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REGULARS

In next month's issue: There is currently much interest in direct conversion SSB radio systems because of their ease of integration onto silicon. N C Hamilton describes an SSB HF receiver using quadrature direct conversion for use with either analogue or DSP design techniques.

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The nature of power

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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 2.5GHz 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.

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UPDATE

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.

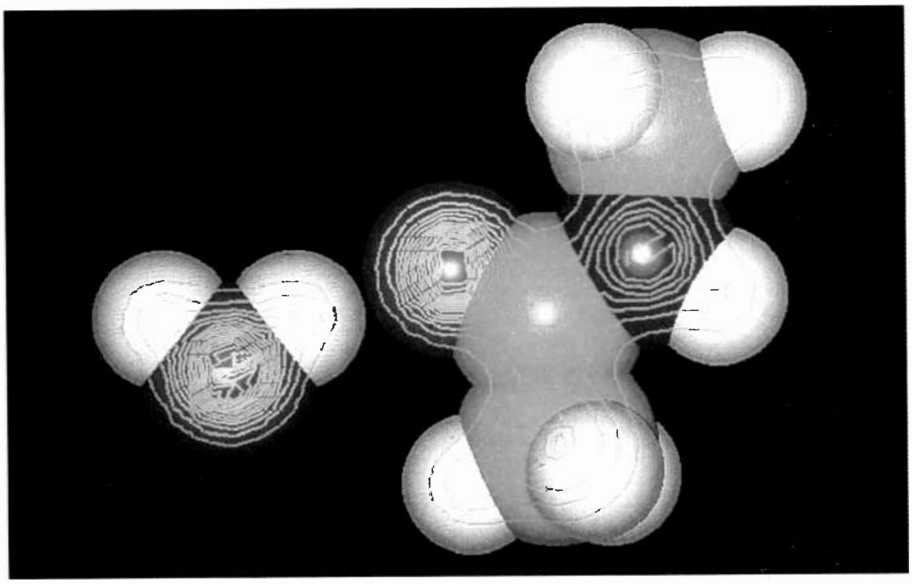
GSM has been slow to take off in Europe because the existing analogue services are too successful, so manufacturers have been looking to export. The name was changed in 1990, to Global System for Mobile Communications, to make this easier.

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

Welcome to Cyberspace: Computer aided design will never be the same again as Autodesk redefines the traditional interface between human and computer.

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. Well 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. Autodesk, the US company famous for its AutoCad design program, was also one of the first to get into VR and created an environment called Cyberspace. This will become available as a development kit later this year, as Autodesk bids to set the *de facto* standard for future VR systems. The kit will come out in the US initially and will not be available in the UK until at least next year.

VR for the home might be slightly longer in coming. Yet Sega promised a virtual game in 1992 which has yet to appear and Nintendo is known to have a top-secret VR project. This particular seam of gold has not been mined yet, but the digging has started.

Dom Pancucci

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 mm^2 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 analogue microwave and optical functions in its new wafer fab; it has established a reputation for manufacturing advanced RF functions such as DBS amplifiers and downconverters, phased array radar functions and discrete microwave devices made with class III/V semiconductor. However, it doesn't expect to make pure logic except possibly as a foundry exercise for another vendor.

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

GEC Plessey's Caswell GaAs wafer production line for high volume microwave devices. Plant manager Fred Myers claims device yields up to 80 per cent on its 3in wafers. The picture shows molecular beam epitaxy equipment used to grow further semiconductor layers on the GaAs wafers.

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 70mile/h 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."

Pushing back the limits on optic fibre comms

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 frequenc·· that can be regenerated in this way.

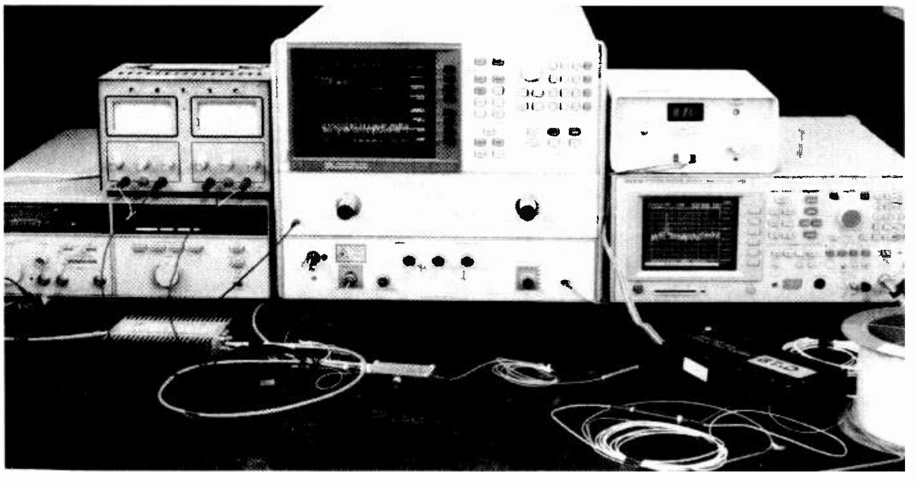
An alternative method, in theory, would be to send two coherent CW laser beams along the same optic fibre and then mix them at the far end to obtain millimetre-wave RF by heterodyning: the necessary stability can certainly be achieved by frequency-locking the two lasers to the required frequency separation.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

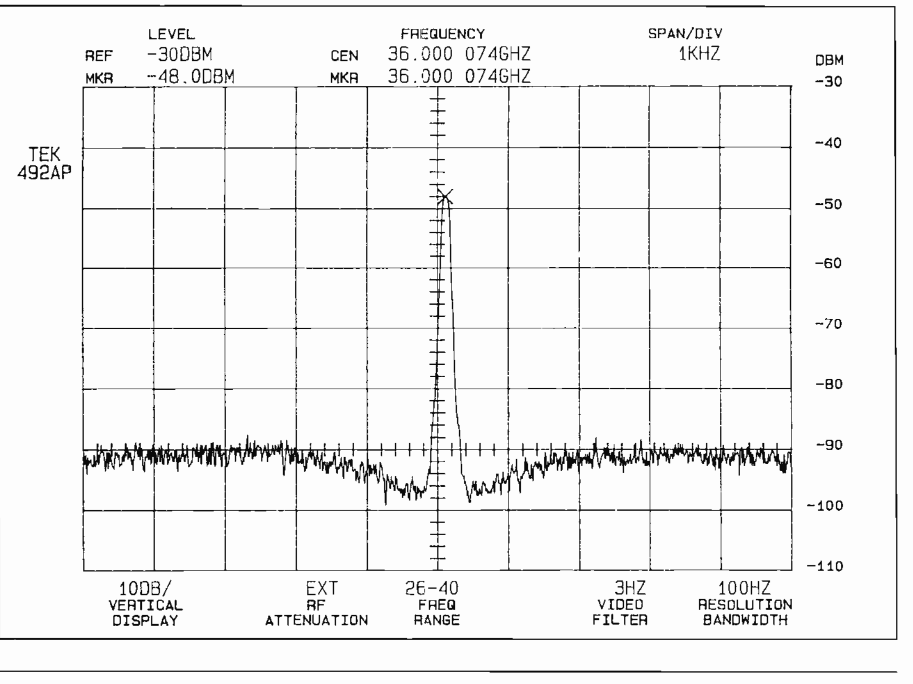
Benefits of the technique do not stop there. Amplification of the optical signals in an erbium-doped fibre amplifier (EDFA) make it possible to regenerate a millimetre-wave RF signal at a relatively long distance from its source. The experimenters tried sending the optical components along 8km. of fibre 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.



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.

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



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.

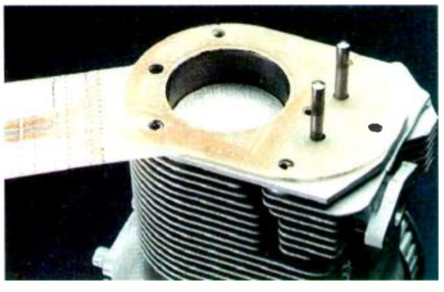
Ionisation probes built into the instrumented gasket are merely exposed conductors carrying a voltage. As soon as a flame contacts the probe, a distinct signal is produced.

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 bogey 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.

Sandia's ion probe flame detector gasket has been successfully tested at General Motors test facilities, Michigan, and is now in manufacture on a small scale.



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 either 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.

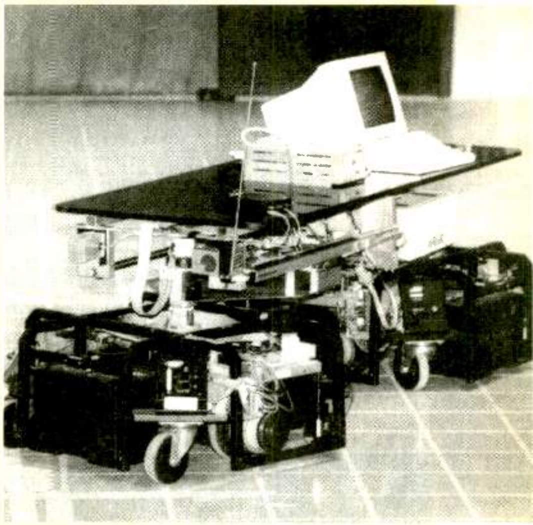
Main problem is that because mechanical components can not react as fast as the computer commands, it is impossible to eliminate all errors in speed and direction. Any practical motor system will always undershoot or overshoot the desired movement.

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.



A dual chassis system keeps a tight leash on any wandering robot.

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, 1669), 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.

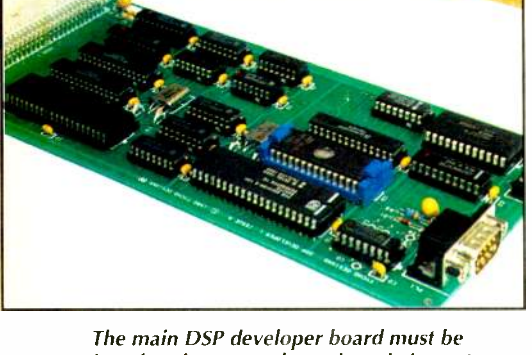
Volunteers participated in experiments to determine how accurately they could recall the loudness of a tone by comparing their memory of it with a "probe" tone played a second or more later. The subjects pressed one computer key if the probe tone seemed louder than the memory tone and another key if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

STARTING OUT WITH DSP

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 main DSP developer board must be interfaced to an analogue board via a 64-way back-plane.

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

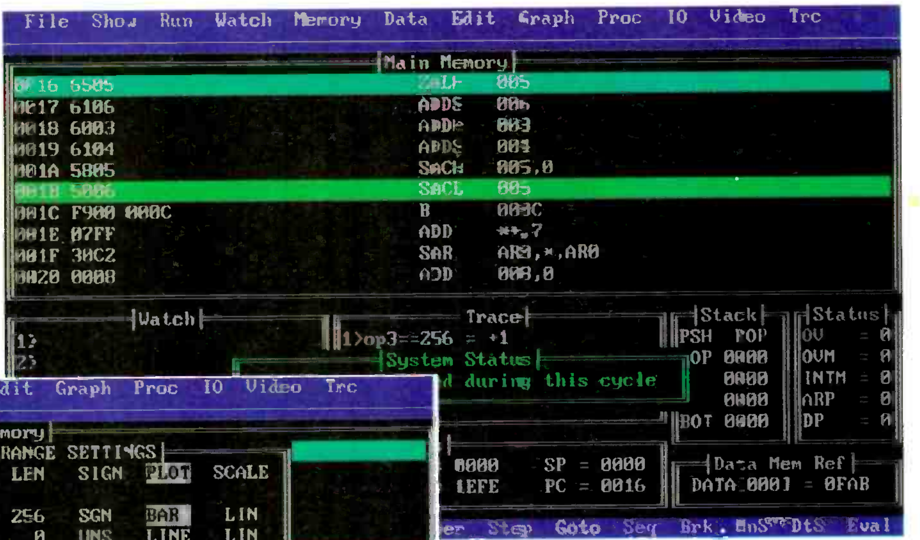
The developer is based on the *TMS320C10* running at 20MHz. An *80C31* microcontroller is used as a master processor to the DSP; its serial port is used to connect to an IBM PC. A software package on the PC communicates to the DSP developer which allows DSP programs to be download to the DSP program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

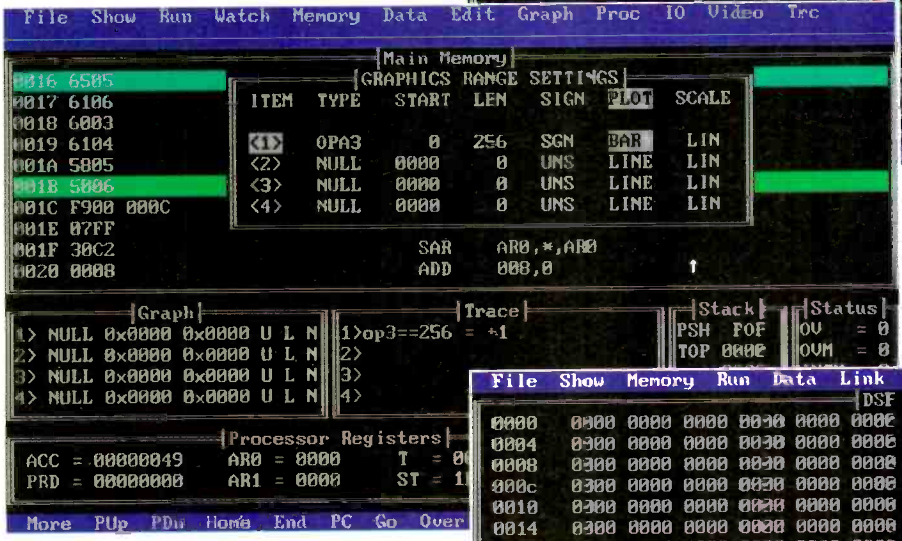
The *TMS320C10* is a first generation DSP chip which was introduced in 1983 by TI. It is an old design but for the purpose of this exercise, the device is perfectly adequate for a demonstration of what DSP can be used for, and it is easy to use.

The DSP developer is constructed on a 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 type 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.

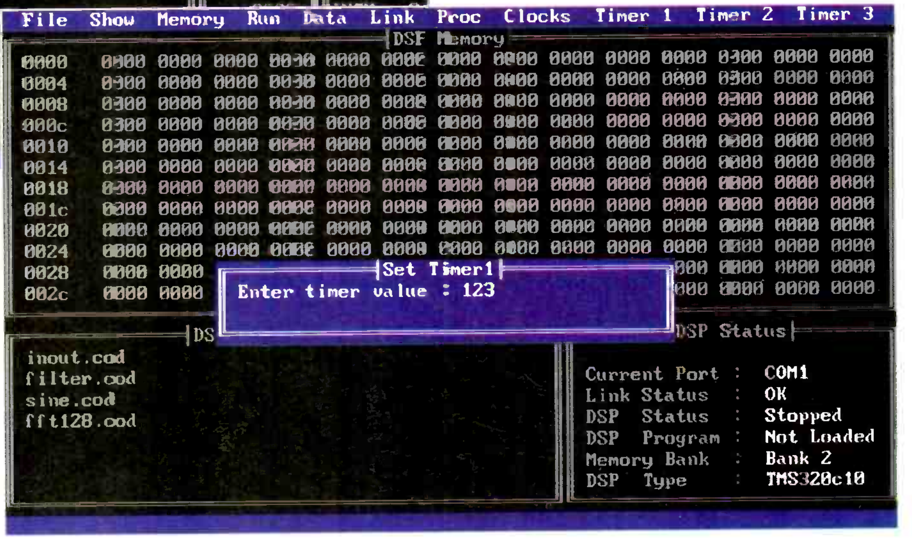
This figure shows the main simulator screen. The large window across the middle of the screen is the memory window. The light blue horizontal cursor shows the location of the program counter. The green cursor is used to set breakpoints. In this example, a tracepoint has forced the simulator to halt.

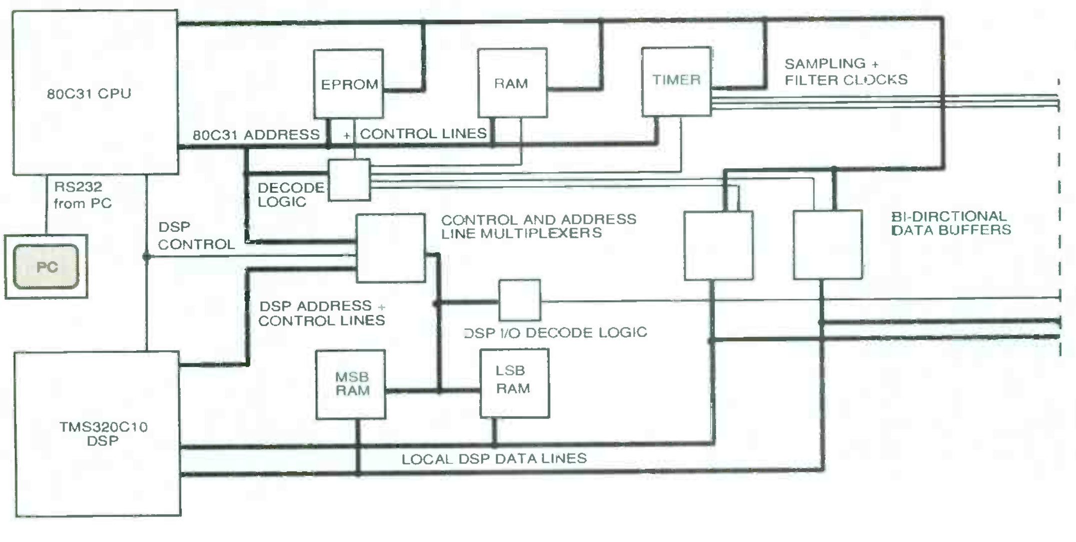




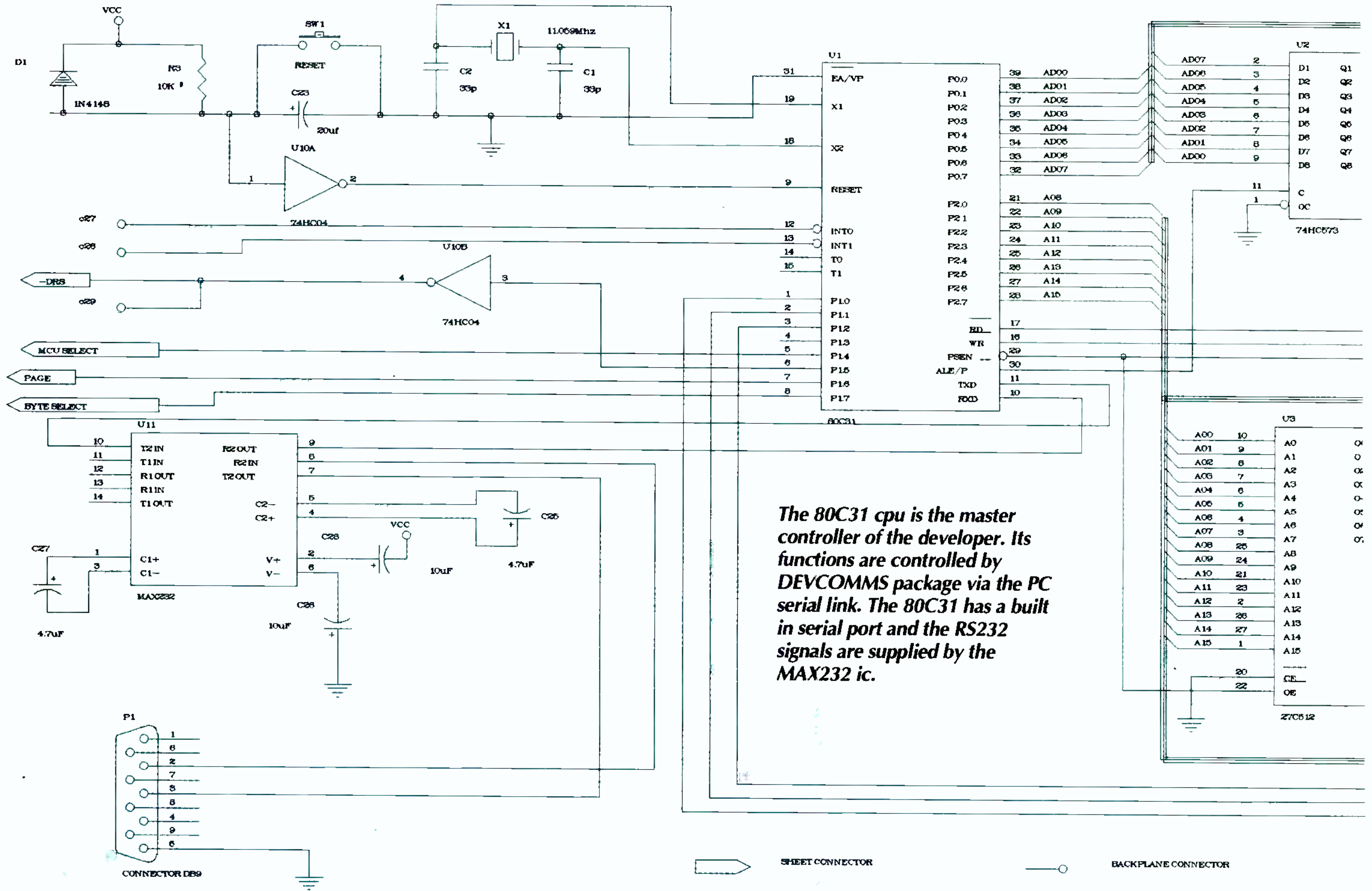
This figure above shows the simulator graph layout window popped up over the main screen. Up to four traces may be plotted, only one has been defined here. Main memory, data memory, input or output ports may be graphed with any start position and trace length. Number representation may be either signed 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 fitering. DEVCOMMS provides the facility to set, start and stop any of the timers.

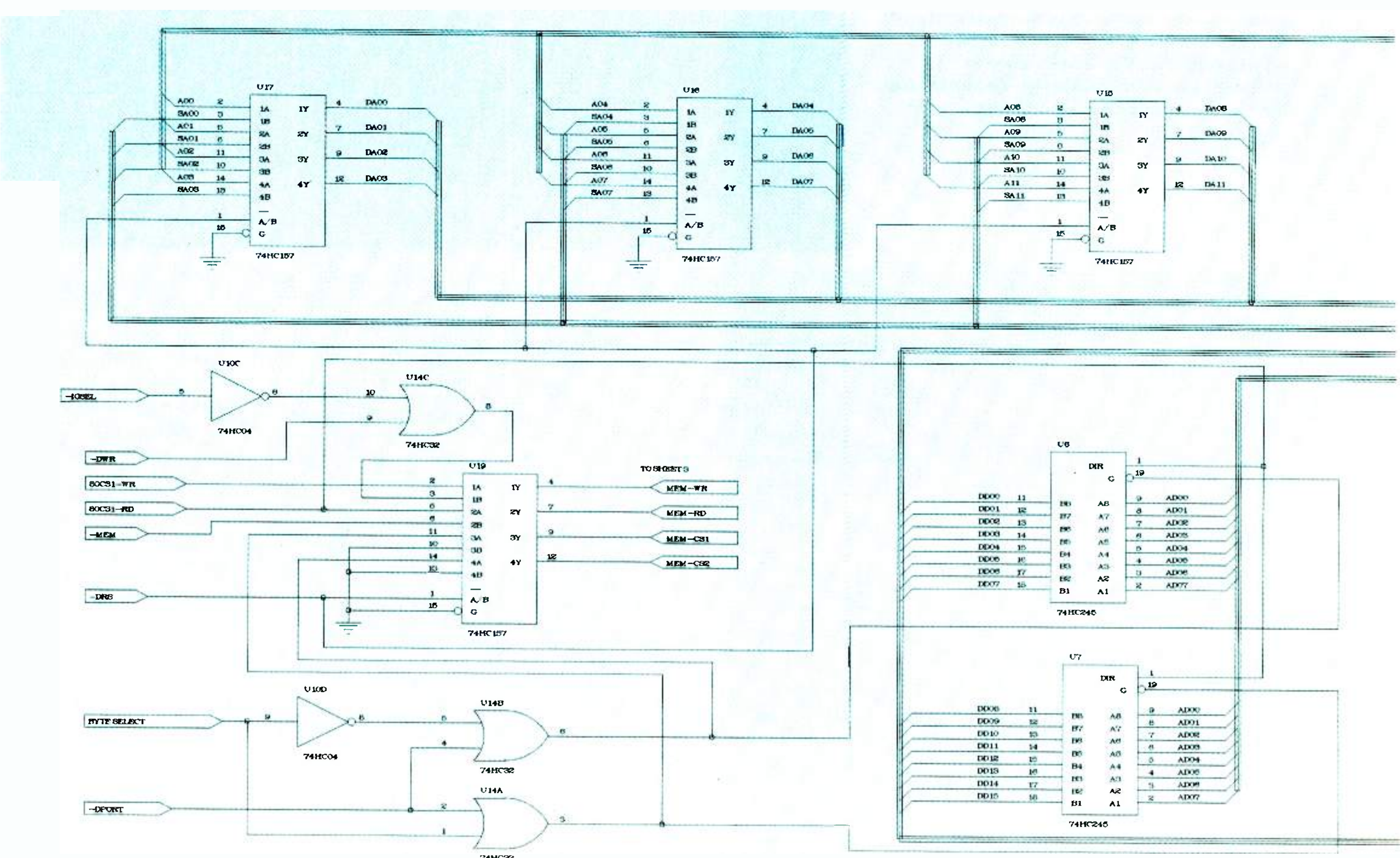


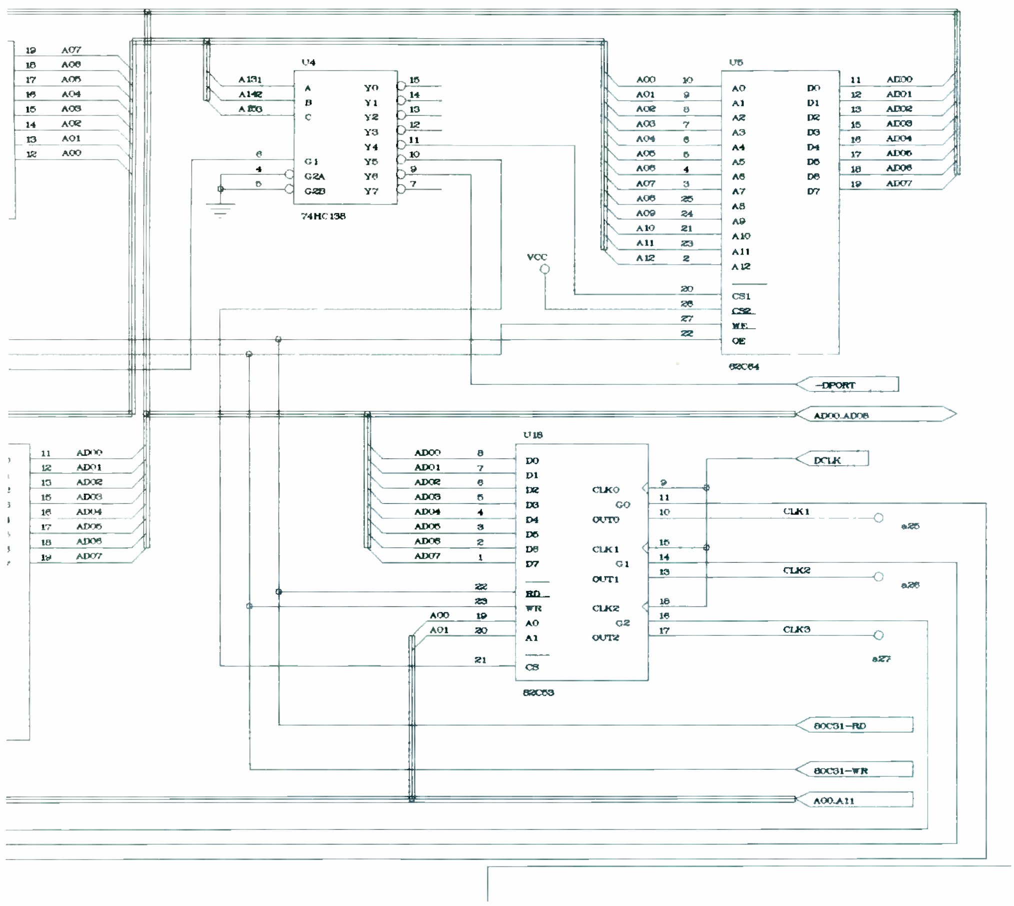


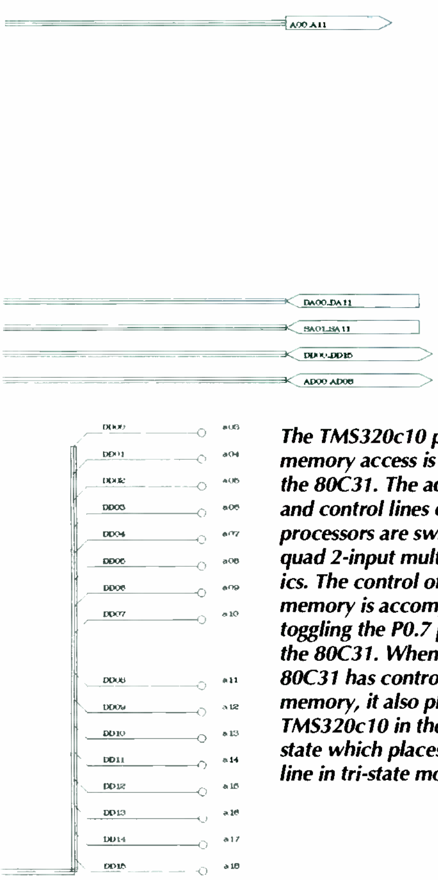
The block diagram of the developer shows that low cost DSP development system can be acheived with minimal chip count. The use of the 80C31 microcontroller as the master controller provides high intergration and interfacing to the PC via a serial link allows user flexibly to experiment without the risk of damage to the PC.



