is childhood support for Alasdair Philipps' research on health effects of electromagnetic fields, especially from electric power lines. The other is to have a presence at this year's Milcom (military computing) exhibition.

ECP is also adding the experience of its Ethics at Work group to that of SGR's Science and Ethics group. I believe this will become one example of several in which the whole is more than the sum of the parts.

Alan Cottee
Scientists for Global Responsibility
London

Beatles in the ground

Andrew Ainger's letter, "Keeping an ear to the ground" (EW + WW, December 1992), concerning communication through the ground using earth currents reminded me of my own experiments in this area.

In the late 1960s, when I was about 16 years old, it was suggested to me by my father that I might try earth current transmission. I set up a transmitter of about 20W output at audio frequencies feeding two aluminium stakes about 0.6m deep into the ground about 1m apart at various sites. This mobile apparatus reproduced a signal at about 300m. My friend also had a 20W transmitter and I still have a recording of the Beatles "Long and winding road" transmitted the 200m from his house to mine.

This means of communication is definitely not new. Louis Meulstee in his article "Earth current signalling" quotes Samuel Morse as the first to achieve electrical signalling without wires. On the December 16 1842 Morse used direct current telegraphy across a 23m wide canal in Washington. The same article gives great detail of the telegraphy signalling sets used in World War I by the British signalling service and similar devices used by the French and German forces.

These devices used an alternating current interrupted by the Morse key and generated by a mechanical vibrator. The note of the buzzer was heard at the receiving station directly in a telephone earpiece. It is reported that British power buzzers gave ranges of 1000 to 5000m.

While I always had ideas of comb filters to reduce mains interference or better still the use of an FM carrier, I never tried these techniques. One phenomenon I do remember is a background noise which occurred randomly lasting one or two seconds and best described as sounding like paper slowly torn.

With ranges for base band audio not exceeding 1000m the hopes of starting a net by using this technique must, unfortunately, be small.

Red Brown
Nottingham

Crash response

With reference to the letter "Crash solution" (EW + WW, November 1992) asking for clarification of C1 and C2 capacitors and flux cancelling inductors. X class capacitors are defined in BS2135 and their use defined in BS613. In simple terms only, they should be connected from a live mains line to earth, on equipment connected via a plug and lead, that is most equipment. Their value is limited to less than 5nF to limit the earth leakage current which could flow via the equipment user if the earth connection was faulty.

They are made with robust dielectric and voltage proofed to 2250V to ensure a minimal risk of breakdown, against stray currents. X class capacitors are connected from live to neutral. As the leakage cannot flow to earth (or a victim) larger values can be used, say 0.1μF, and then the X class accepts as a failure would only bring out a fuse, with little risk to an equipment user. Note there are various grades of X class capacitors X1, X2, X3 designed to meet various specifications on spikes and so on.

The obvious problem with inductors used in a filter is possible saturation of the iron material due to the load current. Using low permeability ferrite or gapped cores is OK but many turns are needed to produce the large value of inductors required to give good filter performance at low frequency.

An alternative solution is to use a high perm material toroid with two windings passing the live current back and forth through the second. The flux then cancels, the core is not saturated, and we have a cheap high value inductor, typically 20mH. This value inductor, together with the larger X class capacitor gives good attenuation for any common mode (symmetric mode) interference, that is a spike that appears on live and neutral together.

Of course any differential or symmetric mode interference is on line and neutral but opposite polarity produced between the lines will suffer the same flux canceling as the main current, that is only a small inductance will be seen, typically 0.1μH. The small inductance together with the larger X capacitor should give equally good attenuation to a different spike.

The values quoted would give a low pass filter providing useful attenuation right across the broadcast radio frequency bands. But such a filter provides little protection from audio and sub-audio clicks and spikes (from Mike Whittaker’s freezer, toaster, and vacuum cleaner for example). Altogether larger and more expensive filters are required incorporating voltage dependent resistors to clip spikes and larger values of inductors and capacitors.

BS613 allows up to 2μF on permanently connected household apparatus.

John Foster
Entelid
Boating continues to be pushed to new heights of high-tech intricacy. Messing about these days usually means unravelling skeins of cabling, connecting meters for speed, wind, depth, plus radar, GPS, VHF, to the chart table, cockpit, autopilot or to each other.

Integrated instrumentation has gone a long way to sort out this complexity - such as the Network system being shown at this year's boat show by Brookes & Gatehouse - but even the B & G display panel does not handle one big electronics feature of the show: Differential GPS.

DGPS is designed to improve accuracy of the positioning system from 100m to 5-10m. Some manufacturers are still arguing that small boats do not really need it. But in a restricted seaway, in fog, knowing precisely where you are could make a difference. A fix to the nearest hundred metres is, potentially, much less safe.

Unfortunately, the irritation for yachters and other civilian users of GPS, is that the satellite-based location network is technically quite capable of delivering pinpoint accuracy - as are the receivers they can be bought for a few hundred pounds. Indeed it was because the commercial receivers proved so unexpectedly precise that the US Defence Department, which developed the system, deliberately degraded the signals in case they proved useful to enemy forces. As a result, interference, known as Selective Availability, reduces accuracy to 100m.

To restore the accuracy, some organisations (US coastguards, initially) retransmit the signals from ground stations, whose known location can be used to identify and eliminate the SA-induced errors, restoring the possibility of a 5-10m fix. The comparatively local nature of the beacons, limited to 400 miles, is something the Pentagon can learn to tolerate.

The first DGPS-ready receivers at this year's show - including a hand-held unit from Magellan (Nav 500D) at £899, teamed with a £499 differential beacon receiver - were perhaps a little premature.

UK waters are at present covered by two privately-operated systems, Scorpio and Seamate, with subscriptions of around £600 a year, protected by smart-cards in leased receivers. Present customers are chiefly commercial shipping and fishing fleets (location of lobster pots is enormously simplified). But, some European countries plan to introduce state-run, free DGPS services in the near future.

UK customers are chiefly commercial shipping and fishing fleets (location of lobster pots is enormously simplified). But, some European countries plan to introduce state-run, free DGPS services in the near future.

DGPS exhibitor Trimble Navigation, is first in the field with a CD-based chart system, using a 7 x 6in LCD display. It costs about twice as much as the rom-cartridge/green CRT type, but can play audio CDs as well. Trimble clearly appreciates what most yachting is about.

Satellite comms

Inmarsat-M is a new service, bringing direct-dial speech telephony to smaller boats - 40ft or under - by using digital voice coding. Required bandwidth is reduced from 25kHz to under 10kHz with a corresponding reduction in antenna size and weight: the 1.2m antenna and 120kg all-up weight of the big-boat Inmarsat-A system is cut down to 50cm and 20kg. Equipment costs still remain high though, at around $20,000. Launched at the show was one of the first systems, the Magnavox MX 3400.

One rethink in satellite boat technology is scrapping of its plans for live video links with boats in the Whitbread Round the World Race, starting in September. The belated discovery that yachts roll, and are thus likely to suffer loss of contact between their Inmarsat-A dishes and the satellite, has forced the rethink.

Instead, BT is perfecting the store-and-forward "videoclip" technology - digitising picture and sound at 384kbit/s in a video codec (coder/decoder) and then feeding it out at 64kbit/s, thus taking 12min to transmit a 2min clip. Picture quality is said to be no worse than S-VHS.

Loss of contact during transmission affects this process too - producing a mosaic of the picture - so BT is now working on error-correction capability.

Development deadline has been put back from January to May, this year.
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Bryan Hart describes a graphical technique for describing feedback amplifiers that reveals fundamental operating characteristics.

Pictorial display of the characteristic features of an electrical system can often provide an insight not immediately apparent from a bare mathematical study. Well-known examples are the reflection chart used to keep track of pulse reflections on lines and cables; the Smith chart for solving problems (eg stub matching) with transmission lines driven by sinusoidal signals; and the Karnaugh map, used to minimise Boolean expressions in digital logic.

But it is also possible to use pictorial display to clarify the operating principle of a negative feedback amplifier, of especial interest to those who prefer diagrams to lengthy algebra. The technique is easily applied to amplifiers with low-frequency open-loop voltage gain, not necessarily large, and a frequency response dominated by a single 3dB roll-off frequency. Application in the case where open-loop gain is large, offers a deeper understanding of the virtual-earth principle employed in the analysis of operational amplifier circuits.

The approach—which includes novel features such as graphical determination of the DC/low-frequency closed-loop gain—might best be described as the "Method of ganged-phasor".

**Basic amplifier**

In the basic amplifier configuration of Fig.1 amplifier A is assumed to be linear and to have, initially, a frequency independent voltage gain $A_o$. The sinusoidal input, error and output voltages are represented by the phasor quantities $V_i$, $e$, and $V_o$ respectively. Phasor symbols are sometimes shown in bold face in texts on circuit theory, but there is no advantage to this convention here, although the phasors themselves will be shown as bold lines on diagrams.

$R_1$ and $R_2$ comprise a summing network. Regarding this as a potential divider with one end held at a potential and the other at $V_o$ gives:

$$e = (V_i - V_o)R_2 \over R_1 + R_2 + V_o$$

reducing to:

$$e = aV_i + bV_o$$

where, by definition:

$$a = R_1 \over R_1 + R_2$$

$$b = R_2 \over R_1 + R_2$$

However the amplifier gives:

$$V_o = -A_o e$$

so by combining equations

$$aV_i + bV_o = e = -A_o e$$

Let $V_o = -V$ and we obtain

$$aV_i - bV = e = -A_o$$

The construction producing the first pictorial display is based on a graphical interpretation of the above equations. Crucial to construction is determination of the input for an assumed output—not usually the approach adopted in texts on circuit analysis but typical of the design process. For a power amplifier the sensible procedure is to design the output stage required to deliver a specified power to a load and then work backwards to the input available from a transducer.

The general procedure is:

- Draw three horizontal axes I, II and III (Fig. 2a).
- Line I at $y = 0$ is intended for the display of $V_i$, II at $y = R_1$ for $e$ and III at $y = -(R_1 + R_2)$ for $V_o$. The vertical scale in kΩ/cm is chosen for easy scaling with the resistor values used in a particular problem.
- Construct a vertical line, IV, to define the zero voltage axis and mark its points of intersection H, J, $K$ with I, II and III."

---

**Diagram Lettering**

- Mark point $L$ at $V_o = -V$ on III, point $M$ at $1/A_o$ on II and construct $v$, the diagonal $LM$.
- This cuts IV at $N$ and, when extended, intersects I at $P$.
- The numerical value used for $V$ (in mV or V) is one of practical convenience.
- Construct IV, the diagonal $PK$, and mark $Q$ where this cuts II.
- Draw VII, the diagonal $HL$, and mark $S$ where this cuts II.

Diagram lettering starts at $H$, rather than $A$, avoiding some of the letters already associated with electrical parameters. For this reason $R$ is also omitted: 0 is missing because it could...
be confused with zero or an origin.
The geometry of the diagram (Fig. 2a) gives the following relationships:

\[ \frac{JQ}{HP} = \frac{R_1}{R_1 + R_2} = a \]
\[ \frac{JS}{KL} = \frac{QM}{KL} = \frac{R_1}{R_1 + R_2} = b \]

Thus, the phasor interpretation of the horizontal line sections is

- \( KL = V (= -V) \)
- \( JS = QM = bV (= -IN) \)
- \( IM = E \)
- \( HP = V \)
- \( JQ = aVi \)

These phasors are shown more clearly in Fig. 2b (Fig. 2a redrawn but with the line lettering omitted). The phasors are spatially separated but geometrically linked. If one changes all the others change in a related way, analogous to the mechanical linkage or "ganging" of electronic components, e.g. capacitors, used in variable frequency oscillators – hence, the suggested name "ganged-phasors". Figure 2a offers some insight into circuit operation, an aspect pursued later, but the diagram need not be drawn in full if the problem is merely graphical determination of \( L_A C_L (0) \), magnitude of the DC and low frequency closed-loop gain.

An example is calculation of \( L_A C_L (0) \) for the case \( R_1 = 10k\Omega, R_2 = 20k\Omega, A_0 = 10 \).

Figure 3, not to scale indicates the method. Construct i, ii and iii with a vertical scale 5k\Omega = 1cm. Now to find point \( V \) = 50mV gives an integral value (5mV) for \( \epsilon \). On a horizontal scale of 1mV = 1mm, mark points \( L \) and \( M \), construct \( LM \) and extend it to \( P \). Measured length of \( HP \) will be 32.5mm, and \( L_A C_L (0) = 50/32.5 = 1.54 \) which agrees with the value obtained by a purely algebraic calculation.

Suppose \( V_o \) had been equal to -30mV instead of -50mV. The modified construction line \( L \) to \( M \) to \( P \), shown dotted in Fig. 3, also passes through \( N \). In fact, from the geometry of the figure, 'the location of \( N \) on \( HK \) is independent of the value of \( V_o \), resulting in two practical consequences. \( LM \) need not be extended to \( P \). If preferred we could measure \( HN \) and \( NK \) and make use of the alternative relationship \( |L_A C_L (0)| = NK/HN \).

Secondly, the output for a given input is easily found once the location of \( N \) is known. Construction of Fig. 4 takes this further, following Fig. 3 up to the marking of points \( M \) and \( N \). Now to find \( V_o \) for \( V_i = 10mV \), say, then mark \( P_N \), draw \( P_N \) to \( L_2 \) and measure the resulting output voltage \( K_L \).

Point \( N \) could be described as an "apparent earth" since it appears to be at earth potential though is not connected to chassis earth. There is no physical location for \( N \) in the circuit, as it stands. But if \( R \) were to be made up of a series combination of resistors of suitably chosen values then \( N \) could be located at the junction point of two resistors in that chain.

**Ideal and practical op-amp**

Introductions to the op-amp inverter circuit, Fig. 1, often show \( \epsilon = 0 \), corresponding to \( A_0 = \infty \), with the advantage of simplifying circuit calculations. But some beginners find the infinite-gain (or "ideal" op-amp) approximation hard to swallow. The zero and infinity of the theoretical argument need some interpretation in practical electronics.

So how large must \( A_0 \) be (or, how small must \( \epsilon \) be) for \( \epsilon \) to be taken as zero without significant error in the calculation of\( L_A C_L (0) \)? A purely algebraic analysis will give an answer, but it can divert attention away from an appreciation of the basic operation of the circuit. This is where the pictorial approach comes into its own, providing some physical insight into the mechanism.

In Fig. 5 \( JM(\epsilon) \) is the phasor sum of two components: \( JQ (= aVi) \), the signal fed in from \( V_i \) with \( V_o = 0 \); and \( QM (= bV (= -IN)) \), the signal fed back from \( V_o \) for \( V_i = 0 \). \( \epsilon \) must be insignificant in comparison with the magnitude of each of its component parts for it to be neglected. As \( bV \) is (marginally) the smaller of the two components the condition is always met if \( \epsilon (= V/A_0) << hV \), or \( A_0 \gg 1 \).