The 80C31 cpu is the master controller of the developer. Its functions are controlled by 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.







The TMS320c10 program memory access is control by the 80C31. The address lines and control lines of both processors are switched via quad 2-input multiplexers ics. The control of the memory is accomplished by toggling the P0.7 port line of the 80C31. When the 80C31 has control of the memory, it also places the TMS320c10 in the reset state which places its data line in tri-state mode.

Circuit description

The *320C10* and the address/control multiplexers (U_{17},U_{16},U_{15} and U_{19}) 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 U_{10b}. 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 (U_6 and U_7). Port P1.7 and the inverter (IC_6) 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 ics. The *80C31* port P1.6 is connected to A12 pin of U_8 and U_9 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 U_{13}, U_{14} and U_{10} 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.

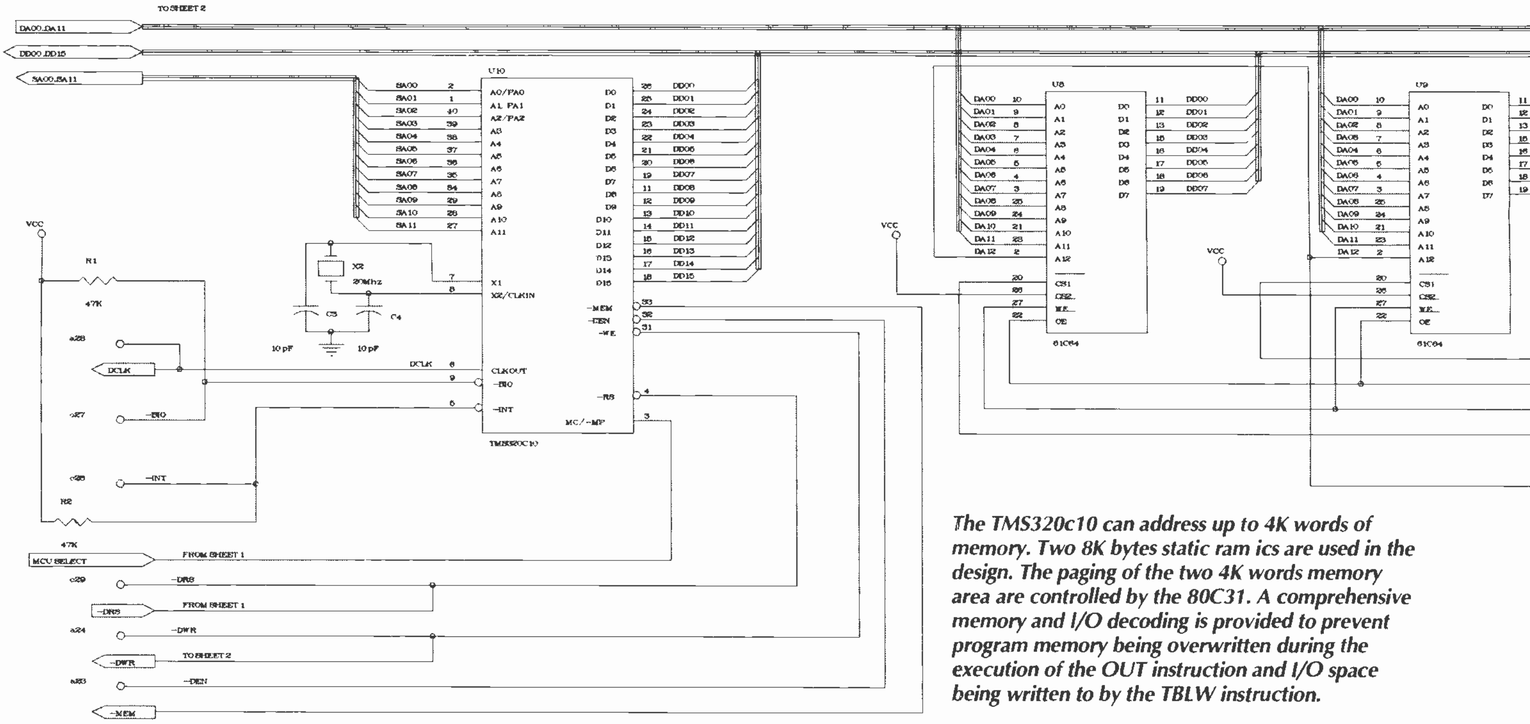
The output of U_{14D}, /IOSEL signal, goes low when address lines DA03 to DA11 go low. This signal must be used together with PA0-PA2 signals to decode i/o on any boards plugged onto the bus.

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 prototcol 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.



The TMS320c10 can address up to 4K words of memory. Two 8K bytes static ram ics are used in the design. The paging of the two 4K words memory area are controlled by the 80C31. A comprehensive memory and I/O decoding is provided to prevent program memory being overwritten during the execution of the OUT instruction and I/O space being written to by the TBLW instruction.

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.

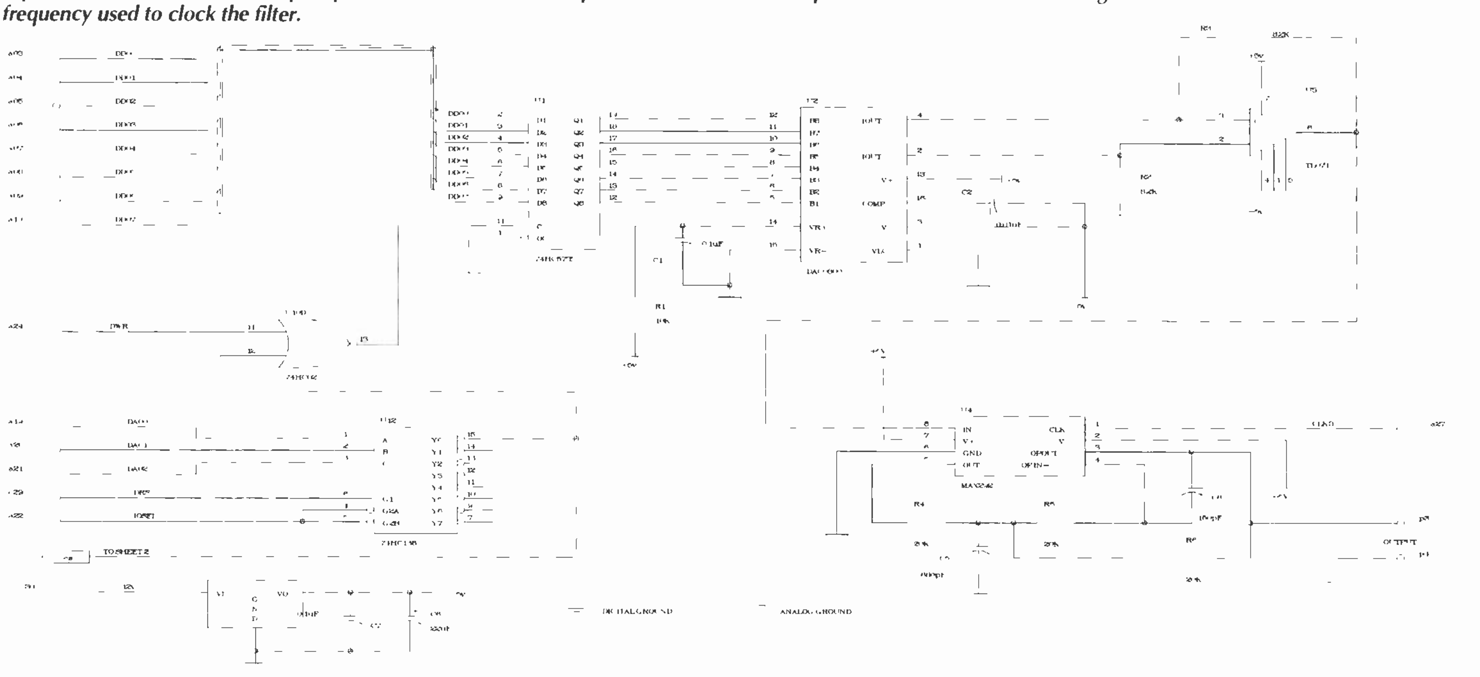
The software system consists of a compiler, assembler, linker and simulator. Broadly speaking, the compiler translates C language subset source into assembly language, these sources are combined with other assembly language sources, assembled and linked into a single executable image which can be either split into high and low byte images with the provided utility U1X 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 feed into a MAX292 8th order low pass switched-capacitor filter ic. The internal op amp of the MAX292 is used to provided a 2nd order low pass filter to remove the clocking frequency used to clock the filter.



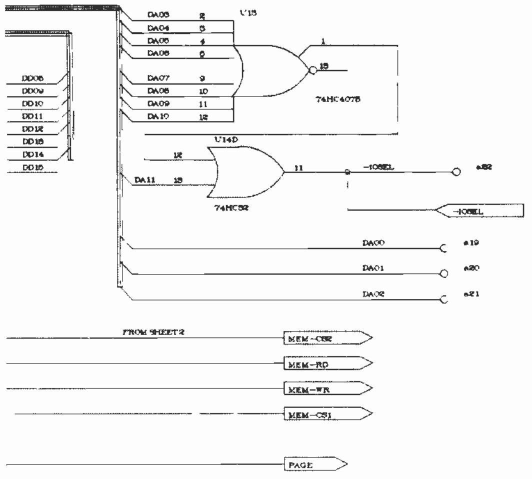


image. In addition, the linker can produce a file containing all the symbols which are externally defined within the link. This file can be used by the simulator to give symbolic debug information.

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 for the target system. Thirdly, the code file can be downloaded into 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.

The simulator has a generalised user interface which means that anywhere numeric input is required, the user can enter complex C language type expressions using any operators 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 *DAC800* 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 320C10'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 DSP developer.

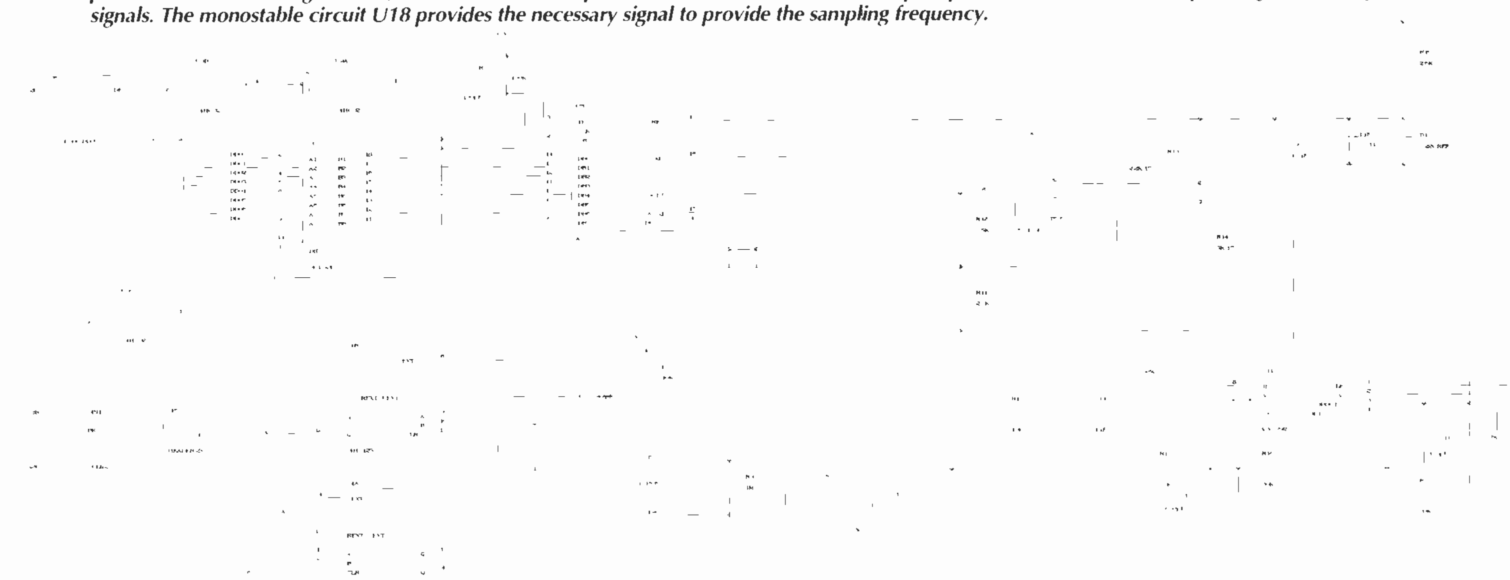
HARDWARE/SOFTWARE SUPPLIER DETAILS

The TMS3201X software tools mentioned in this article can be obtained from: John Mackinnon, Hippo Solutions Ltd, 7 West Preston St, Edinburgh EH8 9PX. Phone 031-668 4701. Total cost is about £150.

The DSP Developer can be obtained from: Tycho Designs 38, Playfields Drive, Branksome, Poole, Dorset BH12 2EQ. Phone 0202-736791. 12-bit and 16-bit ADC/DAC boards are at present being developed and will be available shortly from the same source. The compete hardware cost is also about £100.

Readers requiring further information should apply directly to the addresses above. The magazine has provided these vendor details in good faith but we have not inspected any of the products on offer directly. Readers should assure themselves that any software or hardware relating to this article will work in their intended application before purchase.

A MAX292 8th order low pass switched-capacitor filter ic together with a 2nd order low pass filter using the internal op amp of the MAX292 provides the anti-aliasing of the A/D converter. The output of the filter is feed into the U7 op amp circuit which convert bi- polar signals to uni-polar signals. The monostable circuit U18 provides the necessary signal to provide the sampling frequency.



```
        titl    'ADC input/output program'
        idt     'adc'
        dseg    ; Start a data memory segment
adc     bss     1       ;One word for the adc value
        dend

        aorg    0       ; Link to absolute address zero
        b       main    ; On reset jump to label 'main'
        ret             ; On interrupt do nothing ( just return )


        pseg                    ; Start a relocatable program segment
main    dint                    ; Disable interrupts
        ldpk    0               ; Make the active data memory page, page zero
loop
        bioz    loop1           ; Wait for sample period
        b       loop            ; not ready yet
loop1
        in      adc,0           ; Read ADC and store in adc store
        out     adc,0           ; Write value to DAC
        b       loop
        end
```

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 (eg. ADC) are shifted. At each delay stage a multiplier weights the delayed sample by some factor and the sum of all the scaled past signal samples are accumulated (summed) to produce the filter output. All DSP processors are designed specifically to implement this multiply-accumulate operation as fast and as efficiently as possible since it is a common requirement of a number of signal processing algorithms.

In this filter type, the output signal is only dependent on the past samples of the input signal. If an impulse signal is applied at the filter input (an impulse is a signal comprising a single sample of value 1 and all other samples zero) after an initial transient of N samples the output signal will also become zero, thus its impulse response has a finite duration. By taking the spectrum of this impulse response the frequency characteristic of the filter can be determined.

To design a given filter it is necessary to determine both the appropriate order N for the filter and the values of the multiplier coefficients for each delay stage. For a linear phase filter these must be symmetrical about either side of the $N/2$ multiplier coefficient. A number of filter design techniques and programs are available to tackle this problem. Most of these methods involve a process of iteration from an initial filter specification to an acceptable compromised filter response. In general however the larger the filter order the less the compromise but the greater the computation.

A differentiator

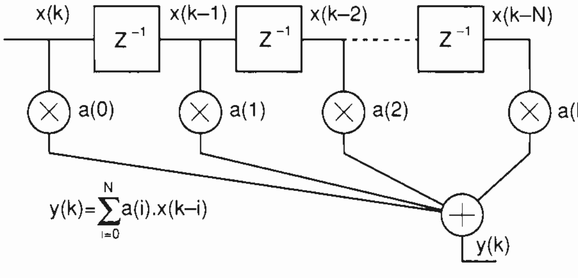
The output from a differentiator is the gradient of the input signal. In the sampled data domain this gradient can be calculated from the input signal samples by assuming that at the given sample rate the gradient can be approximated as the gradient of a straight line connecting successive samples. For this simple differentiator the coefficients would be [1 , -1].f_s. Using 16-bit fractional two's complement integer representation, the coefficients would be [32767, −32767] ie 2^{15}−1. The scaling by f_s 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 −32678 and 32767 in this case. There are a number of approaches which can be taken to this problem:

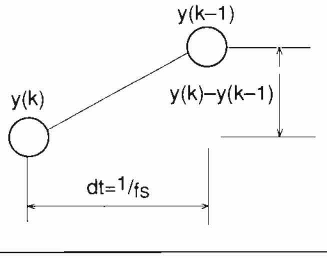
1. Assume that the input sequence which could cause overflow is improbable and make no allowance for its occurrence.

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/sum from i=0 to 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-



Finite Impulse Response (FIR) filter is commonly use to implement digital filter. The filter response is determined by the values of the coefficient a(0) ... a(n-1). High pass, low pass and band pass filters can be all implemented using structure shown. x(k)..x(k-n) are samples of the input signal , which for each output sample y(k) are multiplied by the coefficient a(0)..a(n-1) and summed.



A differentiator finds the gradient of the input signal at each sample point. For example, if the input signal is a triangular wave, the output of the differentiator will be a square wave. A differentiator can be implemented by a FIR filter with coefficients (1,-1). This approximate the gradient at the sample point as the gradient of a line connecting successive points in the input signal.

```
TMS32010 FIR Filter Code

ORDER      equ  2
     tseg
COEFFTABLE
     data -32767,32767
     tend

     dseg
     dorg 0
DELAY      bss  ORDER           ; Delay line starts at location 0
COEFF      bss  ORDER
     dend

     pseg
     sovm                                 ;Saturation Arithmetic
     larp       AR0
     lark       AR1,COEFF+ORDER-1   ; Coefficient Arra
     lark       AR0,DELAY+ORDER-1   ; Data Array
     zac
     lt    *-,AR1     ; Fetch Last Coefficient            mpy   *-,ar0
; Calculate Last Partial Product
; Calculate remaining products in reverse order
loop
     ltd  *-,AR1      ; Accumulate last product & shift delay line
     mpy  *, AR0      ; Calculate next partial product
     banz loop        ; Decrement ar0 and Loop back
     apac             ; Accumulate very last partial product
     pend
```

imum of minimum value depending on the sign of the result.

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 elements, 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.

The ltd instruction is paramount to the efficient calculation of the fir filter on the *C10*. This one instruction causes the pending previous result in the product register in added into the accumulator; the T-register is loaded with 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 *TMS32010*.

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 TMS32010

Traditionally sinewaves are generated by analogue techniques. It is now possible to generates 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 DSP 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 *TMS32010* 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.

■

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

```
Direct Digital Synthesis of Sinewaves program

        TSEG
M1      DATA    #07FF           ; Mask confines addresses to table
STEPL   DATA    #30C2           ; Poke phase acc. fractional 16-bits
STEPH   DATA    #0008           ; Poke phase acc. integer 16-bits
        COPY    'SINE.ASM'      ; 2 k data table
        TEND
        PSEG
SINA    EQU     0
MASK    EQU     1               ;Mask to confine addresses to table
OFSET   EQU     2               ;Start address of sine table
DELTAH  EQU     3               ;Integer part phase step
DELTAL  EQU     4               ;Fractional part phase step
ALPHAH  EQU     5               ;Phase accumulator integer part
ALPHAL  EQU     6               ;Phase accumulator fractional part
;
; Initialise generator
START   LDPK    0
        LACK    M1
        TBLR    MASK            ;Table 2k address mask to data memory
        LACK    SINE
        SACL    OFSET           ;Start of sine table to data memory
        ZAC
        SACH    ALPHAH          ;Initialise starting phase to zero
        SACL    ALPHAL
        LACK    STEPL           ;Phase accumulator fractional 16-bits
        TBLR    DELTAL
        LACK    STEPH           ;Phase accumulator integer 16-bits
        TBLR    DELTAH
; Wait for hardware timer
WAIT    BIOZ    SWAVE1
        B       WAIT;
; Infinite Loop to output sine samples
SWAVE1  ZALS    ALPHAH          ;New phase integer part for masking
        AND     MASK            ;Mask to ensure wraparound in 2k table
        SACL    ALPHAH          ;Save wrapped value
        ADD     OFSET           ;Add offset to start of 2k table
        TBLR    SINA            ;Next sample from table to data memory
        OUT     SINA,PA0        ;Output this sample to the DAC
        ZALH    ALPHAH          ;Load 32 bit phase accumulator
        ADDS    ALPHAL
        ADDH    DELTAH          ;Add next phase step - (don't worry
        ADDS    DELTAL          ;about overrunning end of 2k table)
        SACH    ALPHAH          ;Save new phase value for next sample
        SACL    ALPHAL
        B       WAIT
        PEND
```

PC
ENGINEERING

DADiSP revision with reservations

DADiSP is one of the most flexible and comprehensive signal data processing packages on the market. But Allen Brown wonders if Version 3.0 had been better left in the box.

Signal processing software package *DADiSP* works like a spread sheet, 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.

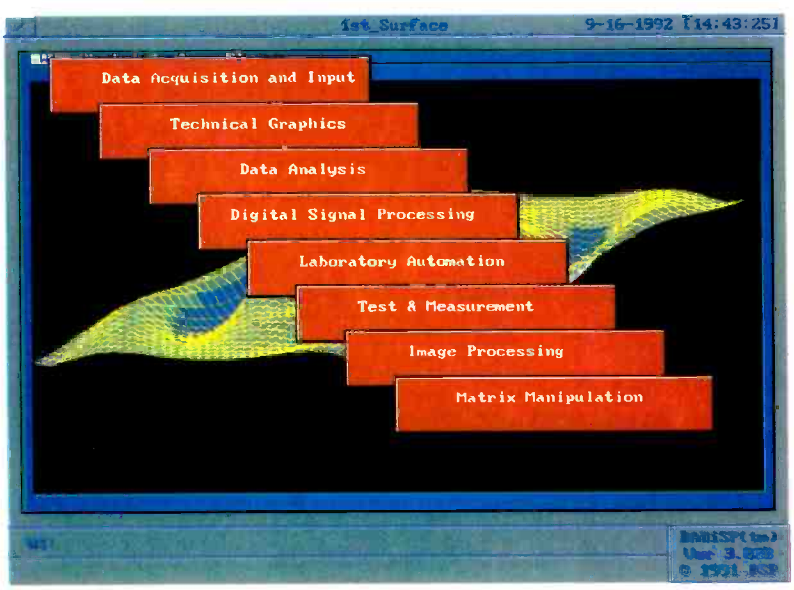
Version 3.0 adds some major extensions to the package. But it is not all good news. The inclusion of a parallel port dongle is a real pain, there because of massive copyright infringements. In one Mediterranean country the software became one of the most popular signal processing packages, yet only a 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 effect *DADiSP* and *Windows 3.1* are incompatible (Now solved – see Update box). 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.

The VGA standard is supported but alas not Super-VGA which is replacing its predecessor. Mouse operation is a welcome addition of Version 3, though its use is very limited. For example, the easy way to alter the colour of a graph display would be for all the colour options to appear under mouse control. But *DADiSP*'s approach is a clumsy alternative, requiring prior knowledge of numbers assigned to colours. Although the mouse does generate a series of drop-down menus, practice is required to appreciate the range of options which can be accessed by it. The user manual is not always helpful in this respect either.

System requirements

286, 386 or 486 PC
Maths coprocessor (optional)
3 Mbyte ram
3 Mbyte hard disc space
Microsoft mouse
VGA or EGA graphics



DADiSP's spreadsheet format uses graphs as elements, as opposed to rows and columns of numbers.

More facilities

The range of functions has been greatly expanded over earlier versions[1] 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 spread sheet 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. A demo program supplied with the package is in effect a macro and illustrates many of the options available in the product.

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 elements 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 a *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

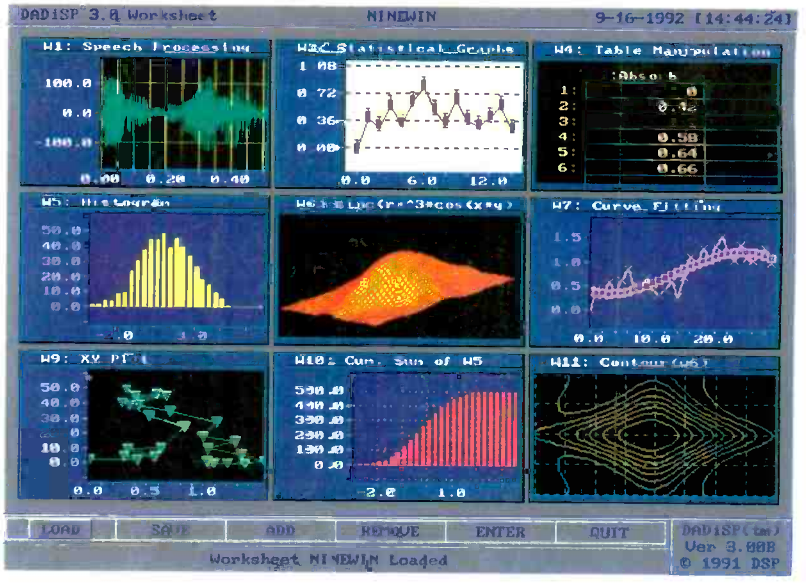
Documentation can be described in a single word – poor. The reference manual is written like a dictionary, and though it might be useful to old hands, it will leave the first time user none the wiser. The user manual is slightly more illuminating – but not much. It serves as a useful introduction to 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.

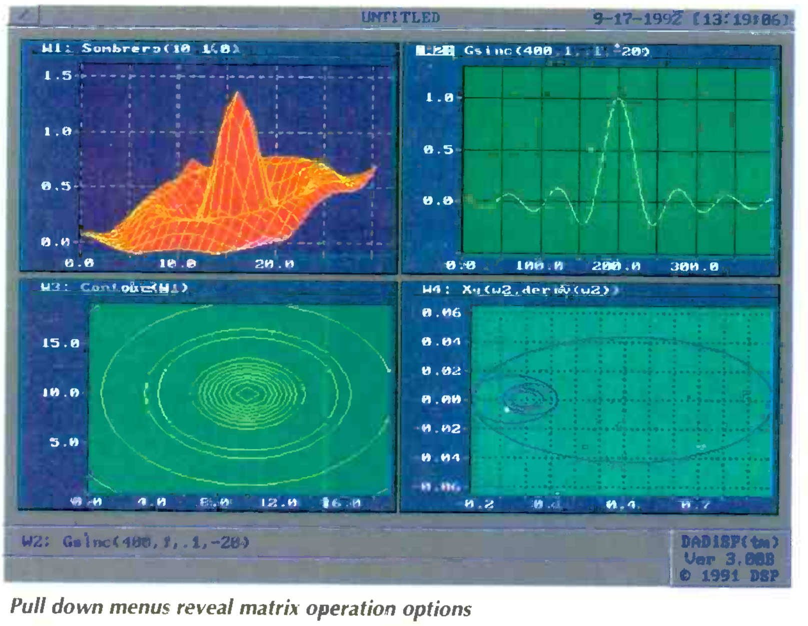
That said, *DADiSP* still remains one of the most flexible and comprehensive signal data processing packages on the market. ■

References

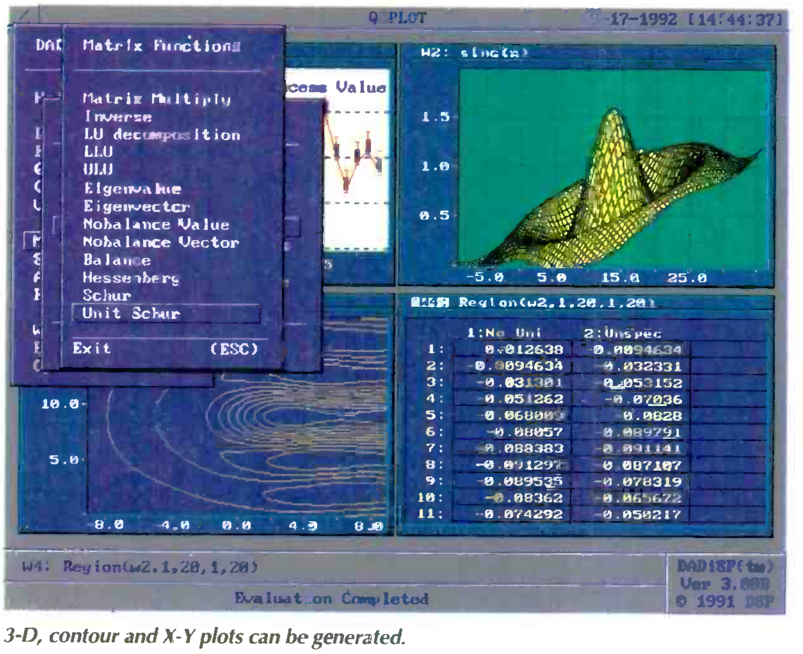
1. DSP Design with DADiSP, *EW + WW*, Dec 1989, p1151



Application areas of DADiSP span the whole of engineering.



Pull down menus reveal matrix operation options



3-D, contour and X-Y plots can be generated.

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.

PC

ENGINEERING

Graphical demonstration of scientific power

If you want to convert your PC into a scientific calculator with powerful graphing at minimum cost, Don Bradbury shows why CC4 is the program to consider.

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.

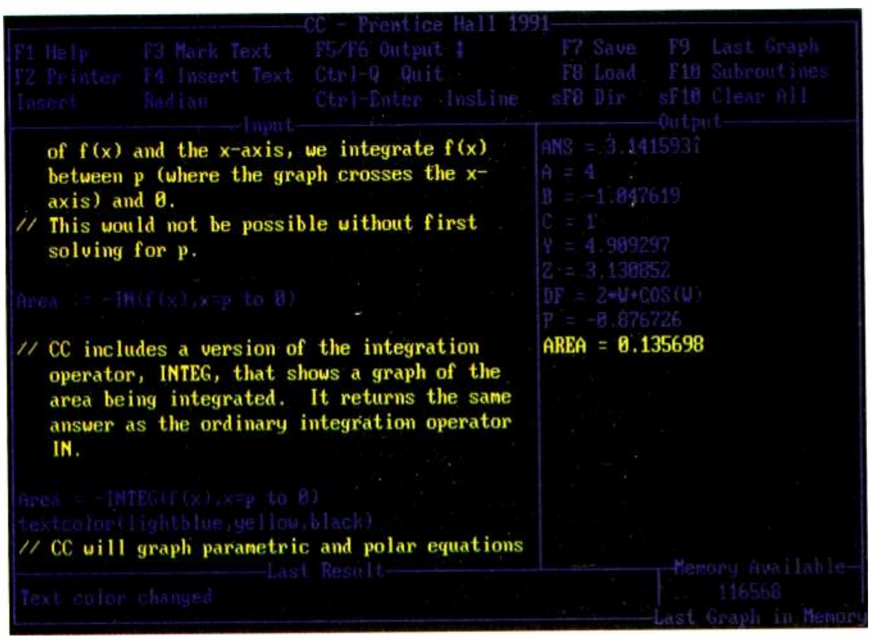
Files can be output to disk or graphic screens to a printer, and all functions are entered as they are written, with other custom procedures created in a *Pascal*-like environment. The considerable power derives mainly from the enormous range of commands that can be used to program not only the calculator 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.

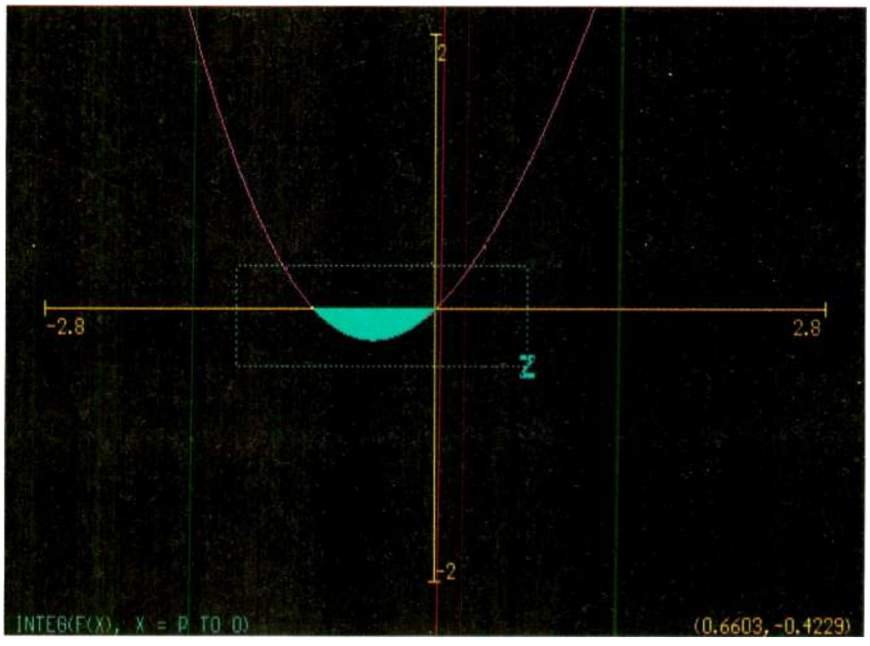
The output window is used to report the computed result of expressions submitted to the calculator. If syntax of an entered expression is incorrect, *CC4* puts up an explanation box, removable with the esc key, calling for another attempt. The program's default insert mode can be switched to overwrite, 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.



CC4 provides all the function of a top market scientific calculator. Both data entry and results are shown on a single screen. Output data may be presented as complex graphics as depicted in other screen shots in this article.

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



2D graphic showing an area being defined for zooming.

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, LNGAMMA, BINOM all have the usual meanings. π 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 fraction, 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.

CRTL/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 Fransisco State University. Its predecessor, *CC3*, is a standard commercial product, 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, trigonomic 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 plot presentation and enhancement.

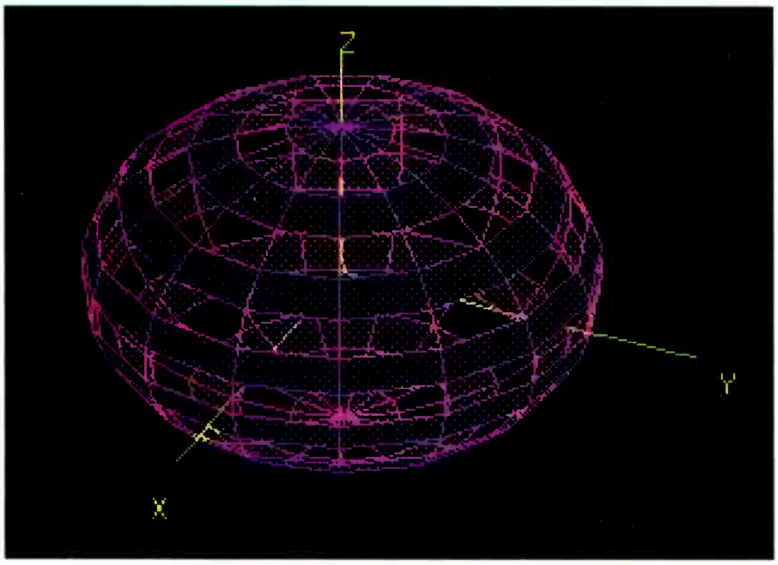
On-line help

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

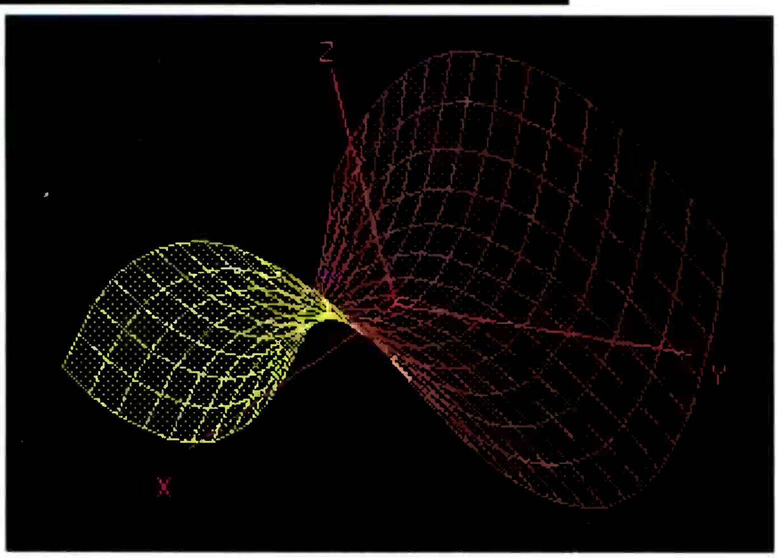
Under broad headings such as "Capabilities of CC", Help for individual commands", "Function keys and other special keys", "Printing", and "Disk and memory management", the help file fully describes the program. Its 84k is well-used disk space, but it can be omitted to operate from the much slower and 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



Mid resolution of a 3D image.



A 3D image in high resolution mode.

Interactive data analyses can be created by subroutines to develop dynamic presentations or interactive mathematical lessons. Where a user needs, for example, to set crosshairs in the graphical mode or to await some input routines can be paused.

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 fineness 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.

CROSSX and CROSSY return the last X and Y coordinates of the crosshairs, so if 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=CROSSX) 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. ■

Total +VAT

CIRCLE NO. 110 ON REPLY CARD

CIRCUIT IDEAS

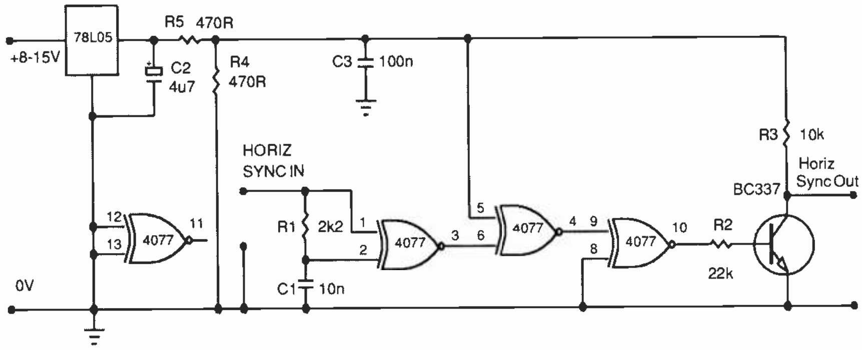
Automatic generators for VGA sync

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_4, R_5 potential divider to 2.5V eliminates the need to amplify the input signal. A CR network R_1C_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



Automatic sync plunger produces negative sync, regardless of the input polarity.

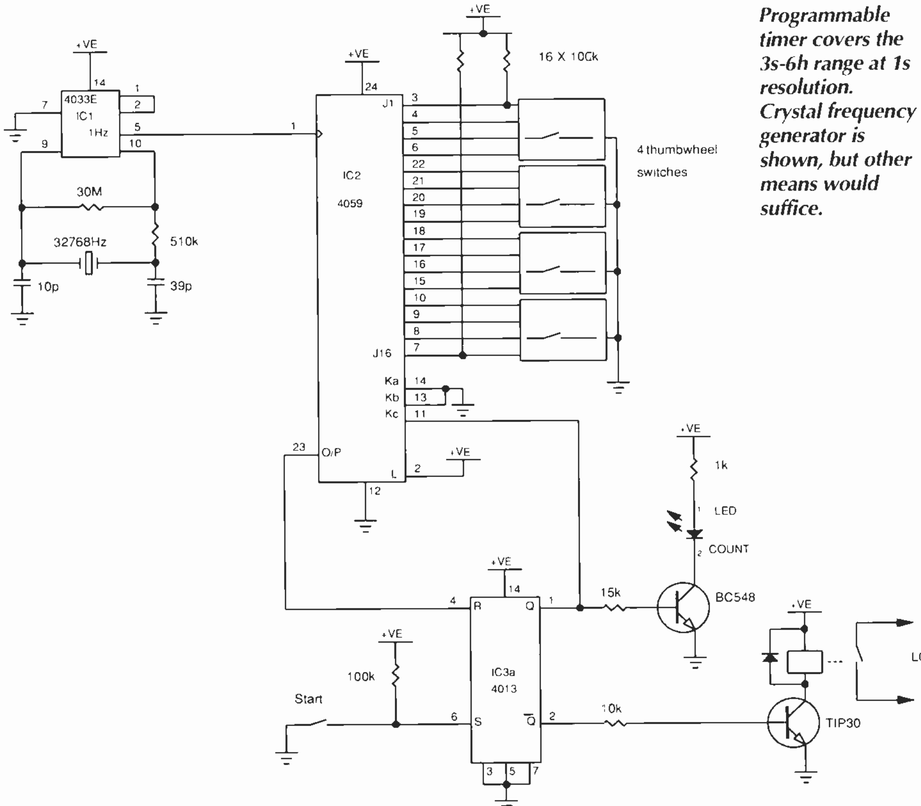
Programmable timer

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 Ka,b,c being low. A start signal from the switch sets the D-type flip-flop IC_3 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



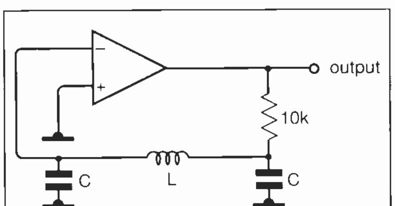
Programmable timer covers the 3s-6h range at 1s resolution. Crystal frequency generator is shown, but other means would suffice.

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.

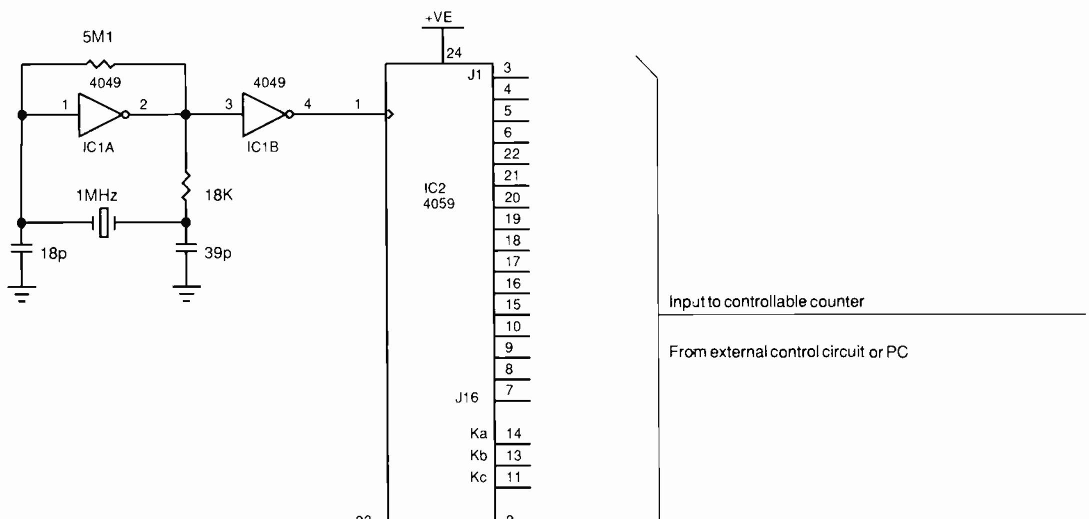
Any of the standard op-amps will work reasonably well at audio frequencies, trace (a) being the output of a *741* at 6kHz; trace (b) is an *LM6361* at the same frequency. The third trace is an *LM6361* at 1.3MHz. Output frequency is $1/(2\pi\sqrt{(LC/2)})$, the two capacitors being effectively in series.

Michael A Covington
Athens
Georgia
USA



opamp	L	C	frequency
Any	33 mH	0.04 μF	6 kHz
LM6361	470 μH	40 pF	1.3 MHz

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.



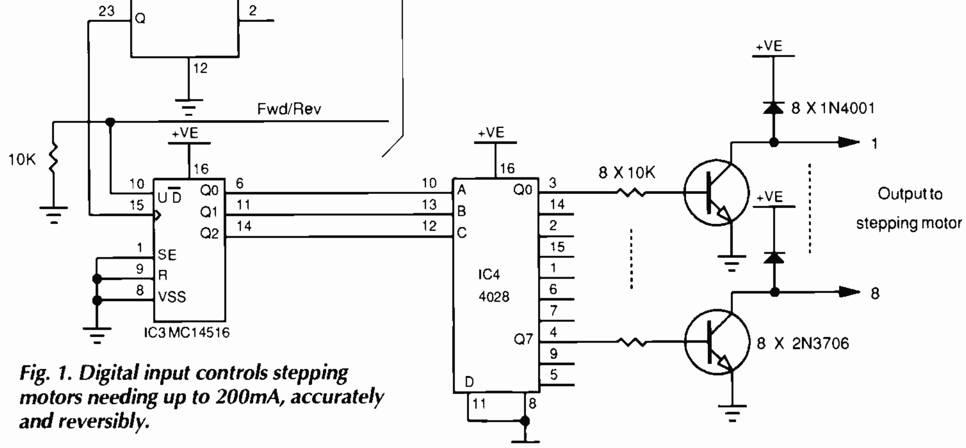


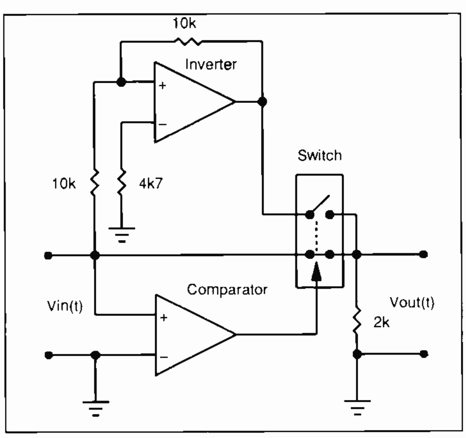
Fig. 1. Digital input controls stepping motors needing up to 200mA, accurately and reversibly.

Fast, small-signal rectifier

For the rectification of signals of less than about 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 *IH5141* 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



Fast rectifier for small signals avoids distortion and lack of bandwidth in usual op-amp with two feedback diodes.

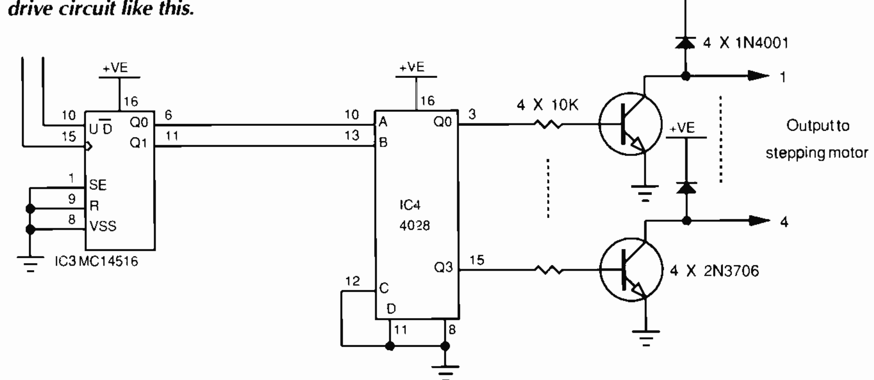
Adjustable, accurate stepper controller

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 IC_2 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. 2. If the motor is an 4-phase type, modify the drive circuit like this.



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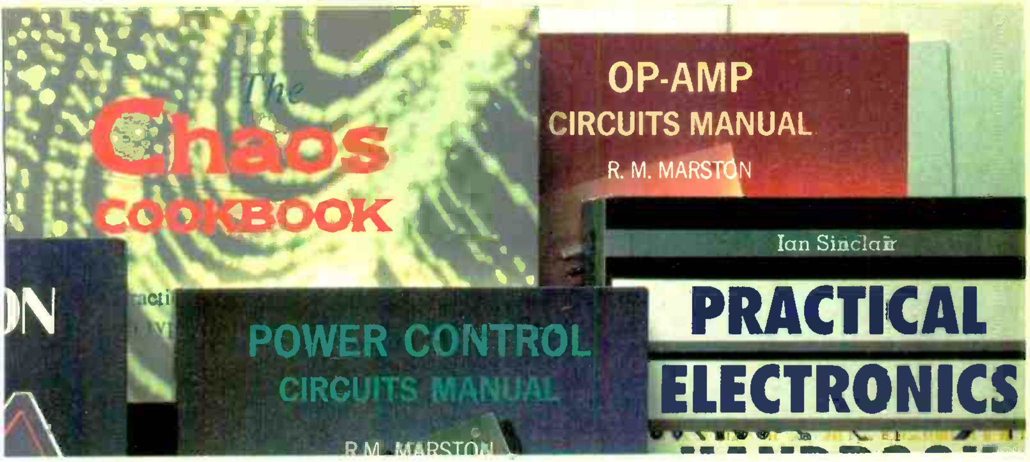

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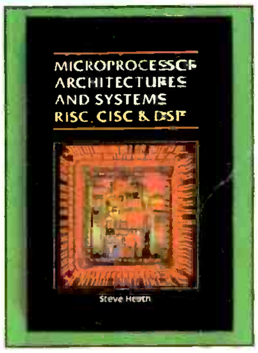

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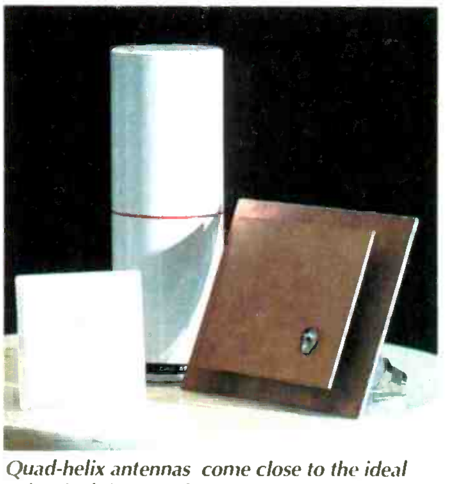
GPS

4: radio architecture

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 astronomic. 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 not likely 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.