This condition is independent of magnitude of the input signal amplitude and relates to the amplifier configuration as a whole, i.e. \( A \) plus resistor network, and not just \( A \) by itself. Hence, the inequality \( A \gg 1 \) – a plausible
prior assumption — is a necessary but not a sufficient condition.

For most practical op amps with $A_0 > 100,000$, the condition for $\varepsilon$ to be neglected reduces to $b > 10^{-5}$. In such cases, point $M$ in Fig. 5 is close to point $J$. The junction point of $R_1$ and $R_2$ is frequently referred to as a virtual earth but it can never be precisely zero for an amplifier with finite gain.

The dotted construction lines $LP'$ and $KP'$ refer to the ideal or limit case $A_0b = \infty$. For that case $J$, $M$ and $N$ coincide. Then, $aV_i + hV_0 = 0$,

$$\frac{|A(s)|}{A_0(s)} = (V_o/V_i)$$

The fractional error in $|A(s)|$ made in assuming $A_0b$ is infinite when it is finite (though perhaps large), is given by,

$$\frac{|A(s)|}{A_0(s)} = \frac{J}{M} = \frac{J}{M} = \frac{1}{A_0b}$$

In practice the near-equality in magnitude of the two opposing components $aV_i$ and $bV_0$ is an interesting parallel in solid-state physical electronics. For a forward-based PN junction the drift and diffusion tendencies within the junction region almost balance. The observed junction current is very small compared with either drift or diffusion current taken alone.

**Frequency response**

So far all discussion has been based on a frequency-independent voltage gain $A_0$. But for the frequently-encountered case of an op amp with a finite DC gain and a single-pole frequency response, dependence of gain $A(jf)$ on frequency ($f$) is:

$$A(jf) = \frac{A_0}{1 + f/f_3}$$

in which $A_0 = \text{DC gain}$ and $f_3$ is the $-3\text{dB}$ roll-off frequency. A Bode gain plot of this equation is shown in Fig. 6. If $|A(jf)| = 1 (0\text{dB})$ at $f = f_3 > 2f_2$ it follows that $f_3 = A_0 f_2$

Substituting in earlier equations to obtain a basis for a modified construction gives:

$$\frac{aV_i}{bV_0} = \frac{f}{f_0}$$

Using $f_2 = A_0 f_3$ and $f_3 = \frac{f}{f_3}$, it follows that $f_3 = A_0 f_2$

Interpreting this graphically calls for a $j$ axis for the quadrature component of $\varepsilon$. This can be drawn perpendicular to the plane of the page.
with the positive direction of \( j \) pointing upwards. Phasors for \( V_a, V_c \), and \( V_v \) can then be shown on planes erected, respectively, at \( \delta = \theta = R_1 \) and \( \gamma = (R_1 + R_2) \).

Looking first at the phasor plot drawn on a plane at \( \gamma = (R_1 + R_2) \) (Fig. 7a), \( a_1 \), is the component of \( V_v \) required to maintain \( \theta \), constant at \( \delta = \theta \). Relocating phasor \( e \) produces Fig. 7b which simplifies discussion.

\[ \theta = \text{the frequency-dependent phase shift due to } A \text{ and the feedback amplifier as a whole. Clearly, } \theta < \theta \text{ if } \delta \text{ increases, illustrating the benefit of phase shift reduction with negative feedback.} \]

The \(-3\text{dB}\) roll-off frequency, \( f_2 \), for closed-loop gain occurs when \( \theta = 45^\circ \). From Fig. 7b the condition \( \delta = \theta \) gives:

\[ \frac{a_1}{f_2} = \frac{a_1(2)}{f_2} \]

Now

\[ \delta = \frac{R_2}{R_1} = \frac{R_2 + R_1}{R_1} = \frac{a_1(2)}{f_2} \]

So for \( \delta = \theta \) \( \theta \gg 1 \)

\[ \frac{a_1(2)}{f_2} = \text{constant} \]

This is the gain-bandwidth product rule. Figure 8 is the three-dimensional equivalent of Fig. 2a for the case of frequency-dependent gain. In addition to the plane erected at \( \gamma = (R_1 + R_2) \) there is a plane at \( \gamma = 0 \) for the display of \( V_c \). A line, rather than a surface, is sufficient if \( \gamma = (R_1 + R_2) \) because \( V_v \) is fixed along the horizontal axis as \( f \) changes.

Just as the locus of \( T \) is a perpendicular erected at \( Q \), so the locus of \( W \) is a perpendicular erected at \( P \). \( HW \) is parallel to \( JT \) and the points \( K, T, W \) lie on a straight line.

Once the general significance of Fig. 8 is appreciated, it can be redrawn as the two-dimensional \((x,y)\) diagram resulting if the vertical planes are imagined pushed flat back onto the page. Then the \( y \) direction is used, as in Fig. 2a, for the display of \( R \) and also when the scale is completely unrelated for the quadratic components of the phasors.

Enhanced understanding

A pictorial approach can help in the introductory study of negative feedback amplifiers. The graphical technique helps particularly with op-amp circuits—emphasise the basic phasor balancing process inherent in the virtual-earth concept and enables some of the benefits of negative feedback to be deduced by visual inspection. Certainly, drawing Fig. 2a to scale for commonly available op-amps will soon convince the most doubting of students of the essential validity of infinite-gain approximation in practice.

Readers with some knowledge of circuit theory might like to reflect that in choosing to plot frequency-dependent input for a fixed output we have effectively converted an output pole into an input zero.
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200MHz FGAs. AT&T's AT73000 series of field-programmable gate arrays now run at 200MHz – a performance attributed to the use of the company's 0.6micron sidential cmos process. Devices toggling at 230MHz are expected in production early in 1993. AT&T Microelectronics, 0344 865927.

Asic libraries. The Liberty series of physical layout libraries and compilers from Compass allows a designer to choose the foundry and cmos process for production. As well as working with the Compass Navigator system, the series supports a variety of design tools such as Mentor Graphics, GenRad and Zycad. The library includes gate array and standard cells, ram and multiplier compilers and the Compass Datapath compiler. Compass Design Automation, 0908 661729.

0.8-micron gate arrays. Four new devices have joined Hitachi's H802G gate array family. Gate counts now reach 54200, 51100 and 70500, with i/o of 264, 288 and 336. The fourth device is for smaller applications and contains 10000 gates with 136 i/o pads. The H802G family offers sub nanosecond operation (0.3ns for a 2 input power Nano) and operates from 2.7V to 5.5V rails. Hitachi Europe Ltd, 021 643 6999.

Zero-drift op-amp. Linear claims its new LTC1250 chopper-stabilised op-amp to be the lowest-noise device of its type available. From 0.1Hz to 10Hz, noise is 0.655V/µpk-pk in the presence of a 4.2V output swing into 1kΩ. Sample-and-hold capacitors are on the chip. Particularly useful for bridge transducers, there is only 50nV/ºC offset drift and maximum offset is 10Vp-p. Linear Technology (UK) Ltd, 0207 677767.

Dual quad op-amps. LT1112/4 dual and quad op-amps from Micro Call are claimed to exhibit the lowest offset voltage of any such non-chopper stabilised amplifier available. Typical and maximum figures are 20µV and 70µV, both with a maximum drift of 0.5µV/ºC/µA. Input bias and offset currents are both 250pA. Noise performance of both devices between 0.1Hz and 10Hz is 0.32pV/√p-p and slew rate is 0.33V/µs. Micro Call Ltd, 0844 261939.