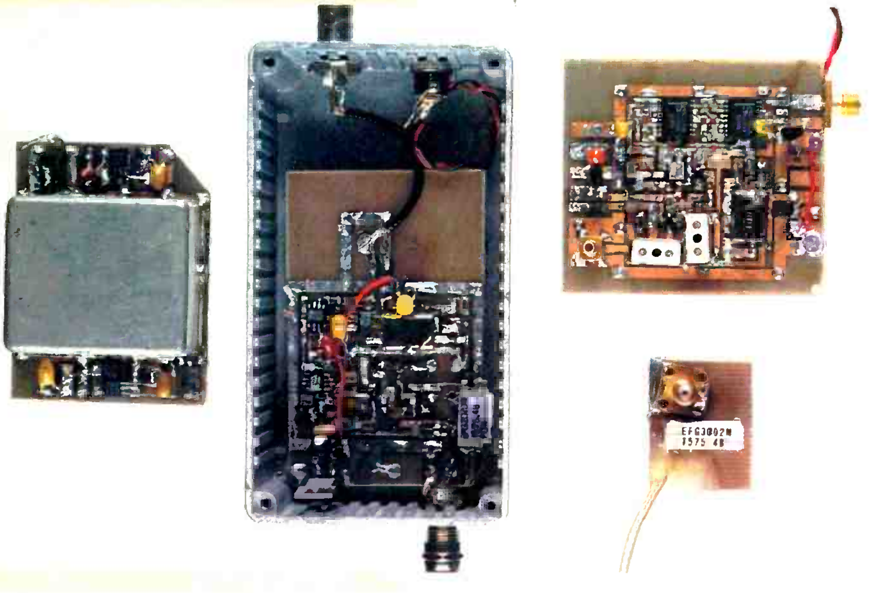


Quad-helix antennas come close to the ideal gain of 3dBiC over the entire hemisphere. The patch is cheaper and its +7dBiC overhead makes it useful on land where low angle satellites can be blocked by trees.

The Avantek *ROC* – receiver on a chip – is not yet available, just vague details in sales presentations, but I have tested the Plessey *GP1010* single chip GPS radio so this will be discussed here along side 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 i/o solution can be replaced with two pals if required, much cheaper but using more power. In the computer side of the hardware, ultimate simplicity is achieved by using a *T805* transputer.

It needs just two address latches to interface with eprom and sram, 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-biassed 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

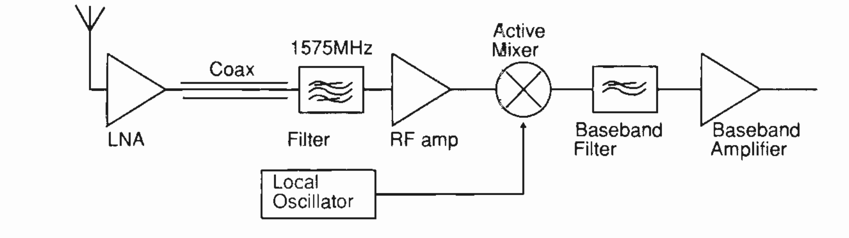


***Fig. 1.** New radio architecture: direct down conversion allows a very simple radio architecture, and makes the local oscillator at signal frequency where low-cost filters are available.*

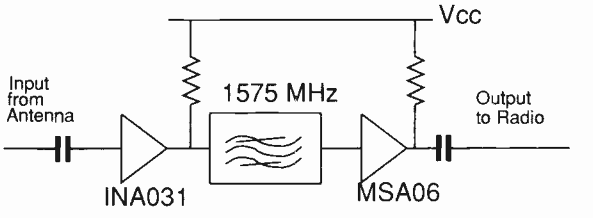


***Fig. 2.** A simple, robust LNA can precede the filter, and shows a better noise figure than a complex GaAsfet circuit placed after the filter.*

33dB 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 30dB 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 30dB 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 so 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 capacitance and has a self-resonant frequency above which it actually looks capacitive. This frequency is below 1.5 GHz for all but the lowest value surface mount inductors. A similar problem arises with coupling capacitors which behave inductively. Thus low value capacitors are chosen, and generally avoided where possible. Ceramic filters block DC, so help considerably.

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 INA031. 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 intermod 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 $f_{intermod}=mf_1 \pm nf_2$ where f_1 and f_2

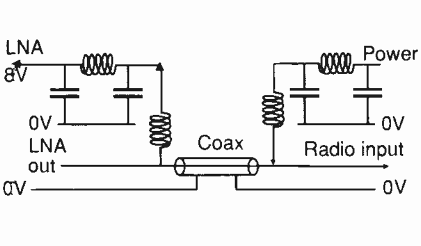


***Fig. 3.** Feeding the LNA up the coax is common practice, but not easy at these frequencies. A cascaded pair of inductors does the job, but to be safe against short circuit, should really be fed from a current limiting transistor. A resistor is not adequate, due to variation in current between different LNAs and lack of voltage headroom.*

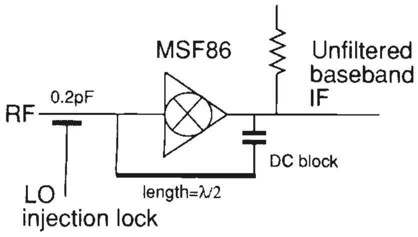


Fig. 4. The self oscillating two-port mixer is extremely simple, but the feedback path length is critical, and not easily managed in production.

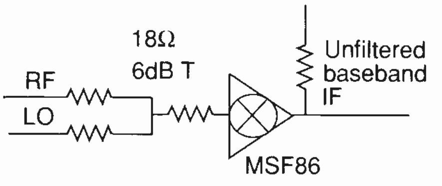


Fig. 5. External LO 2-port mixer: The 2-port mixer works well with an external oscillator, but this simple circuit is not suitable for direct downconversion due to poor input matching at baseband.

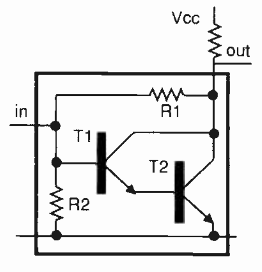


Fig. 6. Mixer internal circuit: Note that the base bias is derived from the collector, without decoupling, so the bias resistors determine both bias, and via feedback, the gain and input impedance.

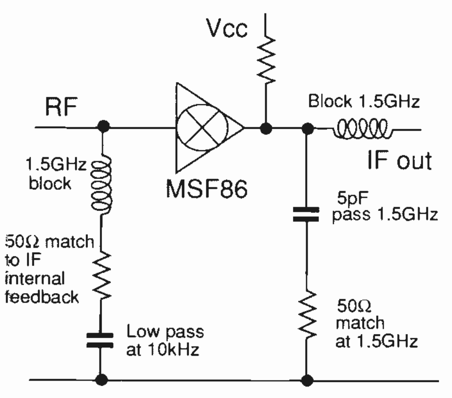


Fig. 7. Mixer matching circuits. To match the mixer to both 1.5GHz and 10kHz at both input and output, considerable complexity is required... a baseband match blocked to RF at the input, and an RF match blocked to baseband at the output.

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 reject 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 *MSA06* stages, preceded by another ceramic filter, both to give further stop band rejection, and to reject any interference picked up on the coax. Two *MSA* stages are used rather than one *INA031*, 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 *INA031* 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 *INA031* 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 *MSF86* 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 are on the same pin.

Also, on the thick low-cost board used, the length of the feedback path making up the 180 degree phase shift required adjusting on each radio, depending on tolerances of aligning the other surface mount components.

As one of the design requirements was an absence of trimming, this was not acceptable. The second approach was to drive the mixer from an external LO, ie no longer self oscillating. (**Fig. 5**).

This worked fine, and a very good radio was demonstrated although the gain was below plan. Then four copies were built, and not one of them behaved satisfactorily.

Two months later, after a lot of blind alleys, always blaming the mixer, the problem was found, not quite in the mixer, but in its input coupling capacitor and in the gain stage following the mixer. The circuit of the mixer is shown in **Fig. 6**.

As can be seen, the collector voltage, both bias and signal, is divided by R_1 and R_2 to give the voltage on the base of T_1. The division ratio is $R_2/(R_1+R_2)$ for the bias, with equilibrium being maintained with about 1.4V on the base and 3.5V at the collectors. The division ratio for the signal however, is different... as the 50Ω source impedance of the previous stage appears across R_2, as the input coupling capacitor blocks DC, but is transparent to the signal. This ratio determines the gain of the stage, and also the sets the input impedance to 50Ω. The problem is, when the mixer has done its job, and mixed two L-band signals to create one at a few kilohertz, that audio signal must also be treated as signal by the input capacitor, or the negative feedback will try to cancel the mixing product. Normally, this is not a problem, as a capacitor can be found that will behave adequately for both signals.

However a capacitor suitable for a 10kHz signal in a 50Ω system is 0.5μF or so, not compatible with a 1.5GHz signal even if a 50pf capacitor, or other "reasonable" value, were placed in parallel. The problem comes from the extreme diversity of the RF and IF frequencies, particularly the true baseband nature of the IF. There is a solution, and it is shown in **Fig. 7**.

A separate path is provided for baseband signals, into a 47Ω resistor, DC blocked by a large capacitor to avoid upsetting the bias, and blocked from the RF signal by a small value, high frequency inductor. With any IF, there is a need to terminate the mixer output correctly. To feed a low frequency amplifier for the baseband chain, a 5pF capacitor and 47Ω resistor shunt the 1.5GHz RF and LO signals, and also the 3GHz sum product. This meant that there were some five components needed above those in the signal path, and still more in 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 MSF86 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 inductor, 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 a cut off point of a few MHz. As there is a considerable phase change at the cut-off frequency, this is set much higher than required, allowing the unavoidable low-pass characteristics of the op-amps to complete the job. With the three port mixer, the 1.5GHz level is about –35dBm, so no longer presents a major problem.

The mixer also includes an output buffer, so is tolerant to mismatch on the output. Thus the baseband filter becomes a simple L-C pair. The other problem in the four production radios related to noise generated in the first op-amp.

After measurements indicated this was not a problem, four op-amp stages were used. However, as is obvious with hindsight, the measurement was erroneous because it was done with the input open circuit. Connecting the input to ground created more noise, which was attributed to poor decoupling of the mid-rail analogue ground.

In truth, the noise was the first op-amp, and with the op-amp input open-circuit, the closed loop gain became the open loop gain allowing the amplifier to cancel all its own noise.

Subsequently all the op-amps were replaced by three transistor stages, as this is both cheaper and uses less board area. The op-amps or the transistor design need to give level gain from almost DC to at least 500KHz, after which a gentle rolloff is acceptable. However, the phase response is important; it must not disperse the signal, so a linear amplitude response out to 1MHz results.

Signetics 5534 (single) and 5532 (dual) op-amps were used, which have a gain bandwidth product of 10MHz. This allows a gain of 10 out to 1 MHz as required, and no low-pass components have to be added. Thus three stages give x1000 voltage gain, or 60dB. (**Fig. 10**).

The same can be achieved with three transistor stages, in common emitter, with an emitter resistor to control the gain. These however yield a frequency response out to tens of MHz, so a large slugging capacitor is added across the second stage.

The 5532/4 op-amps clip cleanly when overdriven, so there is some 60dB of overdrive capability in that design. The transistors are similar, due to the transition from high to low input impedance as the stage saturates, which reduces the gain of the previous stage.

However the op-amps cannot drive a 50Ω coax from rail to rail, and the transistors limit asymmetrically. Thus despite the fact that the only concern is the timing of the zero crossings, we must handle the overdriven state correctly. Both asymmetry and slew rate limiting distort the timing of the zero crossing, the former by creating a DC bias that moves the effective switching point, the latter by delaying steep zero crossings, but not shallow ones. Thus both designs use an emitter follower driver stage before the coax, and diode clippers to limit the output swing symmetrically to 1.2V p-p. Driving into 50Ω would require an unmanageable bias current in the emitter follower without the clipper.

Oscillator chain

The oscillator chain has also passed through many iterations before arriving at a design that performed well without sacrificing simplicity. The initial concept was to use a self oscillating two port mixer injection locked to a subharmonic. The second version was to use an injection-locked oscillator separately, at 525MHz, extracting the third harmonic passively for use as the LO signal. The 525MHz oscillator was locked to a 105MHz 5th overtone crystal oscillator. This version worked perfectly, and used only two stages in total for 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 removed 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 the105MHz 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.

Thus the oscillator frequency is always about 5kHz 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 must be coherent, a different approach is used. Rather than 105MHz (GPS $L1/15$), either $L1/14$=112MHz, or $L1/22$=71.6MHz are used. The oscillator is then locked to the sampling oscillator (10.23MHz), where it is the eleventh or seventh harmonic respectively.

The former case is better as it generates more L-band output amplitude, but the latter has the advantage that by using off-the-shelf 70MHz filters as used in microwave receiver IF strips, it can be ensured that the 10.23 MHz sidebands do not pass through to the LO. If they did, the L-band filter would not be able to reject them, while if the nearest undesired harmonics are 70 or 112 MHz away, the 1575 resonator rejects them.

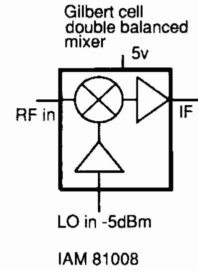


Fig. 8: The 3-port active mixer has only RF at the RF port, only LO at the LO port, and only IF at the IF port. It also tolerates output mismatch.

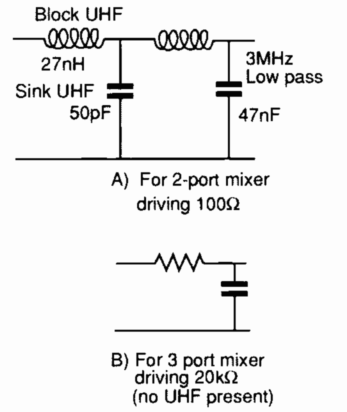


Fig. 9. Baseband filter. The baseband filter must exclude 1.5GHz type signals, but also 3MHz and up. Two inductors, the first to block high frequencies from the second, serve the purpose.

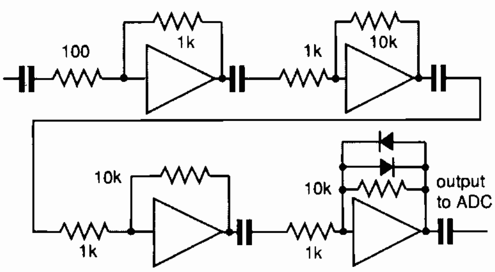


Fig. 10. Op-amp constructed baseband section. Four op-amp stages give 80dB of gain from 0 to 1 MHz, with clean roll-off above that and good phase response out to the required 500kHz of the folded GPS signal. However the noise of the first op-amp was excessive, and with the later mixer design, only 60dB was required.

Performance

The resulting radio performance can be measured in two ways. The first is a comparative measure of recovered satellite power using the same software on many different radio sys-

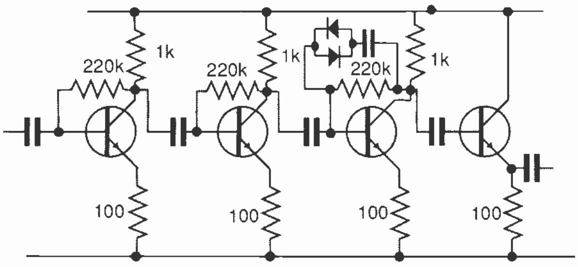


Fig. 11. Discrete version baseband section. Four transistor stages – three for voltage gain, the last for driving the 50Ω coax when the radio is remote from the CPU, replaced the op-amp design.

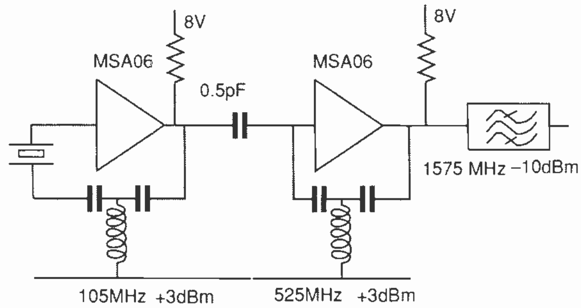


Fig. 12. Full 105/525/1575 Oscillator chain. The first successful oscillator chain used a 105 MHz crystal, driving a x5 multiplier stage at 525MHz, which drove a ceramic filter which picked off the 1575MHz signal passively.

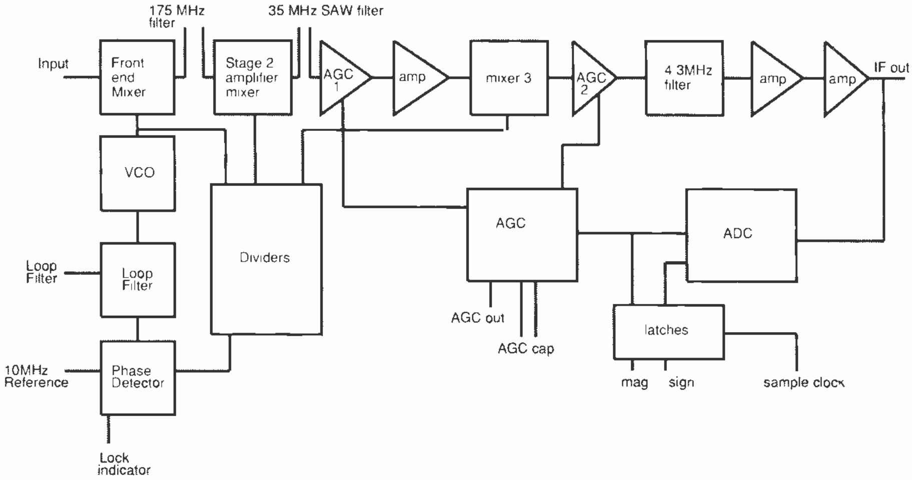
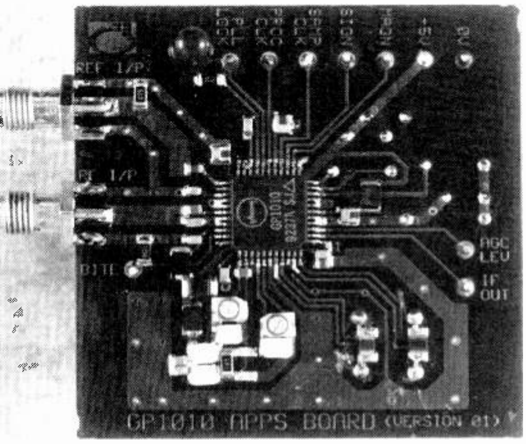


Fig. 13. GEC-Plessey GP1010 single chip GPS radio. The GP1010 has a triple down-conversion, complexity that costs little because it is on-chip. It includes the synthesiser, including VCO and dividers, and even the A-D converter and sampling latches, in this case two-bits wide.

tems, in a benign environment. The second is to consider the effects of hostile environments.

On a recovered power basis, the radio is about 1dB better than any other I have tested. Unfortunately I do not currently have the software to convert this into a carrier-to-noise measurement, so there is no absolute criteria. This simply means that the radio works... It is



GEC-Plessey's GP1010 sample board provides everything except the LNA.

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 combination of tuned patch antenna and ceramic resonator filters means that there are few problems at RF. The interference power equation runs like this. The radio has some 50dB of RF gain, with a maximum permissable 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 continuously would certainly destroy the LNA.

If this radio has a weak point, it is breakthrough 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, provided 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 applications.

However with the target price for car use being so low, a combination of power line filters 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 4.3MHZ. 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 105MHz. 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 1505 MHz, 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.

The final filter at 4.3 MHz cannot shape the band effectively, as a 2MHz bandwidth here is such a high percentage. It is thus allowed to be wider, as its purpose is simply to remove noise outside the band that would otherwise alias about the sampling clock onto the signal.

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 software processing. A circuit diagram of the external circuitry, excluding power rail decoupling is shown in **Fig. 14**.

Power rail decoupling is 3.3μH in series and 1nF + 47pF to ground for each of the four power pins. While the IF output is available, this is intended for test use only. A plot from a spectrum analyser is shown in Fig. 15. The large signal shows the effect of the AGC, as the system shuts down to maintain the correct ratio on the magnitude output bit.

The *GP1010* has a 9dB noise spec. This can be deduced from the small signal curve as the analyser is on 300kHz resolution bandwidth, when thermal noise would be −109dBm. (KTB after Boltzman). The noise trace is 15dB below the −100dBm signal, implying −115dBm for the noise, leaving 6dB for the first stage amplifier noise figure. Thus this sample is well within spec. The chip also includes a 2-bit A-D converter with sampling latches. The system designer must apply an appropriate sampling clock externally.

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 300kHz, ideal for the processing developed for my own radio described earlier. 2.0MHz is easily derived from the 10MHz reference. If synchronous operation is required, 2.046MHz would be generated separately, yielding 254kHz as the perceived carrier, but losing the benefits of coherence. Synchronous and coherent operation can be achieved simultaneously from the same reference using a circuit we developed for a GPS-specific transputer prototyped during 1991.

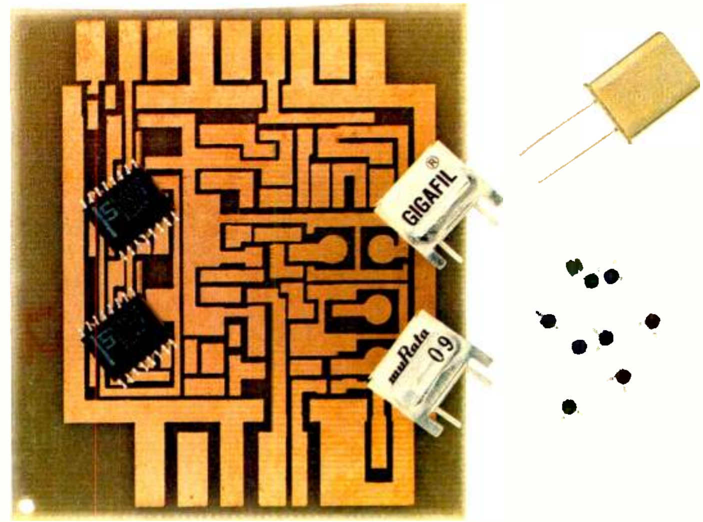
It used a 20MHz reference, the transputer clock, and divides this into two non-overlapping clocks A and B at 10MHz. In general, stream A is used, but if the phase of the tracker has to be delayed, one pulse is deleted from the A stream.

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 20 MHz 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 *16V8* 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 L2 frequency, at 1226 MHz. However it also means that



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.

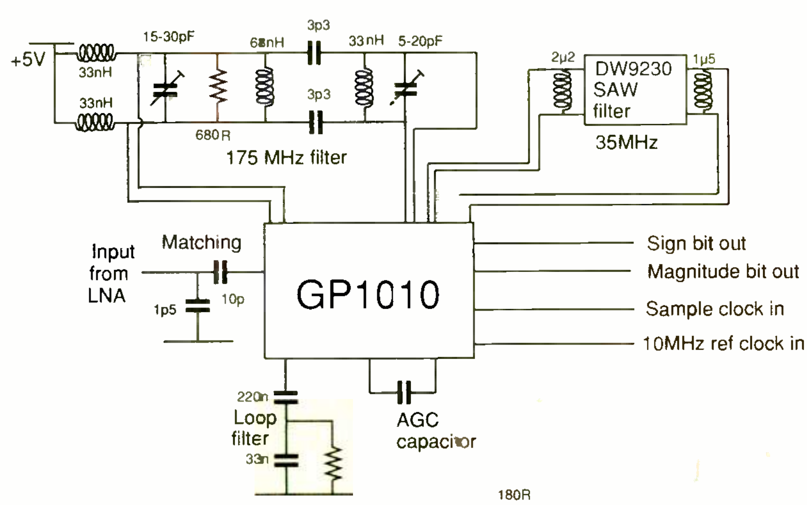


Fig. 14. Radio based on GP1010. All the radio designer has to add, besides clean power rails to each stage, is a reference oscillator and two filters, the first at 175MHz, the second at 35MHz. The VCO loop time-constant components are also off-chip.

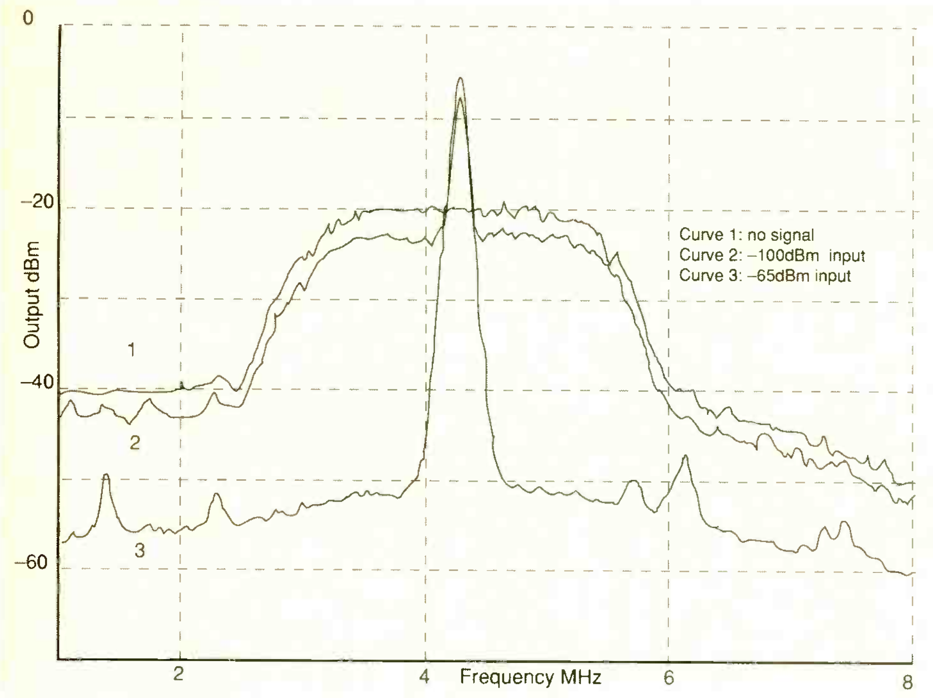


Fig. 15. Test signal through GP1010 based radio. The three curves show the noise response bandwidth, the small signal response (−100dBm), and the large signal response (−65dBm). The small signal is equivalent to GPS via an LNA, but can be seen as it is CW. GPS would be spread across the band almost indistinguishably from the first curve.

there can be no transmissions on the image frequency, as it is reserved for GPS down links and these are below the noise.

The 1400 MHz is divided by ten to give 140MHz as the second LO signal, and by 45 to generate 31.1MHz as the third LO. It is also divided by 140 to lock 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, while the software can absorb carrier shift easily, it is less tolerant of sampler rate change. This is the main reason for the change from 2.046MHz synchronous operation to 2.000 MHz asynchronous, as to maintain the 2.046 at the required 10ppm tolerance requires temperature compensation, while the asynchronous operation can tolerate 250ppm tolerance without problem, up or down. The 250ppm limit relates to requiring an integer number of bytes of samples in each millisecond, as i/o 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 Mattos is a consultant engineer working for Inmos.

Next month: DSP software to extract the GPS signal out of the noise.

IDENTIFYING and avoiding transistor PROBLEMS

Transistors – both bipolars and mosfets – are immune to many problems, but they can still give trouble. Robert Pease shows how robust design methods and proper assumptions about performance characteristics are the keys to avoiding problems. Serialised from his book Troubleshooting Analog Circuits.

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 to blow up, and circuits must be analysed if they do.

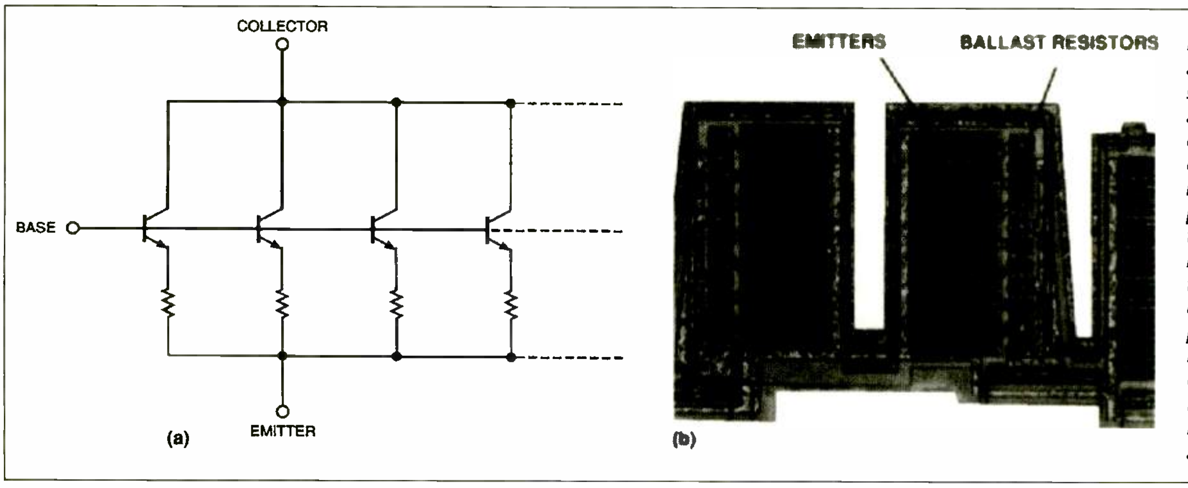
A simple problem is to instal the transistor correctly: three terminals mean the possibility of a wrong connection is considerably greater than with a mere diode. Small-signal transistors are often installed so close to a printed-circuit board that it is not clear whether or not if the leads are crossed 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-217[1] 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 100pF 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. 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.

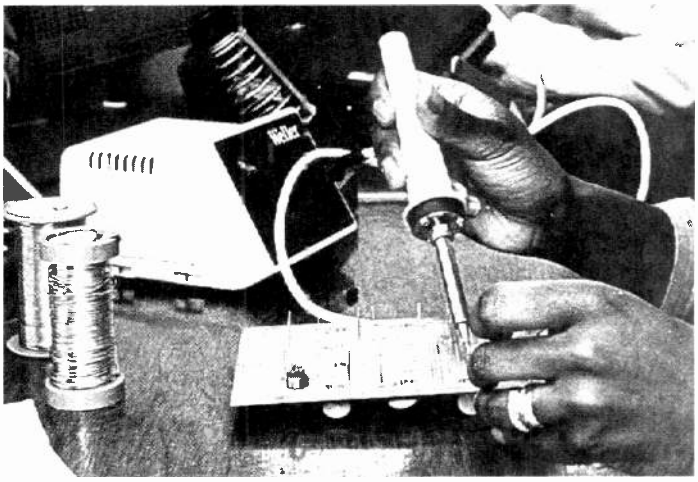
Similarly, transistors from different manufacturers will have different characteristics when going into and coming out of saturation, especially when being driven at high speeds. A components engineer is a very valuable person to have around and can save a lot of grief by preventing unqualified components 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. But unlike a diode, the transistor is normally connected to its leads with relatively small lead-bond wires. So if there is a lot of energy in the power supply, the short circuit will cause large currents to flow, vaporising the lead bonds. As the lead bonds fail, the transistor will ultimately fail as an open circuit.

Beta better?

The h-parameter, h_{rb}, is equal to $\Delta V_{BE} / \Delta V_{CB}$ with the base grounded. Many engineers have learned that as beta rises, so does h_{rb}. As beta rises and h_{rb} 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 around double its low-voltage value, assuming a constant base drive. V_{Early} is approximately equal to 26mV x $(1/h_{rb})$). So, in many circuits there is a point where higher beta simply makes the gain lower, not higher.

Mosfet ICs are static sensitive even though the inputs are diode protected. Cmos logic subjected to static damage may not fail, but develop a fault at some time in the future producing obscure effects.



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 Darlingtons or cascaded followers, and for many years, it's been more important (in most circuits) to have matched betas than to have sky-high betas. You can match betas yourself, or can buy monolithic dual matched transistors like the *LM394*. Or you can buy four or five matched transistors on one monolithic substrate, such as an *LM3045* or *LM3086* monolithic transistor array.

One of the attractions of bipolar transistors is that their transconductance, g_m is quite predictable. At room temperature, $g_m = 38.6 \times I_c$ – much more consistent than the forward conductance of diodes. Since voltage gain is defined as $A_V = g_m \times Z_L$ 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 $1/(R_e + g_m{-}1)$. 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.

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_{GS} of fets 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 bifet process made it feasible to make jfets 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_{OS} 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.

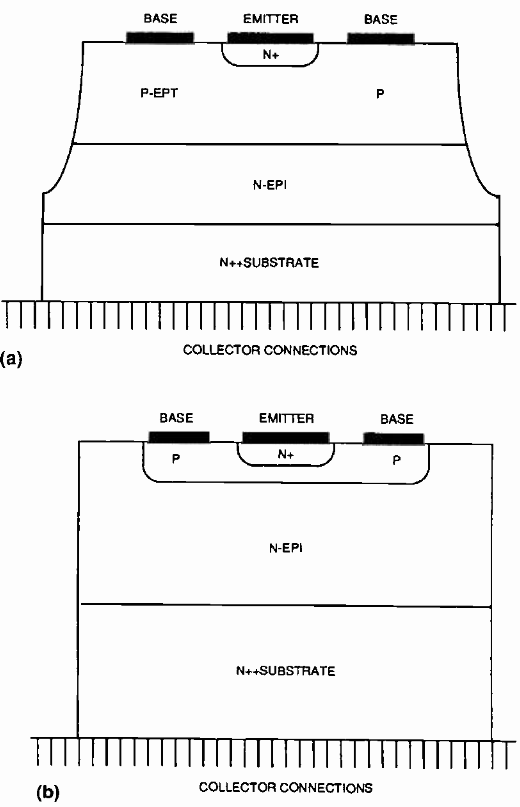
Jfets can have a larger gate current when current flows through the source than when no current flows – called I_{gss}. This looks to be caused by impact ionisation, or "hot carriers." Either way, the gate current has a tendency to increase as a linear function of source current, with an exponential 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 through a large and unspecified impedance.

The substrate of a discrete bipolar transistor's die 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 *Vout*. The *LM196* voltage regulator's substrate is tied to the positive supply voltage, $+V_S$, as are the substrates of the *MM74HC00* family of chips, the NSC *LMC660* and *LPC660* family, and most of the dielectrically isolated op amps from Harris. So, be aware of an IC's substrate connection. If an *LM101AH* op amp's metal can should happen to bump against ground or $+V_S$, there will be a problem. Similarly, the case of an *HA2525* should not be allowed to bump against ground or $-V_S$.

Advantages of mosfets

Mosfets, widely used in digital ICs, are also very useful in analogue circuits such as analogue switches. Quad switches – eg *CD4016* and *CD4066* – are popular because of their



The characteristics of power transistors depend on their fabrication structure. The epitaxial-base structure (a) takes advantage of the properties of several different epitaxial layers to achieve good beta, good speed, low saturation, small size and low cost. This structure involves mesa etching which accounts for the slopes at the die edges. Planar power transistors (b) can achieve very small geometries, small base-widths, and high-frequency responses, but they are less rugged than epitaxial-base types, in terms of forward-biased SOA.

low (typical) leakages and low price. Op amps with mosfet inputs are starting to do well in the general-purpose op-amp market. They used to have a bad reputation for excessive noise, but new IC devices, such as the *LMC662* dual op amp, demonstrate that clean processing can cure the problem, making mosfets competitive with Bifets. Their advantage is a 1000:1 improvement in input current, decreased from 10pA down to 10fA – just be careful not to let ESD near the inputs.

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 *3N160*, 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[2]. So far, I haven't had any blown inputs or bad leakages – at least nothing as bad as 10fA.

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.

Be wary of any devices that manufacturers claim are safe from ESD.

In some cases cmos ICs abused with ESD may not fail instantly, but may become unreliable and fail at a later time. So beware of latent unreliability problems. If you must troubleshoot 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 $I \times R$ 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 $I \times R$ 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.

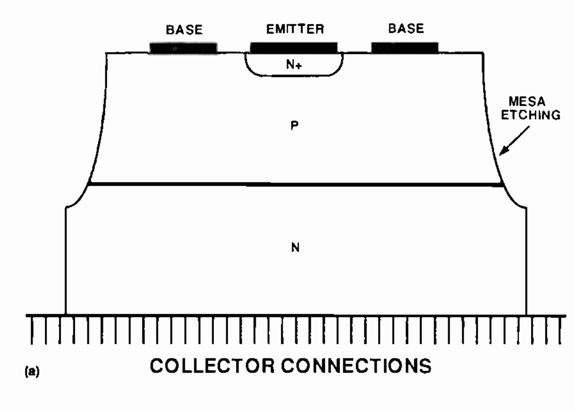
Unless current is turned off promptly, it will continue to increase unchecked, "current hogging" that will cause local overheating, and may cause the area to melt or crater – secondary breakdown, exceeding the secondary breakdown of the device. Designers of linear ICs use ballasting, cellular layouts, and thermal-limiting techniques, all of which can prevent harm in these cases[3]. 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 if 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 overstress and warp the tab 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.



In the old single-diffused structure, n-type dopants were diffused simultaneously into the front and back of a thin p-type wafer. This structure produced rugged transistors with wider safe operating areas than the more modern epitaxial-base transistor types, in terms of forward-biased SOA. But this fabrication has been made obsolete.

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 colour images on the TV screen, and contour maps of isothermal areas. You will learn a lot from those pictures. About twice a year, I wish I could borrow one.

One other point about using bipolar power transistors is that there are two major fabrication structures: the epitaxial base, and the planar structure pioneered by Fairchild Semiconductor[4].

Transistors fabricated with the epi-base structure are usually more rugged and have a wider safe-operating area, while planar devices feature faster switching speeds and higher frequency response, but aren't as rugged. Comparing data sheets for the Motorola *2N3771* and the Harris *2N5039*, shows that the *2N5039* planar device has a current-gain bandwidth ten times greater than the *2N3771* epi-base device. The *2N5039* also has a switching speed faster than the *2N3771* when used as a saturated switch, but the *2N3771* has a considerably larger safe area if used for switching inductive loads. So select the characteristics preferred and order the type needed.

But be careful: breadboard with one type and build in production with the other and you might suddenly find that the bandwidth of the transistor has changed by a factor of ten (or a factor of 0.1) or that the safe area does not match that of the prototypes.

Also be aware that the planar devices, like the familiar *2N2222* and *2N3904*, are quite capable of oscillating at high frequencies in the dozens of megahertz when operated in the linear region. So plan to use beads in the base and/or the emitter, to quash oscillation. The slower epi-base devices don't need that help very often.

There is one tricky problem. Originally the old *2N3771* was a single-diffused part. If you wanted to buy an epi-base part, that was the *MJ3771*. But now ordering a *2N3771*, will bring the epi-base part, which *does* meet and exceed the Jedec *2N3771* specs. But it exceeds them a lot more than might be expected – eg the current-gain-bandwidth is 10 or 20 times higher. So, when replacing an old *2N3771* with a new *2N3771*, bear in mind that they are probably not very similar at all.

Expert power circuits

For many power circuits, transistor choice may not be as clear-cut as in the previous examples, and there are many tricky problems that can challenge even the most experienced designers. For example, adding small ballasting resistors to ensure current sharing between several transistors may still require some transistor matching, which is not easy. Operating conditions need to be considered; decisions made on which parameters, such as beta and V_{BE}, need to be matched; and a way found to avoid the mix-and-match of different manufacturers' devices. Such design questions are not trivial. When the performance or reliability of a power circuit is poor, it is probably not the fault of a bad transistor, but quite possibly the fault of a bad or marginal driver circuit or an inadequate heat sink.

A possible scenario goes something like this. Build ten prototypes, and they seem to work; build 100 more, and half do not. But did they ever work? The circuit design may have been a marginal one and perhaps the prototypes did not really work all that well. If they are still around, it would be useful to check to see if they had any margin to spare. If the ten prototypes had a gain of 22 000, but the current crop of circuits has gains of 18 000 and fails the minimum spec of 20 000, the new units should not be called failures. It is just that expectations were unrealistic.

Mosfets avoid secondary breakdown

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

Mosfetsswitch faster than bipolar transistors, with smaller drive requirements, and mosfets 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[5]. So at high voltages and low current densities, watch out for this possibility.

At high enough V_{DS}, 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 transient milliamps may be needed to turn the gate on or off quickly, unlike with a bipolar transistor, a lot of amps are no 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 exceeding. 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 or 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 task considerably, while improving reliability. ■

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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 much more efficient at doing 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 is on. 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 this 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-compiled, 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 down-loaded via a Braille reader, a voice synthesiser, or a special large print screen.

Initially the reader 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 marking 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 feed. 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 domestic 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.

Macs use 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 be opened from either. "We are changing the dynamics of the broadcast industry", says Schure.

Already 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.

Although 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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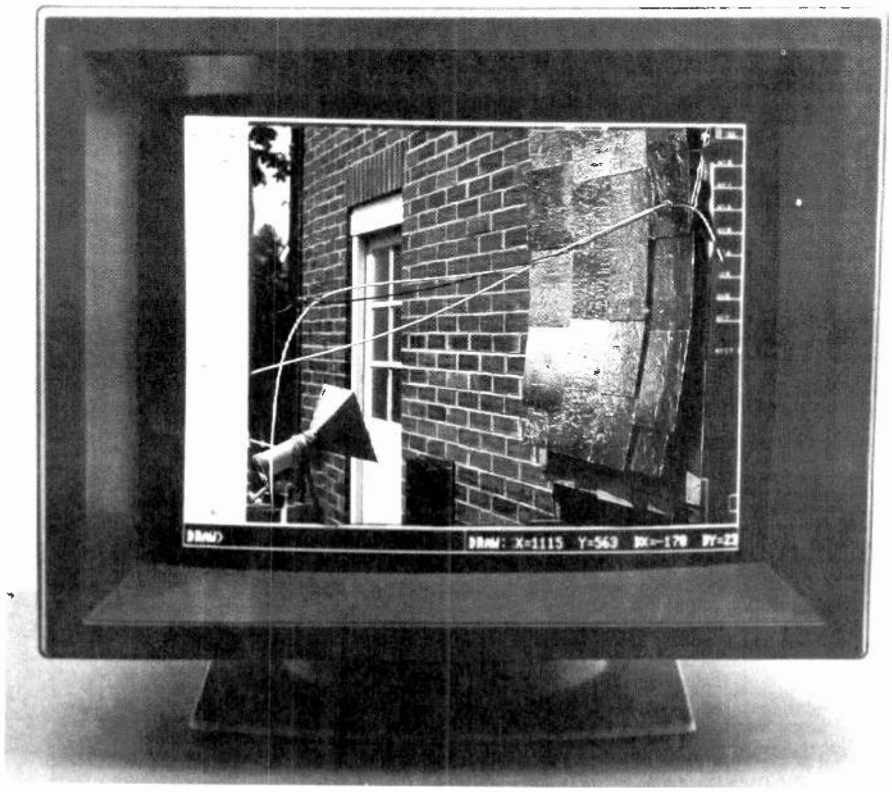


Amateur television is no longer just for the dedicated enthusiast. John Cronk explains how widespread repeaters combined with off-air beacon signals and receiver designs – such as his own – make amateur TV accessible to a much wider audience.

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.

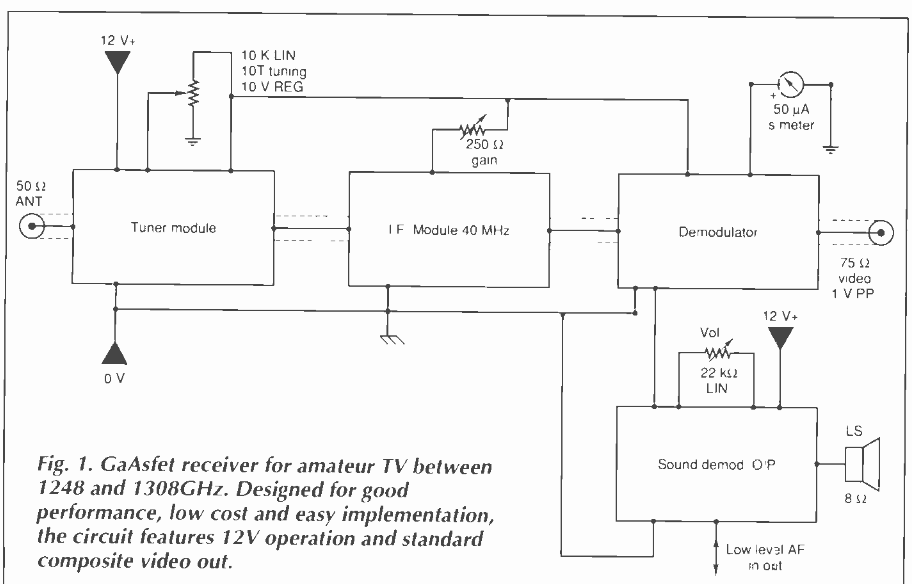


Fig. 1. GaAsfet receiver for amateur TV between 1248 and 1308GHz. Designed for good performance, low cost and easy implementation, the circuit features 12V operation and standard composite video out.

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 50Ω i/o, and at 16mA offers 17dB gain at 1GHz.

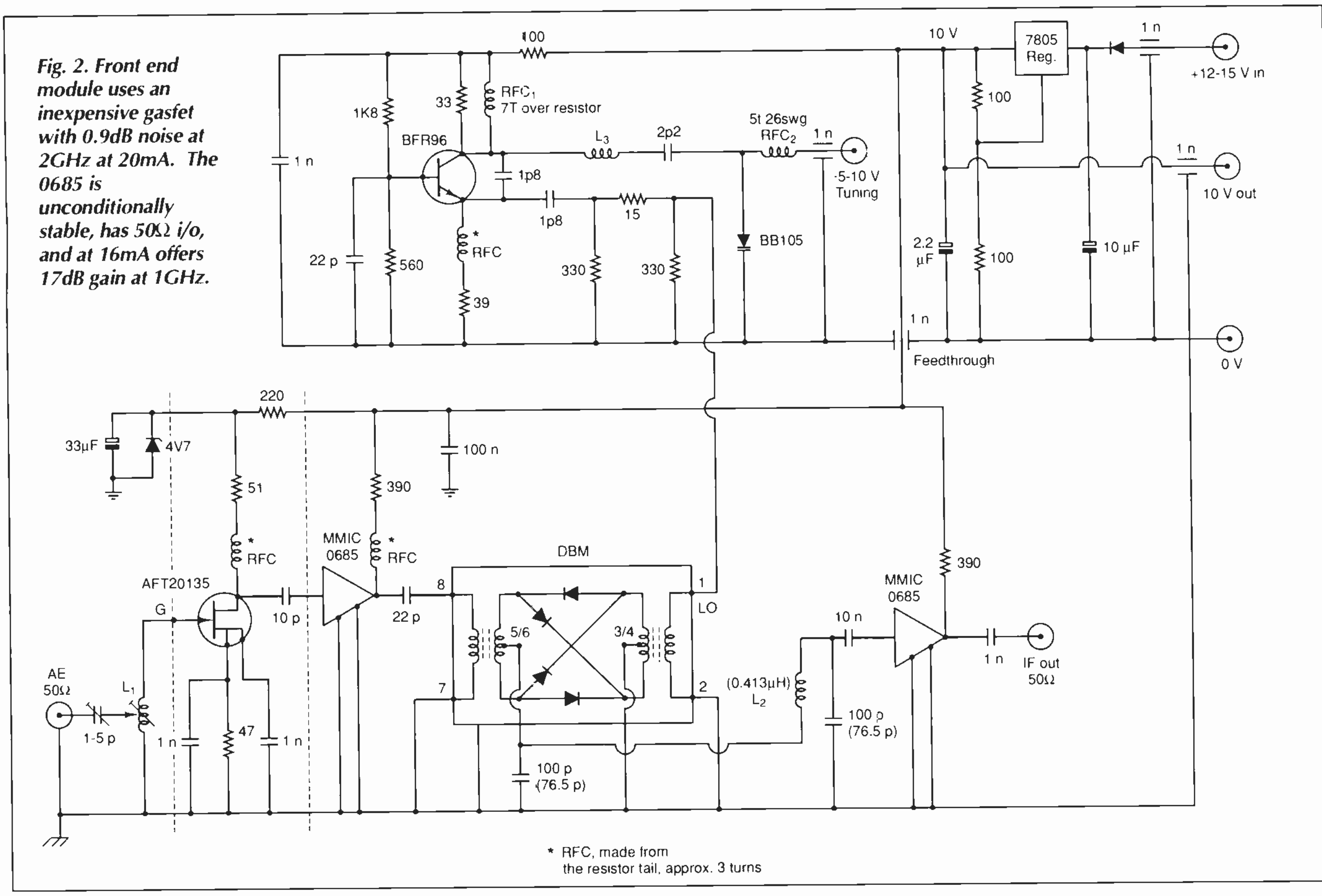
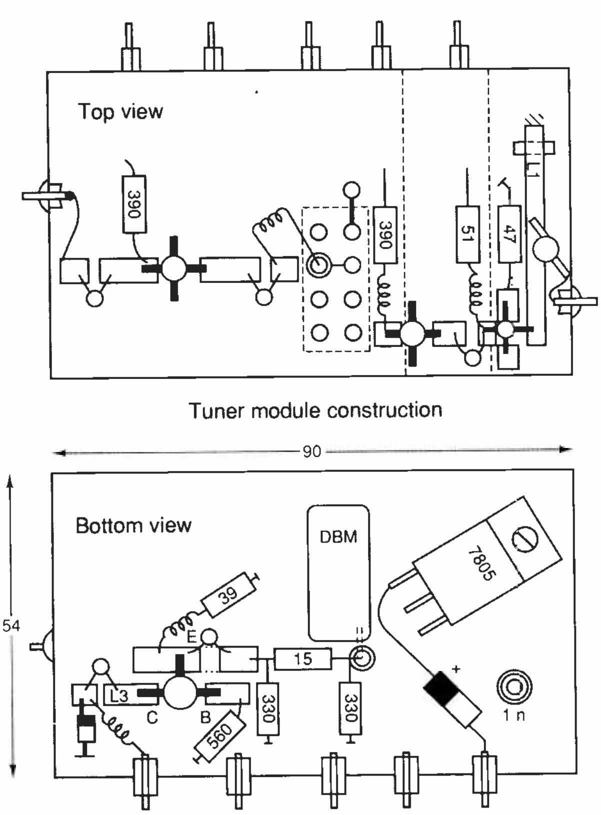
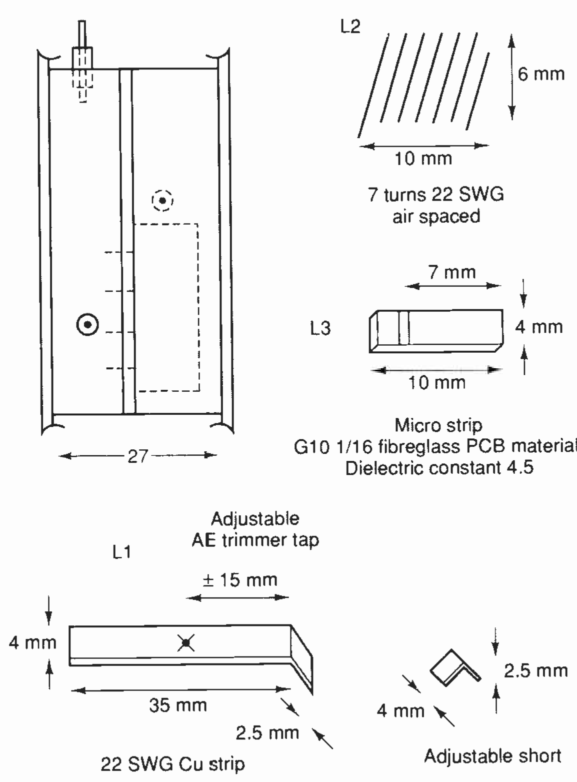


Fig. 3. Tuner layout shows compartments conveniently screened using PCB. Microstrip is lengths of PCB of set width positioned using adhesive.