Triple video amplifier. Elantec's EL422G consists of three 60MHz current feedback amplifiers, each with a DC-restore amplifier activated by a common TTLcomparable control signal and each having a separate restore reference. Response is flat to within 0.1dB to 10MHz and slew rate is 200V/µs. Microelectronics Technology, 0844 278778.

Video distribution. A low-cost video distribution amplifier, the EL2099 from Elan-tec, will drive up to six double terminated cables at a -3dB bandwidth of 60MHz and a gain of 2, giving 1.1V to 250V and slew ing at 900V/µs. Output current is 500mA and differential phase and gain are said to be low. Microelectronics Technology, 0944 279771.

Digital signal processor
Histogrammer. Harris's HSPA4R410 is described as a histogrammer/accumulating buffer and is designed for extremely accurate histogram calculation and image contrast enhancement in machine vision systems, target recognition and medical imaging. It will evaluate contrast in images of up to 4096 by 4096 pixels and generate a histogram of input gray levels for manipulation or analysis, the data being used to modify or enhance the image. Harris Semiconductor (UK) Ltd, 0276 689896.

Lineaar integrated circuits
Video amplifier. Comlinear's CLC411 is a 200MHz video op-amp intended for HDTV, composite video, line driving and D-to-a output buffering. Slew rate is 2300V/µs with a settling time to 0.1% of 15ns. To 30MHz, gain varies less than 0.05dB and diff. gain and phase are within 0.02% and 0.03% respectively. The device has fast, break-before-make enable and disable. Joseph Electronics Ltd, 021 643 6999.

Optical devices
Bright, blue leds. Sharp's GL53X43 silicon carbide light-emitting diode has a luminous intensity of 16mCd at 20mA, emitting at a wavelength of 470nm and a bandwidth of 70nm. Sharp points out that RGB displays using leds are now possible. Sharp Electronics (Europe), 010 49 40 23 76 0.

Memory chips
4Mbit eproms. AMD has announced the Am27C4096 and Am27C400 devices, the former being organised as a 256K by 18bit type and the latter a 2Mbit cmos device user-configured as 512K by 8bit or 256K by 16bit eprom. Both are in 0.85p chip gate arrays and diff. gain and phase are within 0.02% and 0.03% respectively. The device has fast, break-before-make enable and disable. Joseph Electronics Ltd, 021 643 6999.

Logic building blocks
Character display. Two ICs from Philips, the PCA8510 and PCA8516, are designed to generate on screen characters for television displays and camcorders, being the first to incorporate software half-tone colour control to enhance readability. They will show a full screen of up to 13 or 40 lines of 36 or 42, 12 by 18 characters, depending on the television standard. Philips Semiconductors Ltd, 071 436 4144.

Mixed-signal ICs
Electronic digital pot. From Dallas, the...
Switching regulator. Two 200kHz current-mode off-line switching regulators by Linear provide 1% regulation and stabilisation with no opto-coupler for feedback. Power transmission and secondary sensing being done via the transformer. LT1005 has a totem-pole output to drive an external fet while LT1103 has its own fet output, the former being designed for 55W-250W output and the latter for 100W-1000W operation. Linear Technology (UK) Ltd, 0276 677676.

Passive components

Tantalum chip capacitors. Capacitors in Murata’s 227 range measure 1.2mm in height and have a range of values from 0.1µF to 6.8µF in voltage ratings of 4-20V DC. Leakage current is less than 0.4µA at 25°C. Murata Electronics (UK) Ltd, 0252 811666.

Connectors and cabling

2mm IDC connectors. Cambion has the 2630 series of insulated displacement sockets and headers in both straight and right-angle form for termination of 1mm 23awg ribbon cable. Current capacity is 1A per line. These tuning-fork contact sockets weigh 5.4g and take the cover on in 12 to 50-way versions and offer a contact resistance of 20mΩ and insulation resistance of 1GΩ. Interconnection Products Ltd, 0433 21555.

TOUGHC microcable. Internally strengthened microwave cables from Gore are small in diameter, low in weight and flexible, but withstand 175ºC straight line, which means they will take the weight of a forklift truck should the occasion arise. The cables have Gore-Tex expanded PTFE dielectric, giving a temperature range from ~-200°C to 200°C and a range of connectors is available. WL Gore & Associates (UK) Ltd, 0382 561511.

Filters

Chip EMI filter. Murata’s NFM51/52REMI filters offer suppression at 5GHz still being about 40dB. Murata Electronics (UK) Ltd, 0252 811666.

Hardware

Rectangular feedthrough. Possibly the first mass-produced rectangular feedthrough capacitor is produced by Beck. It measures 4.9 by 4.27mm and was designed to fit over the rectangular 30A pin of a car connector. Values of the ceramic capacitors is 1000µF at 100V DC. Beck Electronics Ltd, 0493 856282.

Product labels. Infomark and Appliance-mark by Donprint are label printing devices, using on-screen design and 300dpi printing on material capable of withstand temperatures up to 380°C and a variety of cleaning materials including petrol,degretergents and some fairly hostile chemicals. Donprint Label Systems Ltd, 0355 249191.

RF seals. Conductive, elastomerite materials for Anti-EMI sealing from Dunlop have been tested to US military and commercial spec and the new materials are based on silicon and fluorocarbon rubber with metallic or metal-coated fillers and reinforcements such as metalised fabric, which offers a 0.0012/µin resistivity. These materials will attenuate frequencies up to 18GHz by up to 100dB. Dunlop Precision Rubber, 0509 502151.

Intra-red pocket boxes. Hand held boxes designed to hold IR circuitry for signalling or remote control by OKW are said to be the only ones available with IP65 sealing and battery compartments. They are made in polycarbonate and ABS in three sizes and a belt clip is an option, as is machining and silk screening to order. OKW Enclosures Ltd, 0493 342626.

EMC shielding. Ecosheild is RFI shielding’s new material for the commercial sector where low compression force is needed — round doors, for example. Shielding is effective between 30MHz and 1GHz, providing around 30dB attenuation. The material consists of polyurethane foam coated by a layer of Monel wire, available in a variety of cross-sections, most being available with a pressure-sensitive adhesive backing. RFI Shielding Ltd, 0376 342626.

Instrumentation

Measuring receiver. ITT’s VX600S TV measuring receiver copes with satellite and FM ranges, meeting European standards for ally monitor, video modulation and video/spread separation. As video and sound quality monitor, it reproduces on a screen a zoomable picture, with the input signal and line trigger. It will also operate as a spectrum analyser. Feedback Instruments Ltd, 0892 635322.

EMC components. Global introduce a range of EMC antennas and custom made chokes and filters. Antennas cover the 20MHz-1GHz frequency range in biconical and log, periodic form and are supported by most automated EMC software. Global Specialties, 0978 853920.

Wider-range sig gen. Hewlett-Packard has expanded the frequency range of its HP70340A modular signal generator with the addition of a new module, the HP70340A/1A 10MHz-1GHz unit. Modulation is 60dB log AM,
10MHz peak FM deviation and pulse modulation with less than 10ns rise and fall times. Frequency resolution is 1kHz, or 1Hz as an option, Hewlett-Packard Ltd, 0344 362667.

Digital multimeter. Saje’s 7130 bench multimeter is programmed via RS232 or, as an option, IEEE488, to compare, maximum/minimum and off. All the separate instruments use the oscilloscope’s 3in LC screen and the whole thing is battery-powered. Thurlby-Thandar Instruments, 0480 412451.