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 I, 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**.

Initially I used automatic gain control but the final circuit has manual gain control. Due to the internal limiter in the *NE564* demodulator, adjustment is not critical on normal strength signals, but the control can improve signal-to-noise ratio on very weak signals. It also increases the range of the signal strength meter. The meter can be left out, but it is useful for adjusting antennas as quite considerable changes in signal strength are not always apparent on an FM modulated TV picture.

I compared the performance of this receiver with a leading commercially manufactured ATV receiver. A 10mW beacon transmitter with a 6cm quarter-wave ground plane antenna mounted indoors was used to send a test signal across town. This very weak signal reported as "just identified" was found clearer and less noisy on the receiver described here. Incidentally, a 1W transmitter and small outdoor antenna over this path give perfect colour pictures with impressively low noise – better than many AM broadcast pictures.

Being modular, the design allows each section to be tested and set up separately. Since this is primarily a design article, precise alignment details are not included but anyone with experience of similar circuitry should have few problems.

Amateur TV repeaters and frequencies

Callsign	Channel	Location	Status, April '92
GB3CT	RT2	Crawley, East Sussex	Operational
GB3ET	RMT2	Emiley Moor, Yorkshire	Operational
GB3GT	RT2	Bellahouston, Glasgow	Temp. U/S
GB3GV	RT2	Leicester	Temp. Low Power
GB3HV	RMT3	High Wycombe, Bucks.	Temp. U/S
GB3LO	RT2R	Lowestoft, Suffolk	Temp. Low Power
GB3NV	RT2R	Nottingham	Operational
GB3PV	RMT2	Cambridge	Operational
GB3RT	RT2	Rugby	Operational
GB3TG	RT103 (10GHz)	Milton Keynes	Linked to GB3TV
GB3TN	RT2R	Fakenham	Operational
GB3TT	RT2R	Chesterfield	Operational
GB3TV	RMT2	Dunstable Downs	Operational
GB3UD	RT2	Nr Stoke-on-Trent	Operational
GB3UT	RT1	Bath, Avon	Operational
GB3VI	RMT1	Hastings	Operational
GB3VR	RT2	Brighton	Operational
GB3ZZ	RT2	Bristol	Operational

Channel	Input frequency	Output frequency	Notes
RMT1/RT1	1276.5 MHz	1311.5 MHz	FM or AM Input
RMT2/RT2	1249.0 MHz	1318.5 MHz	FM OnlY
RMT3/RT3	1248.0 MHz	1308.0 MHz	FM Only
RT2R	1249.0 MHz	1316.0 MHz	FM OnlY
RT103	1025.0 MHz	10150.0 MHz	FM Only

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

GaAsfet tuner module

At the heart of the RF amplifier front end is an inexpensive Avantek *20135* GaAsfet, **Fig. 2**. According to the fet 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

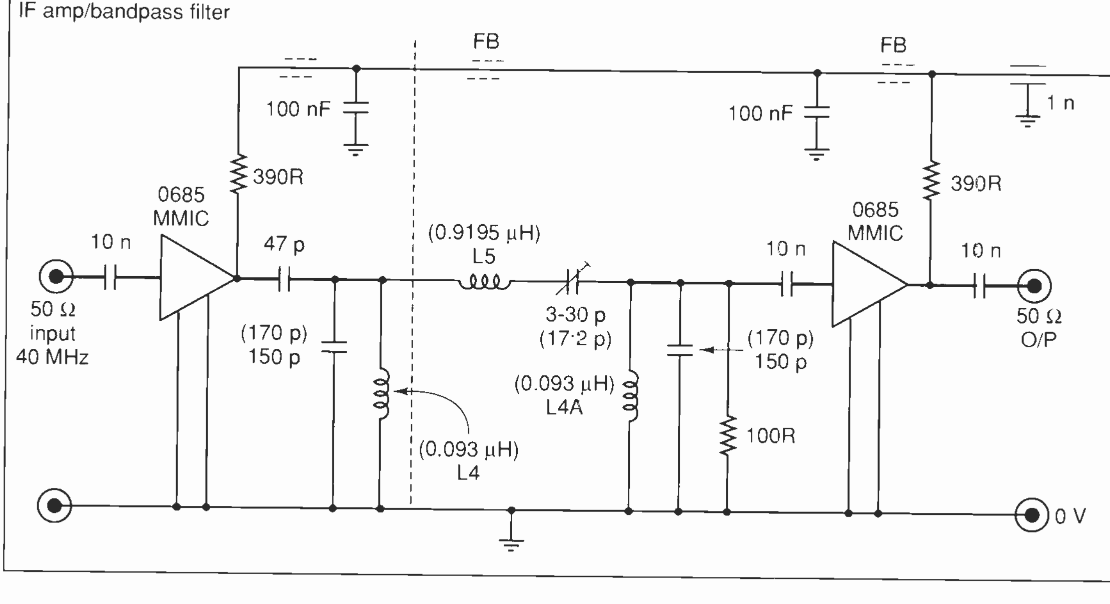


Fig. 4. In the intermediate-frequency amplifier, an easy to align Butterworth band-pass filter with adjustable gain centres on 40MHz. Values in parenthesis are theoretical.

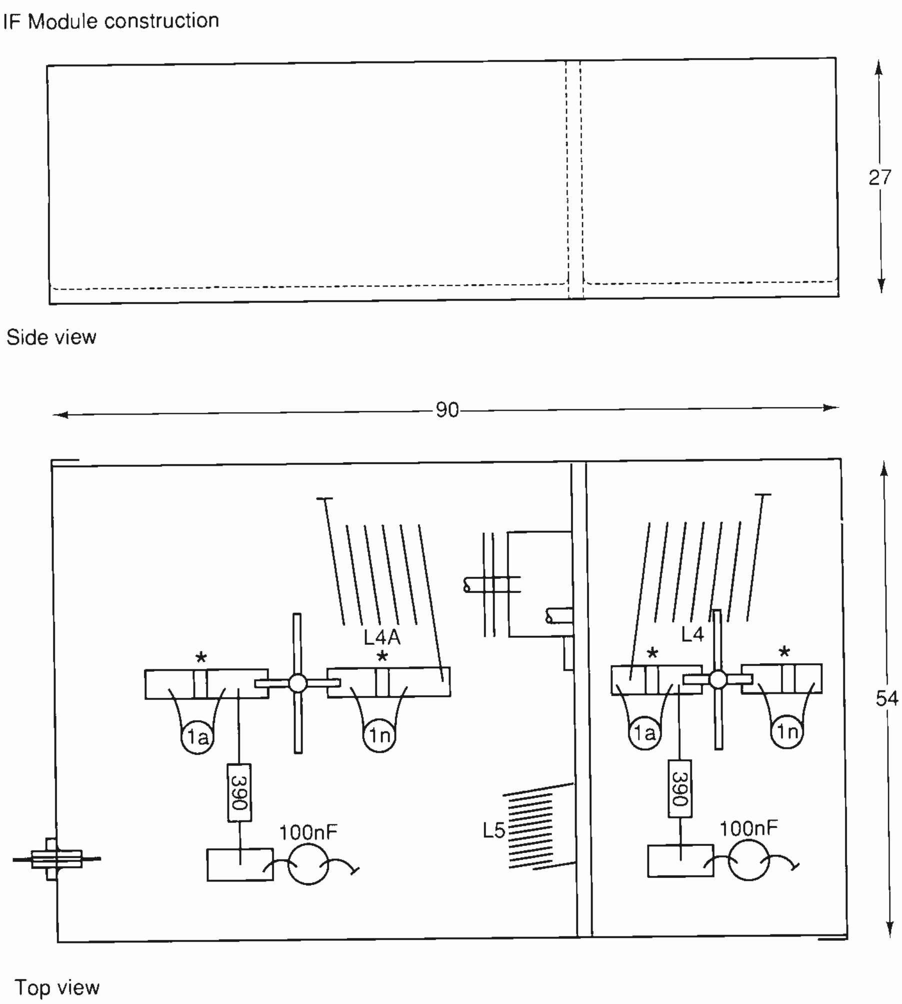
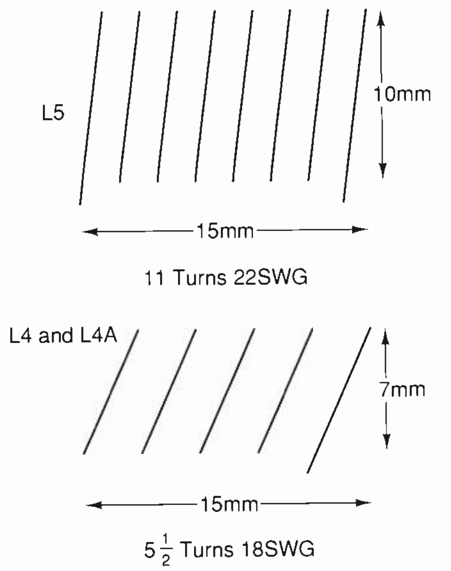
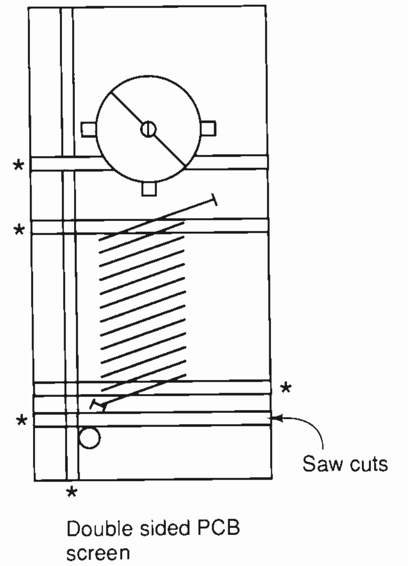


Fig. 5. Layout details of the IF amplifier with bandpass filter. Again, PCB is used for screening but here, one side of a double-sided screen conveniently carries components.





antenna it provides some rejection to out-of-band signals.

A small choke and resistor form a non-resonant load that couples the fet to the *MSA 0685 Modamp*. 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[1] the well known *SBL1-X*, also rated at 1GHz, has proved satisfactory at 1.3GHz.

For the first IF amplifier another MMIC feeds through a 50Ω low-pass pi 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.

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 mechancial details.

The 50Ω microstrip is made from 1.6mm thick glass-fibre single sided PC (G10, .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, L_3, 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 cyanacrylate 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 1208MHz.

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, a series resistor 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 amplitude insensitivity of the FM detector, will appear to have greater bandwidth.

Layout details

This module is the same size as the tuner, but the ground plane is single sided and soldered at the bottom of the module, in **Fig. 5**.

An inter-stage shield can be made from double-sided PC, one side of which has its copper divided with some saw cuts as shown to mount the series filter components. Left intact, the other side acts as the shielding.

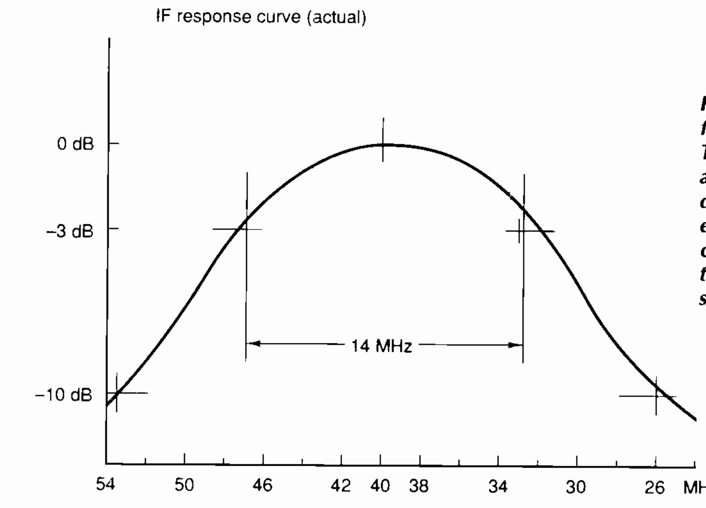


Fig. 6. This curve for the amateur TV receiver IF and filter section, derived empirically, has characteristics typical of FM systems.

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 amplitude 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.

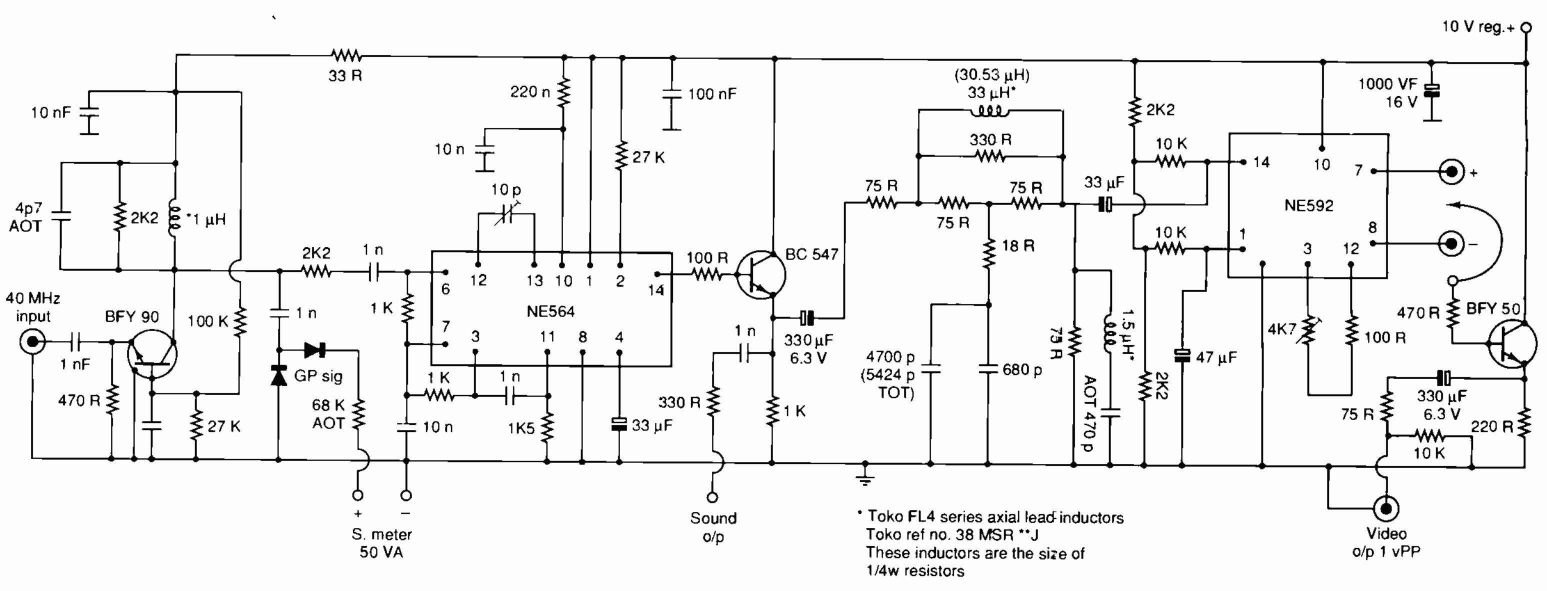
The output is intended to feed a colour or monochrome monitor with 1V composite video when terminated with 75Ω.

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 need only be about 50 by 90mm, the components are close together so

Fig. 7. At the heart of the demodulator section is a proven NE564 PLL configuration feeding a de-emphasis network for correcting video to CCIR 405-1 standard.



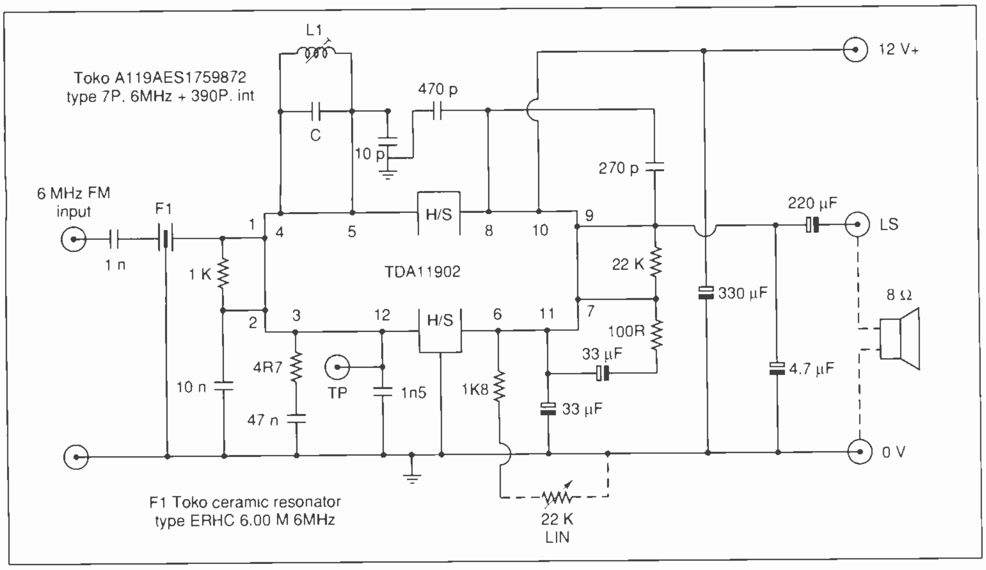
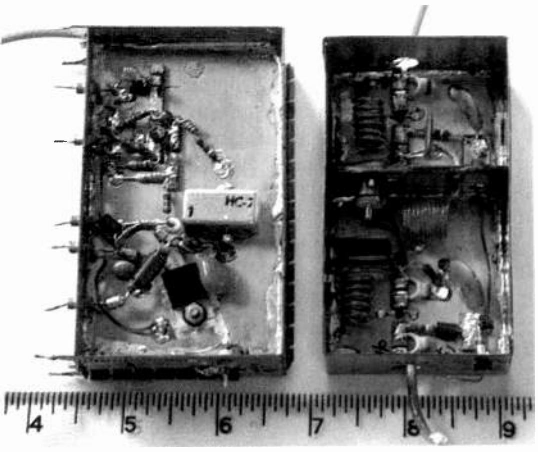
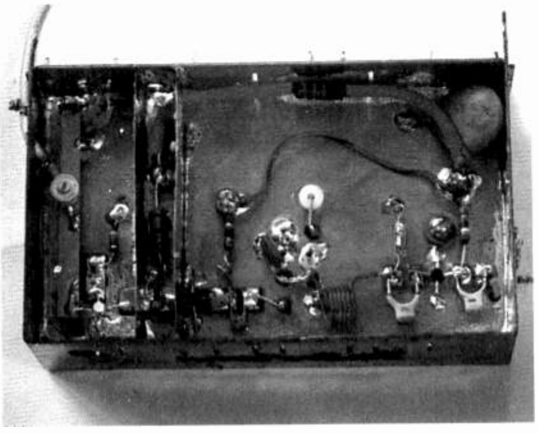


Fig. 8. In the sound section, a single IC converts 6MHz FM from the demodulator into 1.5W of audio power.



Top: Layout of oscillator components (left) and the IF module.

Below: RF side of tuner module.



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.

It is possible to fit this unit on a circuit board of similarsize to that of the video circuit. Used for long periods at high power levels, the IC might need additional heat-sinking but for average use, the ground plane of a double-sided PC should be adequate.

In the prototype, the detector coil between pins 4 and 5 was a Toko product but a custom-built coil is adequate. Note that the Toko coil specified has a built in parallel capacitor so there is no need to fit the external one shown.

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 trap at the de-emphasis network output 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. As the input circuit is loaded by the antenna it is not at all sharply tuned.

■

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 time. Signals can be ducted between these layers with low losses over remarkable distances.

Signals can also reach the other side of large obstructions such as hills or mountains if the angle is shallow by means of an effect called edge diffraction.

Heavy rain showers can obstruct a normally good path. Scatter from a rain shower however may open a new path elsewhere. While it is broadly true that signal paths on 1.3GHz are limited to line of sight, this should not be read too literally – 1600km contacts have been reported under favourable conditions.

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.

Specialist microwave component suppliers

Bonex Ltd., 12 Elder Way, Langley Business Park, Slough, Berkshire, SL3 6EP. Tel: 0753 49502

Mainline Electronics, PO Box 235, Leicester, LE2 9SH. Tel: 0533 777648.

Piper Communications, 4 Severn Road, Chilton, Didcot, Oxon, OX11 0PW. Tel: 0235 834328.

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27S191 PROM £2
IMS1400P-45 .... £2
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P8749H MICRO £5
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D41257C-15 (256K x 1 DRAM) .... £1 ea
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USED 41256-15 .... £1
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For most small users, Seetrax Ranger1 provides a sophisticated system at an affordable price. It is better than EasyPC or Tsien's Boardmaker since it provides a lot more automation and takes the design all the way from schematic to PCB - other packages separate designs for both, that is, no schematic capture. It is more expensive but the ability to draw in the circuit diagram and quickly turn it into a board design easily makes up for this.

Source JUNE 1991 Practical Electronics

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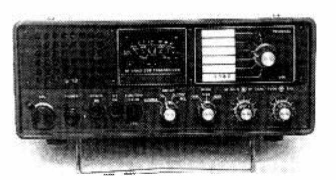

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CIRCLE NO. 125 ON REPLY CARD

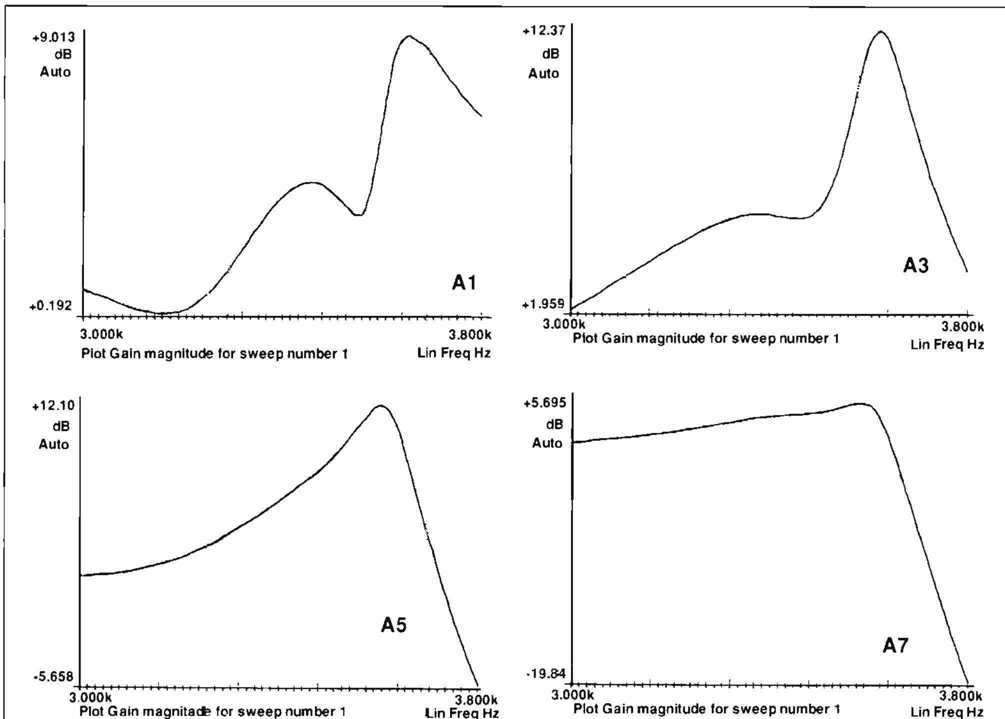
Filter tips

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.

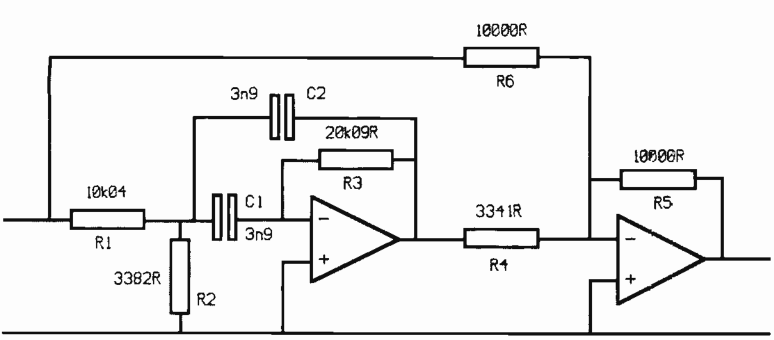


Frequency response at the op-amp outputs at each stop. See **Filter tips.**

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 $F_s/2$ for 8kHz sampling and is easily scaled.

David Markie
Ascot
Berkshire



Sin(x)/x loss correction circuit. Sampling frequency is 8kHz. With values shown resultant error is <0.1dB.

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^{-43}s, and the ratio of gravitational force to electric force of 2.4 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.

One is the idea that the power lines concentrate "the biologically most destructive lower energy particles". Presumably the term lower is relative: low compared to the 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 measured, 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 somehow averaged because the particles travel so fast they respond only to the instantaneous fields. The average value of the alternating electric 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 concentrating 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 rather than "of" normal. I assume that the figure is meant to show half a symmetrical distribution, but perhaps not.

Does not 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 I ▶

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 transoceanic 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 0.02λ 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 was 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 empirical experiments. 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 have been used for ULF signalling instead of Sanguine type antennas with linear elements 160km long.

George Pickworth
Kettering

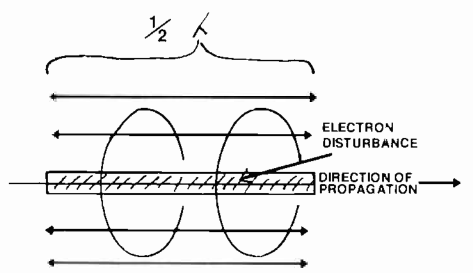


Fig. 1. Electric and magnetic fields surrounding an electron disturbance propagating along a travelling wave antenna

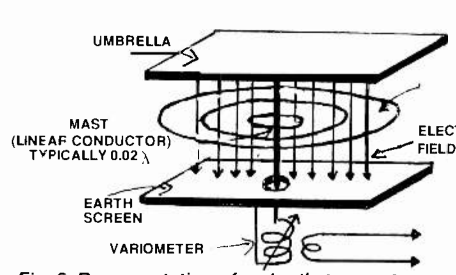


Fig. 2. Representation of umbrella type antenna indicating concentric magnetic field lines.

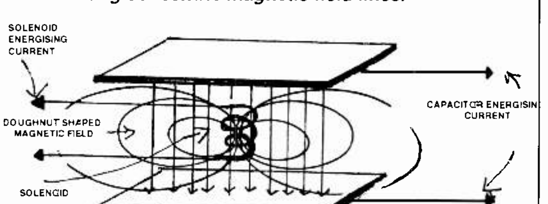


Fig. 3. Effect of substituting a solenoid in place of the linear conductor.

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 II-F3	TV Plus
Tele-X	NRK (Norway – actually D-mac)
Intelsat 512	SVT TV2 (Sweden, D-mac)
	SVT TV1 (Sweden, D-mac)
TV-sat/TDF/Olympus	MCM
	Euromusique
	RTL plus
	Sat 1
	Arte (German)
	3Sat
	France2
	ARD
	Raisat

Ragner Otterstad
Hostterkobvej 10
Denmark

Happiness is an engineer

Against all odds I find myself in full agreement with your editorial "Paying 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 keep 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
South Africa

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*, December 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 leukaemia; 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.

Whatever 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 leukaemia.

John Firth
Ottawa

Rationalising peace

The Scientists for Global Responsibility (SGR) organisation not only warns 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 continuing support for Alasdair Philips' research on health effects of electromagnetic fields, especially from electric power lines. The other is to have a presence at this year's Milcomp (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 Cottey
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 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 long driven into the ground about 10m apart.

It was recognised that the aluminium would oxidise with time but this was all that was available.

A matching transformer was employed and as I recall the resistance was around 100Ω and varied with the water content of the soil. The tappings were changed according to the weather.

From this I progressed to using a heavily stressed pair of 807 valves which gave some 75W output. A friend and I patrolled the area with an amplifier and a pair of metal rods which we would drive a few centimetres into the ground some 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 27m 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.

Rod Brown
Nottingham

Crash response

With reference to the letter "Crash solution" (*EW + WW*, November 1992) asking for clarification of C_x and C_y capacitors and flux cancelling inductors.

$\underline{Y}$ 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, again for safety reasons.

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 thinner dielectric is acceptable 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 through one and the neutral current back through the second. The flux then cancels, the core is not saturated, and we have a cheap high value inductor, typically 20mH. This together with the *Y* class capacitors gives good attenuation for any common mode (asymmetric mode) interference, that is a spike that appears on live and neutral together.

Of course any differential or symmetric mode interference that is on line and neutral but opposite polarity produced between the lines will suffer the same flux cancelling as the main current, that is only a small inductance will be seen, typically 0.1mH. 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 dependant resistors to clip spikes and larger values of inductors and capacitors. BS613 allows up to 2μF on permanently connected household apparatus.

John Foster
Enfield

■

Integration at the helm: Self-contained instruments, linked together to talk to each other are the basis of Brookes & Gatehouse's Network system. The pod contains, from left: VHF; Quad (speed, log, timer, depth); wind; GPS; autopilot. Instruments can be easily switched as new facilities – perhaps DGPS – become available.

Greater GPS accuracy and a compact satellite voice telephone were among the electronic innovations unveiled at this year's London Boat Show. By Peter Willis

ELECTRONICS AFLOAT

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.

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

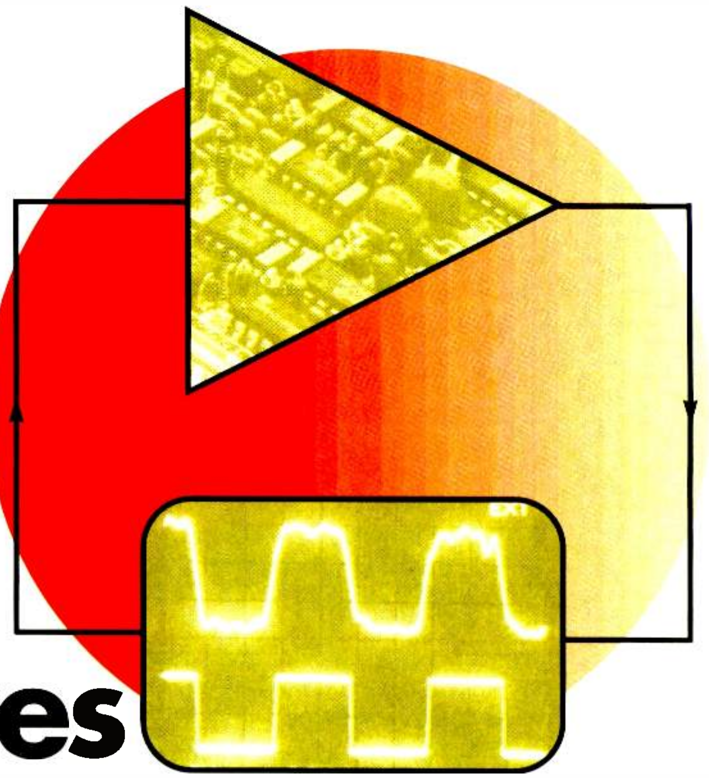
LCD display (640 x 480 pixels) for Trimble's NavGraphic XLGPS. Positions can be plotted direct to screen on the CD-sourced charts, while instrument readings are displayed on the bottom of the screen.

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 the picture – so BT is now working on error-correction capability.

Development deadline bas been put back from January to May, this year. ■



Bryan Hart describes a graphical technique for describing feedback amplifiers that reveals fundamental operating characteristics.

Op amps in pictures

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_0. The sinusoidal input, error and output voltages are represented by the phasor quantities V_i, ε and V_o respectively. Phasor symbols are sometimes shown in bold face in texts on circuit theory, but there is no advantage to that 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,

$$\varepsilon = \frac{(V_i - V_o)R_2}{R_1 + R_2} + V_o$$

reducing to

$$\varepsilon = aV_i + bV_o$$

where, by definition

$$a = \frac{R_2}{R_1 + R_2}$$

$$b = \frac{R_1}{R_1 + R_2}$$

However the amplifier gives

$$V_o = -A_0\varepsilon$$

so by combining equations

$$aV_i + bV_o = \varepsilon = -\frac{V_o}{A_0}$$

Let $V_o = -V$ and we obtain

$$aV_i - bV = \frac{+V}{A_0}$$

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 ε 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.

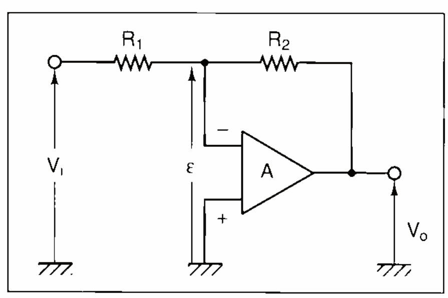


Fig. 1. Basic amplifier configuration

- Mark point *L* at $V_o = -V$ on iii, point *M* at V/A_0 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 point *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:

$$JQ/HP = R_2/(R_1 + R_2) = a$$
$$JS/KL = QM/KL = R_1/(R_1 + R_2) = b$$

Thus, the phasor interpretation of the horizontal line sections is $KL \equiv V_o$ $(= -V)$; $JS = QM = bV'_o$ $(= -bV)$; $JM = \varepsilon$; $HP = V_i$ and $JQ = aV_i$.

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, eg 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 $|A_{CL}(0)|$, magnitude of the DC and low frequency closed-loop gain.

An example is calculation of $|A_{CL}(0)|$ for the case R_1 = 10kΩ, R_2 = 20kΩ, A_0 = 10. **Figure 3**, not to scale indicates, the method. Construct i, ii and iii with a vertical scale 5kΩ ≡ 1cm. The choice V = 50mV gives an integral value (5mV) for ε. 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 $|A_{CL}(0)| = 50/32.5 = 1.54$ – which agrees with the value obtained by a purely algebraic calculation.

Suppose V_0 had been equal to –30mV instead of –50mV. The modified construction line $L_1M_1P_1$ 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, 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 $|A_{CL}(0)| = NK/HN$.

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_0 for V_i = 10mV, say, then mark P_2, draw P_2N and extend it to L_2 and measure the resulting output voltage KL_2.

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_2 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.

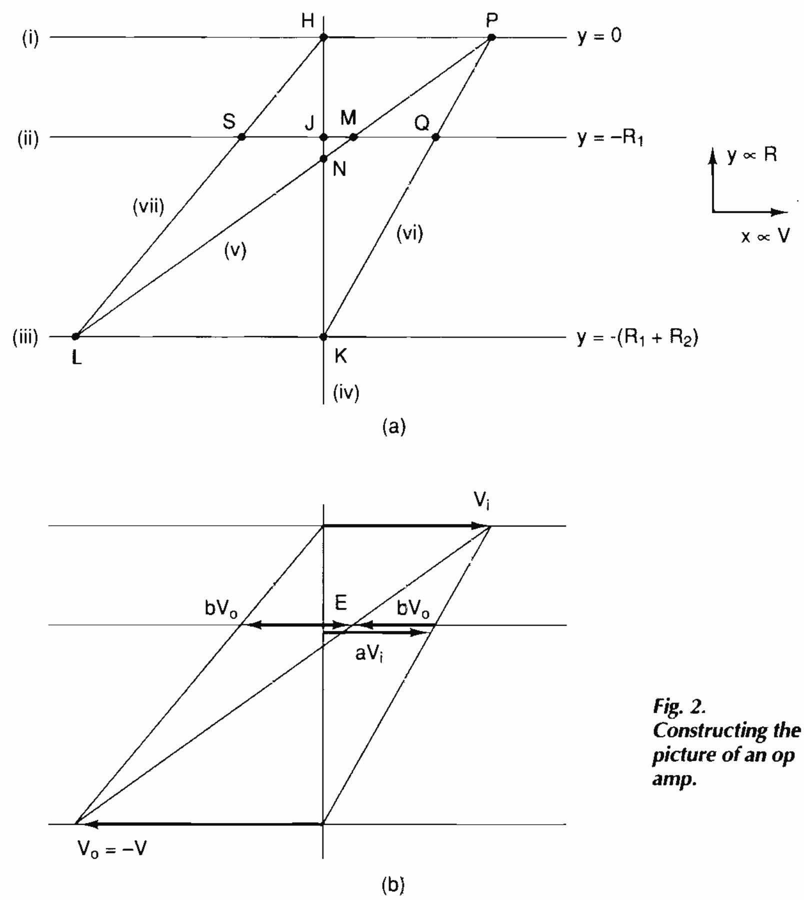


Fig. 2. Constructing the picture of an op amp.

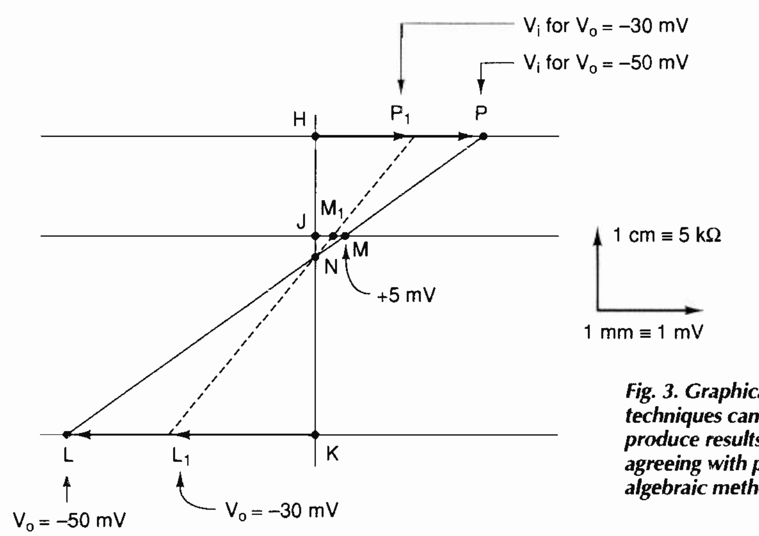


Fig. 3. Graphical techniques can produce results agreeing with purely algebraic methods.

Ideal and practical op-amp

Introductions to the op-amp inverter circuit, Fig. 1, often show $\varepsilon = 0$, 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 theoretical argument need some interpretation in practical electronics.

So how large must A_0 be (or, how small must ε be) for ε to be taken as zero without significant error in the calculation of $|A_{CL}(0)|$? 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, providing some physical insight into the mechanism.

In **Fig. 5** $JM(\varepsilon)$ is the phasor sum of two components: JQ $(= aV_i)$, the signal fed in from V_i, with $V'_o = 0$; and QM $(= bV'_o = -bV)$, the signal fed back from V_0 for $V_i = 0$. ε 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 ε $(= V/A_0) \ll bV$, or $A_0 b \gg 1$. This condition is independent of magnitude of the input signal amplitude and relates to the amplifier configuration as a whole, ie, A plus resistor network, and not just A by itself. Hence, the inequality $A \gg 1$ – a plausible

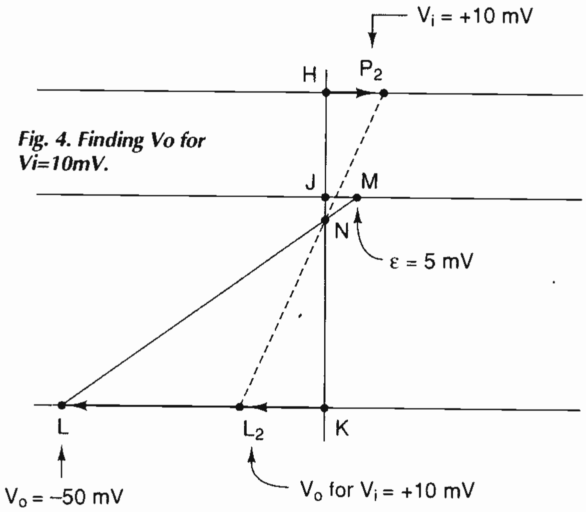


Fig. 4. Finding Vo for Vi=10mV.

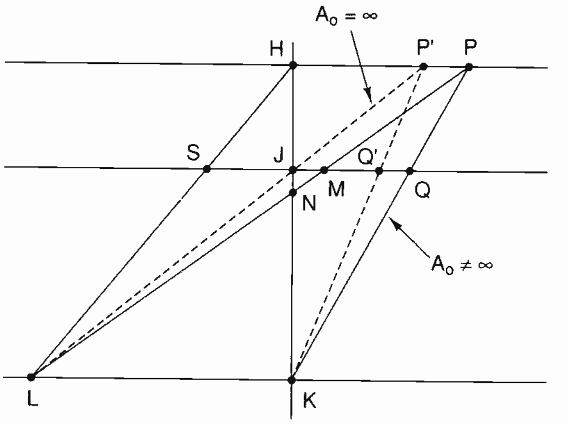


Fig. 5. LP' and KP' refer to the ideal case where A0b=∞.

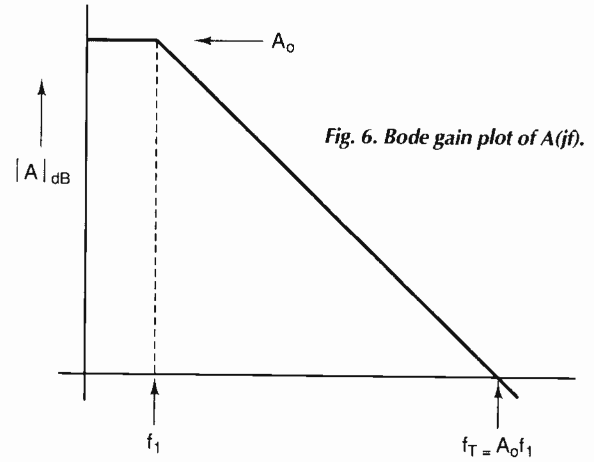


Fig. 6. Bode gain plot of A(jf).

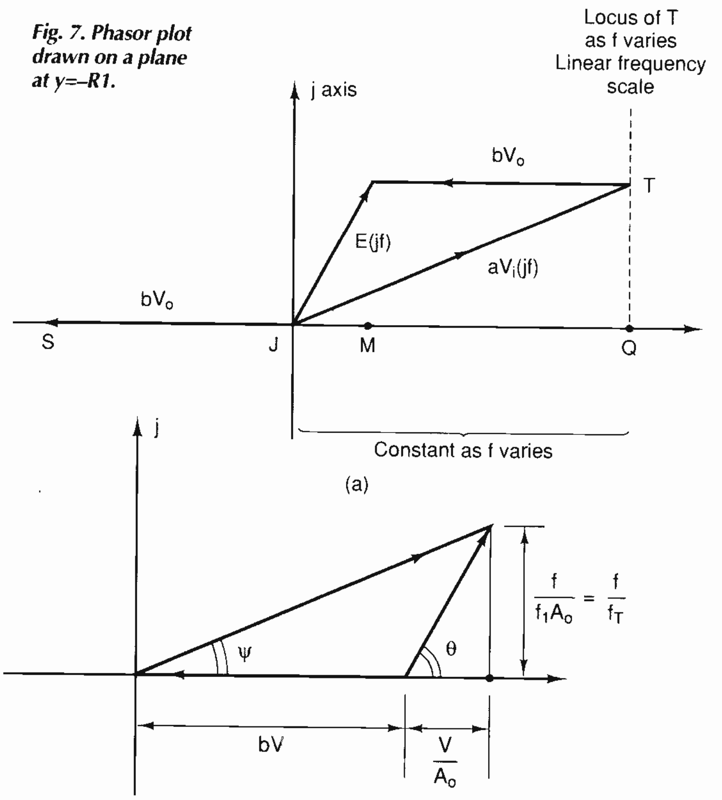


Fig. 7. Phasor plot drawn on a plane at y=–R1.

prior assumption – is a necessary but not a sufficient condition.

For most practical op amps with $A_0 >$ 100,000, the condition for ε 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 $A_0 b = \infty$. For that case J, M and N coincide. Then, $aV_i + bV_0 = 0$,

$|A_{CL}(0)| = (|V_0/V_i|$ for $A_0 b = \infty) = b/a = R_2/R1$

The fractional error in $|A_{CL}(0)|$, made in assuming $A_0 b$ is infinite when it is finite (though perhaps large), is given by,

$$\frac{|\Delta A_{CL}(0)|}{|A_{CL}(0)|} = \frac{JM}{JQ} \approx \frac{JM}{MQ} = \frac{1}{A_0 b}$$

In practice the near-equality in magnitude of the two opposing components aV_i and bV_0 has an interesting parallel in solid-state physical electronics. For a forward-biased 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:

$$A(jf) = \frac{A_0}{1 + j\left(\frac{f}{f_1}\right)}$$

in which A_0 = DC gain and f_1 is the –3dB roll-off frequency. A Bode gain plot of this equation is shown in **Fig. 6**. If $|A(jf)| = 1$ (0dB) at $f = f_T$ ($>>f_1$) it follows that $f_T = A_0 f_1$

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

$$aV_i + bV_o = \varepsilon = -\frac{V_o\left(1 + j\frac{f}{f_1}\right)}{A_0}$$

Using $f_T = A_0 f_1$ and $V_o = -V$

$$aV_i - bV_o = \varepsilon = \frac{V}{A_0} + j\frac{f}{f_T}$$

Interpreting this graphically calls for a j axis for the quadrature component of ε. This can be drawn perpendicular to the plane of the page, with the positive direction of j pointing upwards. Phasors for V_i, ε and V_o can then be shown on planes erected, respectively, at y = 0, $y = -R_1$ and $y = -(R_1 + R_2)$.

Looking first at the phasor plot drawn on a plane at $y = -R_1$ (**Fig. 7a**), aV_i is the component of V_i required to maintain bV_o constant at $-bV$ as f varies. Relocating phasor ε produces **Fig. 7b** which simplifies discussion.

θ and ψ are the frequency-dependent phase shifts due to A and the feedback amplifier as a whole. Clearly, $\psi << \theta$ if $b >> (1/A_0)$, illustrating the benefit of phase shift reduction with negative feedback.

The –3dB roll-off frequency, f_1', for closed-loop gain occurs when $\psi = 45°$. From Fig. 7b the condition $b >> 1/A_0$ gives:

$$\frac{f_1'}{f_1} = b$$

Now

$$b = \frac{R_1}{R_1 + R_2} = \frac{1}{1 + |A_{CL}(0)|}$$

So for $|A_{CL}(0)|_\infty >> 1$

$$f_1'|A_{CL}(0)|_\infty = f_1 = \text{constant}$$

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 $y = -R_1$ there is a plane at y = 0 for the display of V_i. A line, rather than a plane, is sufficient at $y = -(R_1 + R_2)$ because V_o 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 and 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 back flat on to the page. Then the y direction is used, as in Fig. 2a, for the display of R and also – though the scale is completely unrelated – for the quadrature 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. ■

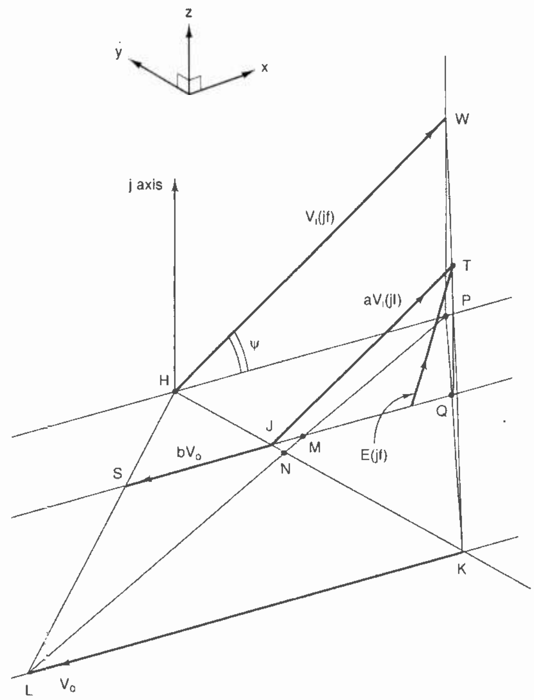


Fig. 8. Three-dimensional equivalent of Fig 2a.

Acknowledgement

Thanks to R H Pearson for perceptive comments.

NEW PRODUCTS CLASSIFIED

ACTIVE

Asics

200MHz FPGAs. AT&T's *ATT3000* series of field-programmable gate arrays now run at 200MHz – a performance attributed to the use of the company's 0.6micron salicided 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 *HG62G* 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 *HG62G* family offers sub-nanosecond operation (0.3ns for a 2-input power Nand) and operates from 2.7V to 5.5V rails. Hitachi Europe Ltd, 0628 585000.

Discrete active devices

Fast-recovery mosfets. *ABB-IXYS HiPerFET* mosfets allow the use of the *ABB-IXYS* intrinsic diode in free-wheeling applicationns due to its fast reverse recovery, which is 60% faster than standard mosfets with no increase in on resistance. The one intrinsic diode eliminates the series schottky and parallel diodes. The range includes 75A/100Z/0.02Ω up to 12A/1000V/1.05Ω types all with reverse recovery time of less than 250ns. Kudos Thame Ltd, 0734 351010.

Digital signal processor

Histogrammer. Harris's *HSP48410* 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 686886.

Linear integrated circuits

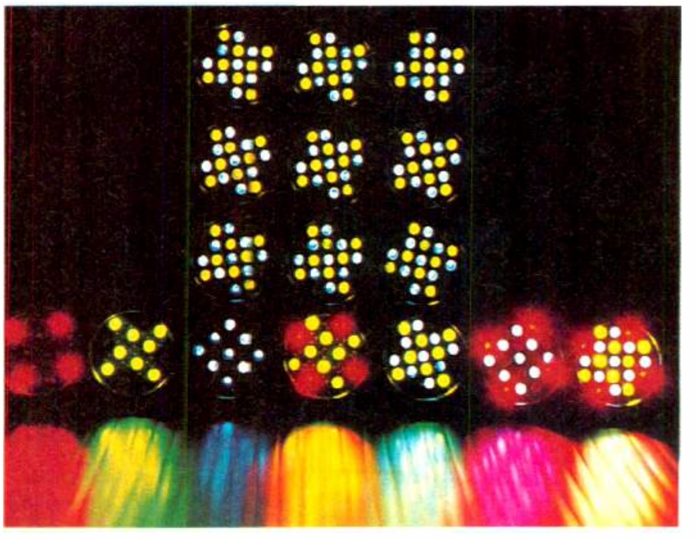
Video amplifier. Comlinear's *CLC411* is a 200MHz video op-amp intended for HDTV, composite video, line driving and D-to-A output buffering. Slew rate is 2300V/µs with a settling time to 0.1% of 15ns. To 30MHz, gain varies less than 0.05dB and diff. gain and phase are within 0.02% and 0.03% respectively.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

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 0.65µV pk-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/°C offset drift and maximum offset is 10µV. Linear Technology (UK) Ltd, 0276 677676.

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/deg.C. Input bias and offset currents are both 250pA. Noise performance of both devices between 0.1Hz and 10Hz is 0.32µVpk-pk and slew rate is 0.3V/µs. Micro Call Ltd, 0844 261939.

Triple video amplifier. Elantec's *EL4390* consists of three 60MHz current feedback amplifiers, each with a DC-restore amplifier activated by a common TTL/cmos-compatible control signal and each having a separate restore reference. Response is flat to within 0.1dB to 10MHz and slew rate is 900V/µs. Microelectronics Technology, 0844 27878[illegible].

Video distribution. A low-cost video distribution amplifier, the *EL2099* from Elantec, will drive up to six double terminated cables at a –3dB bandwidth of 60MHz and a gain of 2, giving ±11V into 25Ω and slewing at 900V/µs. Output current is 500mA and differntial phase and gain are said to be low. Microelectronics Technology, 0844 278781.



Optical devices

Bright, blue leds. Sharp's *GL5BX43* 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.

FN pager receiver. Philips claims its *UAA2080* to be the most advanced single-chip direct-conversion FM receiver for pagers. When used with the *PCF5001T* decoder the result is a two-chip wide-area beep-only pager or a high-performance display pager front end. The chip receives and demodulates FM NRZ FSK data, operating at up to 512MHz with a sensitivity of –124dBm for a bit-error rate of less than 3/100. Active and standby currents are 2.7mA and 3µA. Philips Semiconductors Ltd, 071 436 4144.

Complementary mosfet arrays. Complementary mosfet pairs in the same package from Zetex feature a 60V maximum drain/source volatge and a peak current rating of 3A. The *ZVC2016E* handles 280mA continuously and has a gate/source threshold of 3.5V at 1mA, with an on resistance of 5Ω at 10V and 500mA. Zetex plc, 061 627 4963.