Multi-function measurements. Sensortech from Thurlby-Thandar is a combined oscilloscope, counter-timer, data analyser and multimeter in one case, the whole costing £450. The dual-channel digital storage oscilloscope has a 20MHz bandwidth, sampling at 20Msamples/s, with a repetitive mode to 2.5ns resolution. There are ten waveform stores and a printer output. All the separate instruments use the oscilloscope’s 3in LC screen and the whole thing is battery-powered. Thurlby-Thandar Instruments, 0480 412451.

TV signal-level meter. Leader’s model 952 multi-channel television and satellite signal-level meter is programmed to cover VHF/UHF television and CATV as well as satellite channels. It can cover up to 128 channels simultaneously, with a bargraph display automatically scaling to respond to both smallest and largest signals in a group. Both sound and vision carriers are displayed and the instrument stores in non-volatile memory four sets of smallest and largest signals in any band being shown together. Thurlby-Thandar Ltd, 0480 412451.

Literature

Sensor handbook. Piezoresistive silicon pressure sensors and transmitters, with application notes on signal conditioning and interfacing, are all described in a new handbook from Sensortechnics in Puchheim, Germany. Sensortechnics GmbH, 010 49 69-60 66 30.

Materials

Ceramic HV capacitors. A range of capacitors from the Cera-Mite Corp. handle up to 13kV RMS at 50Hz, using a new ceramic material. Capacitance values are 400pF to 4700pF working at 10kV DC/4kV RMS at 60Hz. The material in the KT series confers improved reliability and stability, tighter tolerances and low dissipation. Acal Electronics Ltd, 0244 722727.

Power supplies

Programmable PSU. Thurlby-Thandar offer the TSP3222, a dual programmable power supply intended as both bench unit and as part of an ATE system. Output levels are both 0-32V, 0-5A, independently regulated to 300V. Both operate in constant-V or constant-I mode with automatic crossover and switching of the display from current to voltage. A GPIB interface is fitted and a LabWindows device is an option. Feedback Instruments Ltd, 0922 653352.

Memory protection. Double-layer capacitors from Surtech are small and light alternatives to the batteries usually used for memory protection, offering a reduced backup time and reduced board space. The AC300 or Ace Caps work in the –20°C to 70°C temperature range. Surtech Interconnection Ltd, 0256 51221.

Battery protection. Thurlby-Thandar offer the BMP – a bistable protector for rechargeable batteries. It guards against both thermal overload and short circuits, replacing the two devices normally used. Texas Instruments, 0234 223252.

Radio communications products

Linear amplifier. AML’s new APO150/1640 1.5-1 GHz linear power amplifier provides a 10W output with a gain of 22dB, flat to within ±0.2dB. A feature is gate-current monitoring to achieve enhanced reliability. European Microwave Components, 0376 515200.

Telemetry RX module. SILRFX-418-A is a s-variant UHF radio telemetry receiver module made by Radiometrix. It is meant for use with the TXM-418-A transmitter module in low data rate paging applications and is a PCB-mounted 418MHz receiver, needing only an antenna. The double-conversion FM superhet and data slicer driven by the AF output will drive a digital decoder for secure links. Maximum active-mode current is 15mA-130µA when standby.

Switches and relays

Low-noise relay. CRX-12X from BLG is a relay designed for the car industry, having a low level of acoustic noise (60dBA) so that intermittent windscreen wipers, for example, do not cause too much of a racket. It has silver alloy contacts and operate and release times of 6ms and 2ms respectively. BLP Components Ltd, 0638 665161.

Miniature RF relay. A 12V miniature PCB-mounted RF coaxial relay marketed by Cirkit will switch up to 50W CW at 1GHz, with a maximum insertion loss of 0.3dB at 1.8GHz and an SWR of 1:1.3 at 1.8GHz. Line impedance is 50Ω and contact form is single-pole changeover. Cirkit Distribution Ltd, 0922 444111.

PCB connector. A surface-mounting, dual-row reverse connector from Methods has an off-the-board height of 7.1mm. It comes in two forms: one with vertical holes and the other with horizontal holes for insertion from either side, both having from two to forty positions on each row at 0.1in pitch. Contacts handle up to 3A. Methodo Electronics Ltd, 0535 603282.

PCB switches. A new family of miniature switches by Elma, the Type 09, are meant for PCB-mouting or for through-panel use. Either toggle or push-button variants are made the toggle type with two or three positions and the button type in latching or momentary contact form, with the option of illumination. Radiatron Components Ltd, 081-891 1221.

Gyrostat is a miniature version of the company’s triangular piezoelectric vibratory instrument, which uses a prism to increase sensitivity by 100 times compared with a tuning-fork type. The triangular structure overcomes the vulnerability to vibration suffered by earlier designs. Bandwidth is 50Hz, the maximum angular velocity of ±50°/s producing a 72mV DC variation about the 2.5V reference. Murata Electronics (UK) Ltd, 0252 811666.

Position transducer. Rayevco’s miniature position transducers by Magnetek use the extension of a spring-loaded cable to drive the fixed sensor, the cable being capable of mounting in variable directions. Five models in the range cover 0-2in to 0.25in at ±0.15% of full scale maximum error, while withstanding 20g and 0-200°C. Powertronic International Ltd, 0438 759377.

Radio data link. A range of UHF communications equipment by Wood & Douglas is intended for remote outdoor use. The SurTel data link provides simplex or semi-duplex communication over a 20km line-of-sight range, internal modems transferring data at 1200 and 2400baud. The required supply is either 12V DC or mains and a standby circuit maintains only the oscillator in operation, thereby ensuring minimum frequency drift with a short power-up delay. Wood & Douglas, 0734 811444.
NEW PRODUCTS CLASSIFIED

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COMPUTER

Computer board level products

Data acquisition. Two boards from Amplicon, the C10-AD16/jr-AT, are for use with PC XT/AT and compatibles and are 100% compatible with the DAS16G board. One is a half-length 140mm board taking 300mA from the computer and having 16 single-ended or eight differential analogue inputs. The 3V5 converter providing and transferring at up to 100kHz. The **AT is three times faster and is able to use the Repeat Input String for 286+ computers for input/output at up to 330kHz. Amplicon Liveline Ltd. (Free)0800 525 335.

Software

Mathematical modelling. VisSim from Adept Scientific is an interactive, intuitive maths modelling program, running under Windows, for use on animated simulation, real-time control and dynamic analysis. It requires no knowledge of programming and is entirely graphic in presentation, including over 70 linear and non-linear function blocks to be “wired” together on screen to form graphical equations defining the process, at which point a mouse click runs the simulation. Adept Scientific Microsystems, 0402 480055.

Development and evaluation

PIC16CXX emulator. Running on PC 286 upwards, Microchip’s Flomaster universal in-circuit emulator now handles the company’s PIC16CXX range of microcontrollers, the industry’s only 8-bit risc family. It will support different family members by means of a probe card change. Arizona Microchip Technology, 0628 850303.

Universal programmer. BP-1200 from BP Microsystems Inc, is claimed to be the first programmer capable of programming and testing up to 240 devices in DIP, PLCC, LCC, QFP, PGA, SOIC and TSOP packages. Attention has been paid to the reduction of ground bounce, so that the units cope with fast clocks, PLDs and FPGAs. Direct Insight Ltd, 0455 538854.

Image compression. The JPEG image-compression board for PICs by C-Cube Microsystems is now available in Europe. It is an ISA half-card running at 10MHz and supporting grey scale, YLV (4:2:2 and 4:4:4), CMYK and 24-bit RGB. Data compression rate is more than 1Mbyte/s, allowing the unit to compress or decompress a 24-bit, 640 by 480 image in 0.7s on a 386DX. The still-image development kit includes a maths, PAL equations, BMP-compatible link library source and full documentation. Kudos Thame Ltd, 0734 351010.

Eeprom emulator. Opticron from Raisonance will emulate 8-bit Eeproms from the 27C116 to the 27C1080 with no hardware upgrading, and handles empor 16 and 32-bit mode up to 4 by 4Mbit using an add-on board. Data can be down-loaded from the serial port of a PC at 115Kbaud in binary. Intel-lex, Tektronix and Motorola formats over a high-speed opto-isolated RS232 link. Logicom Communications Ltd, 081 756 1284.

Turbo C debugger. ChipView-51 is Nohau’s new C source-level turbo C debugger for the company’s range of EMUL51-PC in-circuit emulators, compatible with the Borland Turbo Debugger. The unit provides up to 16K source lines using the standard trace board with a mix of detailed functions. Breakpoints can be to stop at the breakpoint, rather than at one op-code after it, as is usual. Over 150K of context-sensitive help is provided. Nohau UK Ltd, 0962 733140.

16-bit GPIB controller. Amplicon’s INES-AT IEEE 488.2 GPIB controller board performs all the IEEE talker, listener and controller functions to make an AT PC either the bus controller or appear as a listener. DMA transfer mode is automatic and transfer rate is 1Mbyte/s. A command interpreter supplied allows calls of the IEEE routines from any programming language and Windows 3 and above is supported by dynamic link libraries. Languages included are most of the Basics, Pascal, Forti, Fortran and C in their various quick and turbo guises. Amplicon Liveline Ltd. (Free)0800 525 335.

Computer peripherals

Paradise upgraded. Western Digital’s improved software drivers for its Paradise accelerator card for Windows delivers performance gains of up to 77% over the previous version 1.0. Utilities include drivers for Lotus 3.1, Microstation 4.03 and VersaCad 6.0. Current Paradise users will be able to obtain the drivers free. Western Digital (UK) Ltd, 0372 742955.
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March 1993 ELECTRONICS WORLD+WIRELESS WORLD 251
Fast – and safe – charging of NiCds with DSP

Cellular telephones and faxes, laptop computers, camcorders and similar fashion-accessories are lumped together in a note from Integrated Circuit Systems as “nomadic products” – a term worth a note in its own right. However, the piece is all about rapid NiCd charging without blowing up the NiCds in the process.

The company offers the ICSI700 rapid charge controller, which is an IC designed to charge batteries intelligently, using the Christie Electric Reflex principle to charge batteries in 20min-1h instead of many hours.

Conventionally, NiCds makers say their cells should be charged at the 10h rate, which means that 14h is usual, to take account of losses. Faster charging allows too little time for hydrogen/oxygen recombination and increases gas pressure to the point where the vent opens and electrolyte is lost. Some expensive batteries are around that will take a charge in about five hours. Memory effects, caused by partial discharging and full recharging reduces capacity. But it is restorable by several deep discharge/recharge cycles.

Several methods of rapid charging have been put forward, but most have drawbacks, mainly concerned with difficulties in determining the end-of-charge point.

In the Reflex method, a high charging current is interrupted once a second by negative current pulses, which strip accumulated oxygen bubbles from the plates and assist recombination. It has been shown that a high rate of charge increases charge acceptance, which reduces heat generation, which increases charge acceptance, and so on. In a paper referred to in the note, it is said that the Reflex principle allows rapid charging to full capacity at an efficiency of over 95% with low cell heating.

The IC decides when to stop charging in eight different ways, none of which depends on the battery being overcharged. Chief among these is inflexion-point measurement, which stops charging when overcharge is starting and before internal pressure rises.

Voltage is measured during the quiet period after the short discharge pulse. This accurately represents true charge state and, since there is no current flowing, is not distorted by internal resistance drops or plate surface charge. An infinite impulse-response filter eliminates random system noise and a linear regression algorithm obtains the best fit for the voltage samples. Cut-off can then be determined by comparing the first derivative with a reference slope. The algorithm is stored in the chip’s microcode rom.

In the event that unmatched cells are under charge, the inflexion point may be indeterminate, in which case the negative derivative termination stops charging at the maximum point of the charge curve.

When fully charged, rested batteries that have lost surface charge are put under charge and there is a rapid initial rise in voltage, which is detected, and charging stopped. If they have not rested long enough to lose the surface charge, they can produce a negative slope initially, which again is detected and charging stopped.

If a battery is of high impedance, or its contacts not making properly, this is also detected, charging stopped and a fault signal generated.

A consistent low charging voltage denotes a shorted cell or contacts. Charge stops and a fault signal produced.

Additionally, a timer and a thermal switch are present to stop charging after a preset time or in an over-temperature condition.

After normal charging is complete, a maintenance mode takes over, the same regime applying an equivalent average maintenance current to keep the battery free from dendritic formation and the plates in correct crystalline structure.

Amega Electronics Ltd, Armstrong Road, Daneshill East, Basingstoke, Hampshire RG24 0PF. Telephone 0256 843166.
Programmed delays

For applications such as multiple signal path de-skewing, programmable oscillators and pulse generators, delay detection and settling-time measurement, the Analog Devices AD9501 offers 10ps resolution at delays between 2.5ns, up to 50MHz. Figure 1 is its internal block diagram. A positive-going input pulse triggers the ramp generator, whose output is compared with the output of a D-to-A converter, the inputs of which are set by the user; timing diagram Fig. 2 shows what happens. Ramp slope is set by external RC, which governs maximum available delay, actual delay being the total of two circuit delays – trigger and ramp delays – and that programmed by the user. The ramp resets itself, returns past the zero reference and settles before a new cycle can start, the time between the comparator's being triggered and reset being the output pulse width. Data is held while Latch is high; when this pin is low, the D-to-A follows the inputs if the relevant timing is observed.

De-skewing is accomplished by the circuit of Fig. 3. When signal paths in parallel carry high-speed data, delay matching must be precise, but can be difficult to achieve with varying lead lengths and impedance changes. The skew is removed by this circuit, in which one stimulus is applied to all AD9501s in use, delays for each path measured and adjusted by the digital inputs.

An oscillator with programmed frequency and duty cycle is made as in Fig. 4. Changing programmed delay in each AD9501 changes frequency and duty cycle, frequency being f = 1/(2tps1 + tps2 + tP0), tps1, tps2 being the two programmed delays and tP0 the minimum propagation delay.

The random pulse generator of Fig. 5 again uses two AD9501s, triggered together.
and driving an RS flip-flop. If the microprocessor bus varies the digital from clock to clock, a pulse train with varying duty cycle and pulse width, is produced.

To measure an unknown delay, use the circuit of Fig. 6, which is a little like a successive-approximation A-to-D converter, except that the flip-flop is used instead of the A-to-D’s comparator. In calibration, short out the unknown delay with clock at both AD9501/s, the top one being programmed for a delay longer than the zero-set programmed delay of the lower one, which is done by incrementing the data into the top delay generator until the SAR outputs 02H or more. Delay through the top generator is now a little longer than through the other, so that the SAR output is the reference for measuring when it is reinstated. All this compensates for stray delays and setup times.

**Filtering reference voltages**

Although the Burr-Brown REF102 buried-zener 10V reference lays claim to better stability and five times lower noise than a bandgap reference, noise is still around 600μV pk-pk at a noise bandwidth of 1MHz. As is pointed out in Application Bulletin AB-003, Vol.1, filters and buffers go some way towards reducing noise and its bandwidth, but not far enough in many applications.

Figure 1 shows the usual sort of thing -- a single-pole filter and an op-amp buffer. One problem with this is that capacitor leakage current goes through $R_2$ and is variable with temperature, particularly in large capacitors needed for this job. You then have a DC error, which will drift. Then, again, the buffer puts its penn’orth of noise in, over its full unity-gain bandwidth, so even if the filter output is silent as the grave, unacceptable noise still appears at the circuit output.

To solve both problems at a stroke, use the circuit of Fig. 2. The filter is now at the output of the buffer, where its -3dB point is $2RC_1$ (reducing noise bandwidth by, say, 100 reduces noise by10). The $R_2D_2$ arrangement maintains stability and $R_2D_2$ should equal $2RC_1$ to escape amplifier noise gain peaks. Resistor $R_2$ should be kept fairly low, since it takes bias current and could cause DC error and noise; $R_2$ should also be low, since it takes load current, its volts drop increasing the required output swing. It should drop less than 1V full load.

Since the filter is now in the feedback loop, leakage current volts drop across $R_1$ is divided by the loop gain, the DC output impedance is very low and the voltage across $C_1$ is almost nothing, giving rise to negligible leakage current. When driving large capacitive loads, $(CLOAD+C_1)R_1$ must be less than $0.5R_1C_1$.

**Analog Devices Ltd, Station Avenue, Walton-on-Thames, Surrey KT12 1PF. Tel: 09322 232222**

**Burr-Brown International Ltd, 1 Millfield House, Woodshots Meadow, Watford, Hertfordshire, WD1 8YX Telephone 0923 233837**

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**Fig. 1.** Obvious reference voltage filters has its drawbacks – DC error that varies with temperature and noise from the buffer.

**Fig. 2.** Improved filter avoids both problems of Fig. 1. Filter reduces noise both from reference and op-amp and output impedance is low over most of frequency range. Leakage current from $C_1$ is no longer a problem. Peak in output impedance near filter pole frequency of about 35 is reduced by reducing $R_2$ and increasing $C_1$ – peak is $0.7R_2$.  

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**Applications**

**CLOCK IN**

**Trigger AD9501 #1**

**Digital Data Output**

**Latch**

**Reset**

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**CLEC IN**

**Decoder**

**Trigger AD9501 #2**

**Digital Data Output**

**Latch**

**Reset**

---

**Unknown Delay**

**LATCH**

**AD9501**

**Digital Data Output**

**Trigger**

**Reset**

---

**Fig. 5**

**Fig. 6**
INTERFACING WITH C
by HOWARD HUTCHINGS

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This is a practical guide to real-time programming, the programs provided having been tested and proved. It is a distillation of the teaching of computer-assisted engineering at Humberside Polytechnic, at which Dr Hutchings is a senior lecturer.
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INSTRUMENTS TO BUY

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SC-130 and SC-40 are full featured, microprocessor-based, handheld frequency counters providing portability and high performance. Both instruments provide measurements of frequency, period, count and RPM plus a view facility enabling min, max, ave and difference readings:
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187: As 185 except auto ranging. £39.50 plus VAT (£46.61).
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NEGATIVE approach to POSITIVE thinking

Negative characteristic components may not be called for every day, but Ian Hickman shows why they should be regarded as a standard tool for the professional circuit designer.

There is often felt to be something odd about negative components. The circuit designer in the development labs of a large firm knows there will be no obstacle to going along to stores to draw a dozen 100k resistors or half a dozen 10µF tantalums for example. But however handy it would be, drawing a -4.7k resistor would be a different problem.

Yet negative resistors would be so useful; for example when using mismatch pads to bridge the interfaces between two systems with different characteristic impedances. Even when the difference is not very great, eg testing a 75Ω bandpass filter using a 50Ω network analyser, the loss associated with each pad is around 6dB, immediately cutting 12dB off measurements into the stopband. But a few negative resistors from the junk box could make a pair of mismatch pads with 0dB insertion loss each.

In circuit design, negative component values do turn up from time to time and the experienced designer knows when to accommodate them, and when to redesign to avoid them. For example, a filter may call for a -3pF capacitor, say, added between nodes X and Y. Provided that an earlier stage of the computation has resulted in a capacitance of more than this value appearing between those nodes, there is no problem; it is simply reduced by 3pF to give the final value. But where the final value is still negative, redesign may be necessary to avoid the problem, particularly at UHF and above. Lower frequencies allow the option of using a "real" negative capacitor (or something that behaves exactly like one), easily implemented with an ordinary (positive) capacitor and an op-amp or two, as are negative resistors and inductors.

But before looking at negative components using active devices, they can be implemented in entirely passive circuits – if you know how.

I first came across this some time ago. Figure 1a shows a parallel tuned circuit in series with a signal path – to act as a trap, notch or rejector circuit. Clearly it only works well if the load resistance \( R_L \) is low compared with the tuned circuit’s dynamic impedance \( R_d \). If \( R_L \) is near infinite, the trap makes no difference, so \( R_d \) should be much greater than \( R_L \). Indeed the ideal would be to make \( R_d \) infinite by using an inductor (and capacitor) with infinite \( Q \). An equally effective ploy would be to connect a resistance of \(-R_d\) in parallel with the capacitor, exactly cancelling out the coil’s loss and effectively raising \( Q \) to infinity. This is quite easily done, as in Fig. 1b, where the capacitor has been split in two, and the tuned circuit’s dynamic resistance \( R_d \) \((R_d = QoL_c, assuming the capacitor is perfect) replaced by an equivalent series loss component \( r \) associated with the coil (\( r = Qo/L_c \)). From
DESIGN BRIEF

the junction of the two capacitors, a resistor $R$ has been connected to ground. This forms a star network with the two capacitors, and must be transformed to a delta network using the star-delta equivalence formula. The result is as in Fig. 1c and the circuit can now provide a deep notch even if $R_0$ is infinite, owing to the presence of the shunt impedance $Z_s$ across the output, if the right value is chosen for $R$. So, let $R' = -r$, making the resistive component of $Z_s$ (in parallel form) equal to $-R_r$. Now $R'$ turns out to be $-1/(4\omega C R)$ and equating this to $-r$ gives $R = R_r/4$.

Negative inductor

Now for a negative inductor – and all entirely passive, not an op-amp in sight. Figure 2a shows a section of a constant-$K$ lowpass filter acting as a lumped passive delay line. It provides a group delay $dB/d\omega$ of $\sqrt{LC}$ seconds per section. Fig. 2b, at DC and low frequencies, maintained fairly constant over much of the pass band of the filter.

A constant group delay (also known as envelope delay) means that all frequency components passing through the delay line (or through a filter of any sort) emerge at the same time as each other at the far end, implying that the phase delay $\phi = \omega(\sqrt{LC})$ radians per section is proportional to frequency. Thus a complex waveform such as an AM signal with 100% modulation will emerge unscathed, with its envelope delayed but otherwise preserved unchanged. Similarly, a square-wave will be undistorted provided all the significant harmonics lie within the range of frequencies for which a filter exhibits a constant group delay. Constant group delay is thus particularly important for an IF bandpass filter handling phase modulated signals.

Connecting an inductance $L'$ (of suitable value) in series with each of the shunt capacitors, will cause the line to become an "$m$-derived" low pass filter instead of a constant-$K$ filter. The result is that the increase of attenuation beyond the cut-off frequency is much more rapid. But that is of no great benefit in this application. A delay line is desired above all to provide a constant group delay over a given bandwidth and the variation in group delay of an $m$-derived filter is much worse even than that of a constant-$K$ type.

$L'$ may not be a separate physical component at all. It could be due to mutual coupling between adjacent sections of series inductance, often wound one after the other, between tapping points on a cylindrical former in one long continuous winding. If the presence of shunt inductive components $L'$ makes matters worse than the constant-$K$ case, addition of negative $L'$ improves matters. This is easily arranged. Fig. c, simply by winding each series section of inductance in the opposite sense to the previous one.

Real pictures

To picture negative components that may seem more "real" implemented using active circuitry, imagine connecting the output of an adjustable power supply to a $1\Omega$ resistor whose other end, like that of the supply's return lead, is connected to ground. Then for every volt positive (or negative) applied to the resistor, 1A will flow into (or out of) it. Without changing the supply's connections, arrange that the previously earthy end of the resistor is automatically jubbed up to twice the power supply output voltage – whatever that happens to be. The voltage across the resistor is always equal to the power supply output voltage, but of the opposite polarity. So when, previously, current flowed into the resistor, it now supplies an output current, and vice versa. With
the current always of the wrong sign. Ohm's law will still hold if the value of the resistor is labelled as $-1\Omega$.

**Figure 3** shows the scheme, this time put to use to provide a capacitance of $-C_3\mu F$, and clearly substituting $L$ for $C$ will give a negative inductance.

For a constant applied AC voltage, a negative inductance will draw a current leading by 90° like a capacitor, rather than lagging like a positive inductor. But like a positive inductor, its impedance will still rise with frequency. Figure 3 also shows how a negative component can be balanced, or even floating. Clearly, if in Fig. 3a, $C$ is 99pF and the circuit is connected in parallel with a 100pF capacitor, 99% of the current that would have been drawn from an AC source in parallel with the 100pF capacitor will now be supplied by the op-amp via $C$, leaving the source "seeing" only 1pF. Equally, if the circuit is connected in parallel with an impedance which, at some frequency, is higher than the reactance of $C$, the circuit will oscillate; the circuit is "short circuit stable".

**Negative capacitance**

A negative capacitance can be used to exterminate an unwanted positive capacitance – useful in applications where stray capacitance is deleterious to performance yet unavoidable. A good example is the N-path (commutating) bandpass filter. Far from being an academic curiosity as some suppose, this has been used both in commercial applications, such as FSK modems for the HF band, and in military applications.

One of its disadvantages is that the output waveform is a fairly crude, N-step approximation to the input, N being typically 4, requiring a good post filter to clean things up. But on the other hand, it offers exceptional values of $Q$. Figure 4a illustrates the basic scheme, using a first-order section.

Apply at $V_1$ a sinusoidal input at exactly a quarter of the clock frequency (Fig. 4a), so that the right hand switch closes for a quarter of a cycle, spanning the negative peak of the input, and the switch second from left acting similarly on the positive peak.

The capacitors will charge up so that $V_o$ is a stepwise approximation to a sinewave, as in Fig. 4b, bottom left. The time-constant will be not $CR$ but $3CR$, since each capacitor is connected via the resistor to the input for only 5% of the time. If the frequency of the input sinewave differs from $F_{clock}/4$ (either above or below)
by an amount less than 1/(πfCR), the filter will be able to pass it. But if the frequency offset is greater, then the output will be attenuated, as shown in Fig. 4a. Depending on the devices used to implement the filter, particularly the switches, f_{\text{switch}} could be as high as tens of kHz, whereas C and R could be as large as 10µF and 10MΩ, giving (in principle) a Q of over ten million.

**Kundert filter**

The same scheme can be applied to a Kundert filter section, giving a four pole bandpass (two pole LPE - lowpass equivalent) section, Figs 4c and 4d. Figure 5a shows the response of a five pole LPE 0.5dB ripple Chebychev N-path filter based on a Salen and Key lowpass prototype, with a 100Hz bandwidth centred on 5kHz.

The 6 to 60dB shape factor is under 3:1 with an ultimate rejection of well over 80dB. However, the weak point in this type of filter is stray capacitance across each group of switched capacitors. The stray causes “smearing” of charge from one capacitor into the next. In high Q second order sections this has the effect of slightly lowering the frequency of the two peaks and also of unbalancing their amplitude. The higher the centre frequency, smaller the value of switched capacitors, narrower the bandwidth or higher the section Q, the more pronounced is the effect. The result is a crowding together of the peaks of the response at the higher frequency side of the passband and a spreading of them further apart on the lower, producing a slope up across the passband (Fig. 5a), amounting in this case to 1dB. Increasing the clock frequency to a 20kHz centre frequency results in a severely degraded passband shape, due to the effect.

Changing the second order stage to the Kundert circuit, Fig. 5b, improves matters by permitting the use of larger capacitors; C can be as large as C_{\text{T}} in the Kundert circuit whereas in the Salen and Key circuit, the ratio is defined by the desired stage Q. With this modification, the filter’s response is as in Fig. 5b.

The modification restores the correct response of the high Q pole output section, but the downward shift of the peaks provided by the three-pole input section results in a downward overall passband slope with increasing frequency. Note the absence of any pip in the centre of the passband due to switching frequency breakthrough.

If the charge injection via each of the switches was identical, there would be no centre frequency component, only a component at four times the centre frequency, ie at the switching frequency. Special measures, not described here, are available to reduce the switching frequency breakthrough. Without these, the usable dynamic range of an N-path filter may be limited to as little as 40dB or less.

With them the breakthrough is reduced to -90dBV.

Figure 5b was recorded after the adjustment had been made. The slope across the passband is shown in greater detail in Fig. 5c (lower trace), recorded before the adjustment, the centre frequency breakthrough providing a convenient “birdie marker” indicating the exact centre of the passband. Upper trace shows the result of connecting -39pF to ground from point C, 1dB/div vertical, 2kHz/div horizontal. Note: the gain was unchanged; the traces have been separated vertically for clarity.

Fig. 5b. The response of a five pole LPE 0.5dB ripple Chebychev N-path filter based on a Salen and Key lowpass prototype, with a 100Hz bandwidth centred on 5kHz, 10dB/div vertical, 50Hz/div horizontal. (At a 20kHz centre frequency, its performance was grossly degraded.)

b) A five pole LPE Chebychev N-path filter with a 100Hz bandwidth centred on 20kHz, using the Kundert circuit for the two pole stage, and its response (10dB and 1dB/div vertical, 50Hz/div horizontal).

c) The passband of b) in more detail, with (upper trace) and without -39pF to ground from point C. 1dB/div vertical, 2kHz/div horizontal. Note: the gain was unchanged; the traces have been separated vertically for clarity.

d) The passband of b) in more detail, with -39pF (upper trace) and with -100pF to ground from point C; over compensation reverses the slope.

**Optoisolator correction**

In the Design Brief “Bringing the optoisolator into line” (EW + WW December 1992 pp. 1050 - 1051), Fig. 6a also appeared in error in place of Figs. 4 and 8. The correct Fig. 4 was rather similar to Fig. 3 except that the residual contained a substantial component at the fundamental in addition to the second harmonic content. Figure 8 was also rather similar to Fig. 3 except that the residual was larger and distinctly more triangular in form.
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