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 40, 12 by 18 characters, depending on the television standard. Philips Semiconductors Ltd, 071 436 4144.

Memory chips

4Mbit eproms. AMD has announced the *Am27C4096* and *Am27C400* devices, the former being organised as a 256K by 16bit type and the latter a rom-compatible cmos device user-configured as 512K by 8bit or 256K by 16bit eprom. Both are in 0.85µ cmos and possess an access time of 120ns. Advance Micro Devices UK Ltd, 0483 740440.

4Mbit word-wide drams. Hitachi's second generation of devices, the *HM514260A* 256K by 16-bit type and the *HM514280A256* by 18-bit unit are dynamic rams in 0.7µ cmos, with access times down to 70ns. The *60A* uses only 825mW when active, reducing to 1.1mW from 5V on standby. Hitachi Europe Ltd, 0628 585000.

Mixed-signal ICs.

Electronic digital pot. From Dallas, the

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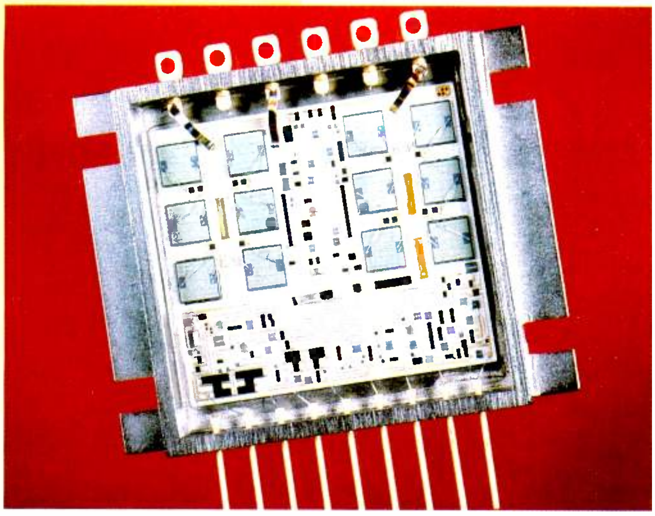


2kW linear amplifier. A 2000W linear amplifier by Apex, the *PA30*, produces up to 8kW pulse output and the 2kW continuous from zero to 40kHz. Output current is 50A continuous (100A pk) on rail-to-rail supplies of up to 200V. Power mosfet outputs avoid secondary breakdown and on-chip sensors provide thermal protection. The package measures 2.8 by 2.2 by 0.4in. Microelectronics Technology, 0844 278781.

Dallastat *DS1668/1669* is a digital intewrface and non-volatile memory, forming an alternative to mechanical potentiometers. *DS1668* has a manual interface and the *1669* is configurable for two-button opeartion, both types being digitally controlled by means of the interface. The memory retains last setting when power is removed. Values are in three ranges up to 10kΩ, 50kΩ and 100kΩ. Joseph Electronics Ltd, 021 643 6999.

Distance sensor. The *GP2D02* distane sensor by Sharp contains an infra-red led, a position sensor and signal-processing to measure distances between 4cm and 30cm or 10cm-80cm, depending on the version. Output is in the form of an 8-bit code. Current consumption is 7ma or 4.5mA, dropping to 2µA in the absence of a reflective object. Sharp Electronics (Europe), 010 49 40 23 76-0.

Power semiconductors

Power mosfets. Hitachi's fourth generation of power mosfets, the *DIV-L* series, are smaller in size and exhibits smaller conduction and switching losses. On resistance is now down to 18mΩ for a 60V device and fall time has been reduced by 30%. Reverse recovery time is typically 125ns. The range includes more than 20 types of n and p channel devices at 60V and lower breakdown. Hitachi Europe Ltd, 0628 585000.

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. *LT1105* has a totem-pole output to drive an external fet, while *LT1103* has its own fet output, the former being designed for 50W-250W output and the latter for 10W-100W operation. Linear Technology (UK) Ltd, 0276 677676.

Power mosfets. *APT10050JN* is an n-channel 1kV power mosfet from APT, rated at 520W dissipation and packaged in SOT-227. The single die is rated at 20A, with a drain/source on-stae resistance of 0.5Ω, meeting 2.5kV RMS isolation. Microelectronics Technology, 0844 278781.

Regulators. Sharp's *PQ* series of voltage regulators drops only 0.5V from input to output, reducing power loss and therefore the need for large heat sinks. Output current is up to 2A; regulation is 0.2% for a 2000-fold load current increase and stabilisation gives a 0.4% output voltage change for a two-fold input change. Sharp Electronics (Europe), 010 49 40 23 76-0.

DIP power. Power+Logic is a new range of TI peripheral drivers in dip packages — the first to combine power transistors and cmos logic on the same substrate. *TPIC6259* is an 8-bit addressable latch; *TPIC6273* an octal D-type latch; and *TPIC6595* an 8-bit shift register, all working in the range –40°C to 125°C. Texas Instruments Ltd, 0234 223252.

PASSIVE

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 insulation-displacement sockets and headers in straight and right-angle form for termination of 1mm 28awg ribbon cable. Current capacity is 1A per line. These tuning-fork contact sockets with a barb to retain the cover come in 12 to 50-way versions and offer a contact resistance of 20mΩ and insulation resistance of 1GΩ. Interconnection Products Ltd, 0433 21555.

Tough microwave cable. Internally strengthened microwave cables from Gore are small in diameter, low in weight and flexible, but withstand 175lb/linear inch, 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. A range of connectors is available. WL Gore & Associates (UK) Ltd, 0382 561511.

Filters

Chip EMI filter. Murata's *NFM51/52REMI* filters offer 100dB/decade damping and are intended for use with digital circuitry, eliminating noise from clock and video RGB circuits or that from DSP and image processing equipment. Four cut-offs are available: 50MHz, 100MHz, 200MHz and 500MHz, suppression at 5GHz still being about –40dB. Murata Electronics (UK) Ltd, 0252 811666.

Hardware

Rectangular feedthrough C. Possibly the first mass-produced rectangular feedthrough capacitor is produced by Beck. It measures 4.9 by 4 by 2.7mm and was designed to fit over the rectangular 30A pin of a car component. Values of the ceramic capacitors is 1000pF at 100V DC. Beck Electronics Ltd, 0493 856282.

Product labels. *Info-mark* and *Appliance-mark* by Donprint are label-printing devices, using on-screen design and 300dot/in printing on material capable of withstanding temperatures up to 380°C and a variety of cleaning materials including petrol,detergents and some fairly hostile chemicals. Donprint Label Systems Ltd, 0355 249191.

RF seals. Conductive, elastomeric materials for Anti-EMI sealing from Dunlop have been tested to US military spec. MIL-G-83528A. The new materials are based on silicone and fluorosilicone rubber with metallic or metal-coated fillers and reinforcements such as metallised fabric, which offers a 0.001Ω/cm resistivity. These materials will attenuate frequencies up to 18GHz by up to 100dB. Dunlop Precision Rubber, 0509 502151.

Infra-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, 0489 583858.

EMC shielding. Econoshield is RFI Shielding's new material for the commercial sector where low compreesion force is needed — round doors, for example. Shielding is effective between 30MHz and 1GHz, providing around 30dB attenuation. The material consists of polyurethane foam covered 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 ality monitor, video modulation and video/sound separation. As video and sound quality monitor, it reproduces on screen a zoomable picture, with the input signal and line trigger. It will also operate as a spectrum analyser. Feedback Instruments Ltd, 0892 653322.

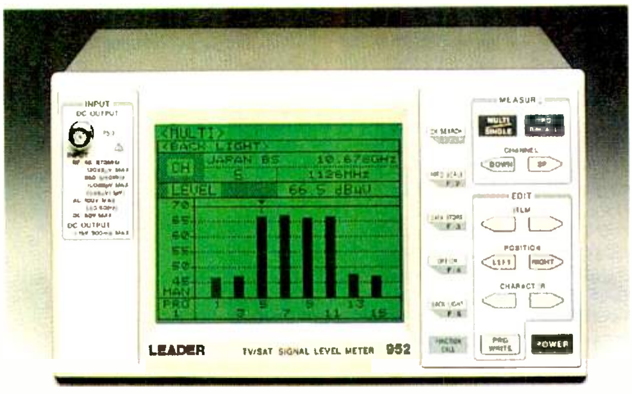
EMC components. Global introduce a range of EMC antennas and custom made chokes and filters. Antennas cover the 20MHz-1GHz frequency range in biconical and log. periodic form and are supported by most automated EMC software. Global Specialities, 0978 853920.

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 *HP70341A* 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 362867.

Digital multimeter. Saje's *7130* bench multimeter is programmed via RS232 or, as an option, IEEE488, to measure alternating and direct voltage and current, frequency and power in VA. Mathematical facilities offered are percentage, compare, difference, maximum/minimum and relative, these being intended to ease production batch testing. Saje Electronics, 0223 425440.

Multi-function measurements. *Scopemaster* 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 20Msample/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 wholething 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 bar-graph dispaly automatically scaling to respond to both snallest and largest signals in a group. Both sound and vision carriers are displayed and the instrument stores in non-volatile memory four sets of eight channels, channels from 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 89-80 08 30.

Materials

Ceramic HV capacitors. A range of capacitors from the Cera-Mite Corp. handle up to 13kV RMS at 60Hz, using a new ceramic material. Capacitance values are 400pF to 4700pF working at 10kV DC/4kV RMS at 60Hz to 40kV DC/13kV RMS at 60Hz. The material in the KT series confers improved reliability and stability, tighter tolerances and low dissipation. Acal Electronics Ltd, 0344 727272.

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. Outputs are both 0-32V, 0-2A, independent and isolated to 300V. Both operate in constant-I or constant-V 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, 0892 653322.

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. TI has the *4MM* — a bimetallic protector for re-chargeable 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 *APO1501640* 1.5-1.6GHz 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. *SILRX-418-A* is a sil-outline 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 on standby. Quantelec Ltd, 0993 776488.

Switches and relays

Low-noise relay. *CRX-12X* from BLP 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, 0992 444111.

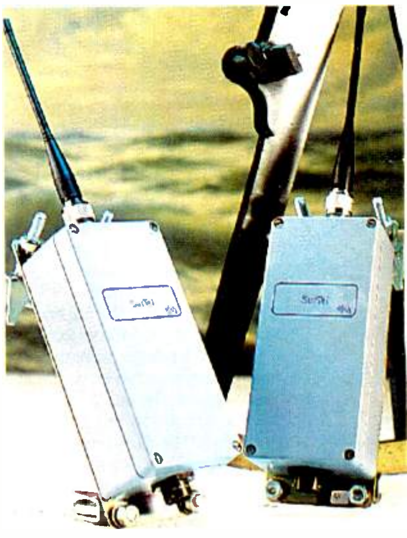
PCB connector. A surface-mounting, dual-row inverse connector from Methode has an off-the-board height of 7.1mm. It comes in two forms: one with vertical holes and the other with horizontal holes for insertin from either side, both having from two to forty positions on each row at 0.1in pitch. Contacts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

PCB switches. A new family of sub-miniature switches by Elma, thr Type *09*, are meant 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.

Transducers and sensors

Miniature gyro. Murata's *ENC-05S* *Gyrostar* 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 ±90°/s producing a 72mV DC variation about the 2.5V reference. Murata Electronics (UK) Ltd, 0252 811666.

Position transducer. *Rayelco* 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 0-20in 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 communiactions 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.

Please quote "Electronics World + Wireless World" when seeking further information

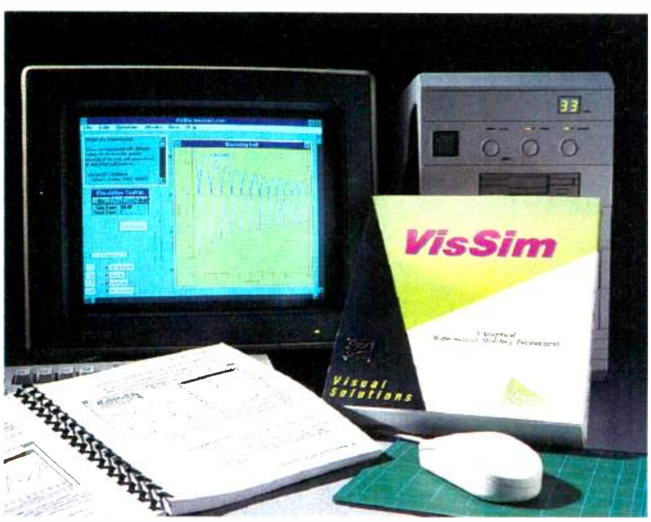
COMPUTER

Computer board level products

Data acquisition. Two boards freom Amplicon, the *C10-AD16jr/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 diff. analogue inputs, the 3μs converter acquiring and transferring at up to 100kHz. The ****jr-AT* is three times faster and is able to use the Repeat Input String for 286+ computers for data transfer 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 work on animated simulation, real-time control and dynamic analysis. It require 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 run s the simulation. Adept Scientific Microsystems, 0462 480055.



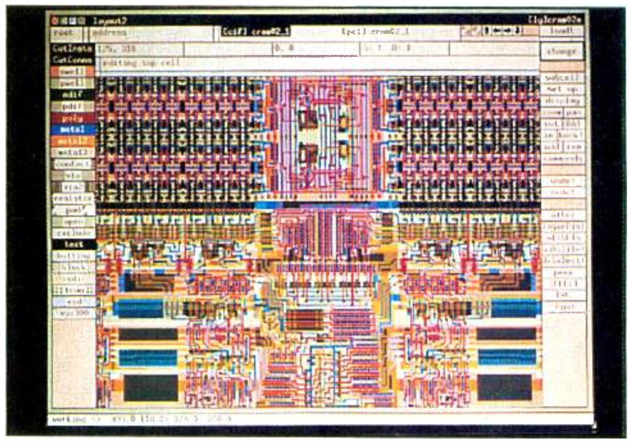
STEbus processor board. Arcom's new *SCIM486SLC* is a high-performance embedded computer based on the Cyrix *Cx486SLC* processor. It is the first *486SLC* to run on STEbus and is wholly contained on a 100 by 160mm Eurocard. The processor has a 32-bit internal data path and 1K of on-chip cache to allow risc-like single-cycle execution. A 16-bit hardware multiplier performs integer multiplication up to eight times faster than in a 386 to improve video performance by 100%. The unit runs at 25MHz and, with an SVGA adaptor fitted, is effectively a PC AT. Arcom Control Systems Ltd, 0223 411200.

Analogue, digital and timing i/o. Three multifunction boards from National for the PC offer analogue, digital and timing i/o data acquisition. *AT-MIO-64F-5* has 64 single-ended analogue inputs to 12 bits, eight TTL digital i/o lines, three counter timers and 16-bit DMA, all at 200kHz. *AT-MIO-16X* has similar facilities, with a 16-bit sampling ADC and AT-MIO-16D a 12-bit ADC. National also has the Lab-PC+, an improved version of the Lab-PC providing differential inputs a no increase in cost. National Instruments UK Ltd, 0800 289877.

Card memories. Triangle point out that the demand for card memories for notebooks has greatly reduced the price and that they are therefore worth considering for industrial use. The company offers the *TDS 20200*-based computer module and *TMB-200-03* or *TMB-55-03* card memory readers, which attach to a PC and appear as another disk drive to the computer, reading data in the sameway as from a dos file. Triangle Digital Services Ltd, 081-539 0285.

Development and evaluation

PIC16CXX emulator. Running on PC 286 upwards, Microchip's *Picmaster* 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 pins in DIP, PLCC, LCC, QFP, PGA, SOIC and TSOP packages. Attention has been paid to the resuction of ground bounce, so that the units cope with fast cmos PLDs and FPGAs. Direct Insight Ltd, 0455 558854.

Image compression. The JPEG image-compression board for PCs by C-Cube Microsystems is now available in Europe. It is an ISA half-card running at 10MHz and supporting grey scale, YUV (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 386SX. The still-image development kit includes schematics, PAL equations, .BMP-compatible link library source and full documentation. Kudos Thame Ltd, 0734 351010.

Eprom emulator. *Optorom* from Raisonance will emulate 8-bit eproms from the 27(C)16 to the 27(C)080 with no hardware upgrading, and handles eproms in 16 and 32-bit mode up to 4 by 4Mbit using an add-on board. Data can be down-loaded from from the serial port of a PC at 115Kbaud in binary, Intel-hex, 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, Forth, 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.

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 *ICS1700* 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.

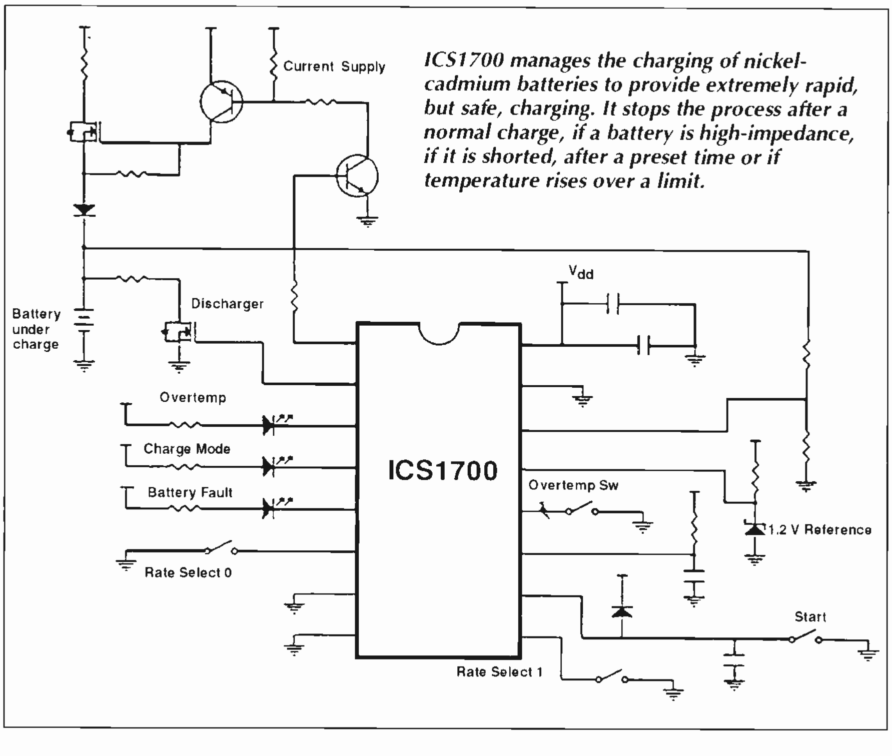
When fully charged, rested batteries that have lost surface charge are put under charge and there is a rapid initial rise in voltage, which is detected, and charging stopped. If they have not rested long enough to lose the surface charge, they can produce a negative slope initially, which again is detected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



ICS1700 manages the charging of nickel-cadmium batteries to provide extremely rapid, but safe, charging. It stops the process after a normal charge, if a battery is high-impedance, if it is shorted, after a preset time or if temperature rises over a limit.

***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, whoseoutput 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 newcycle 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 theinputs 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 *AD9501*s 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 $f=1/2t_{PD} + t_{P1} + t_{D2}$), $t_{D1,2}$ being the two programmed delays and t_{PD} the minmum propagation delay.

The random pulse generator of **Fig. 5** again uses two AD9501s, triggered together

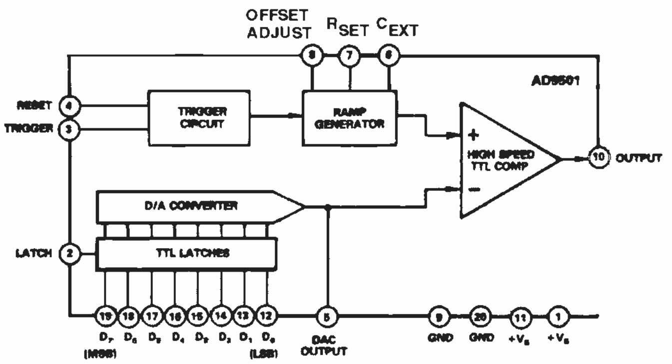


Fig. 1

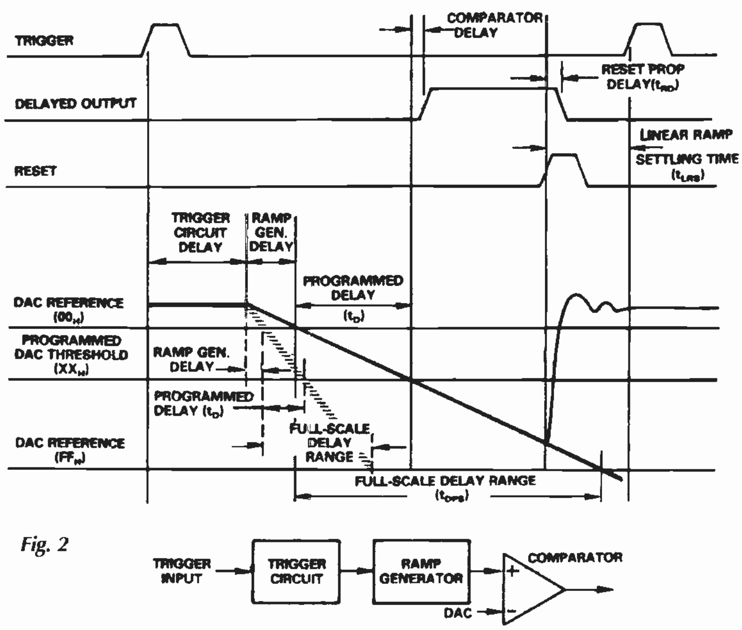


Fig. 2

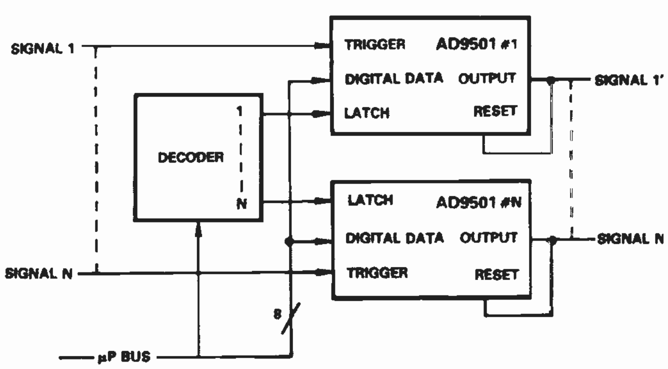


Fig. 3

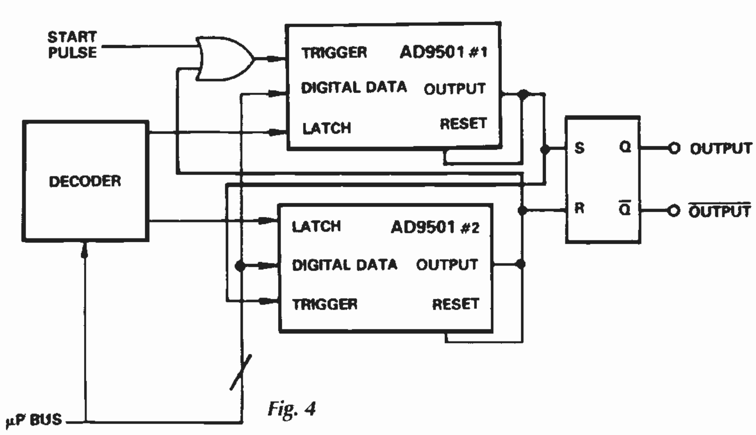


Fig. 4

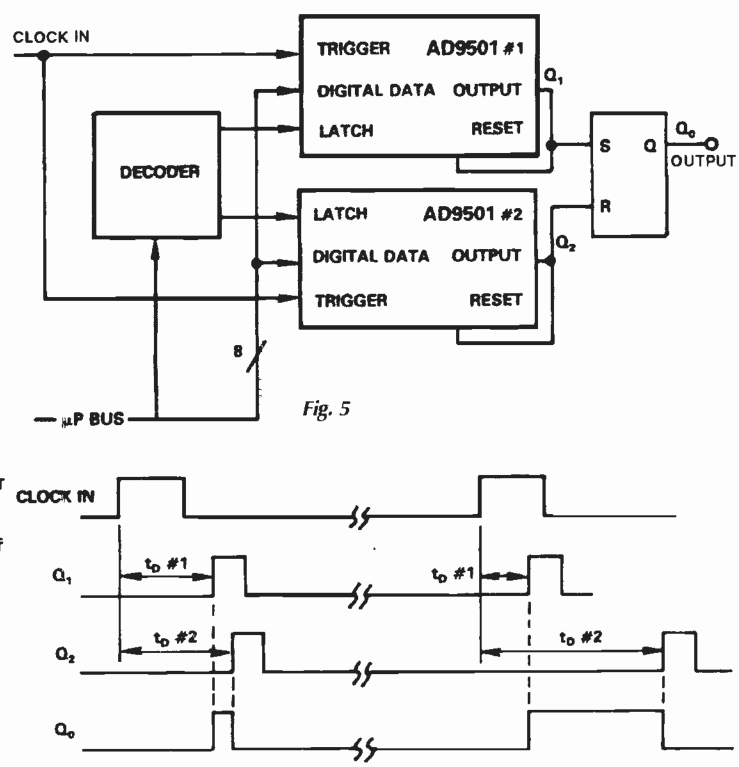
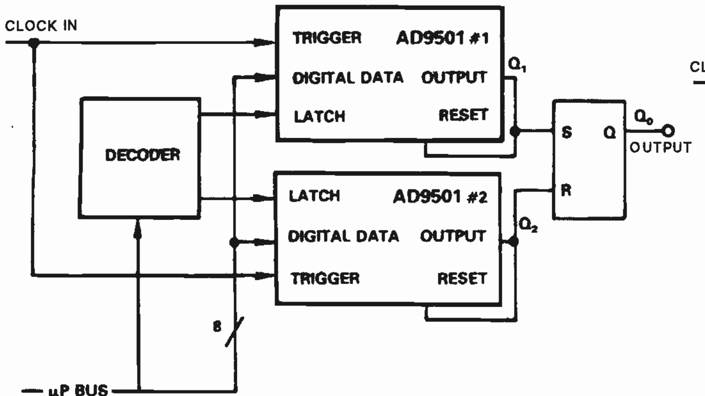


Fig. 5



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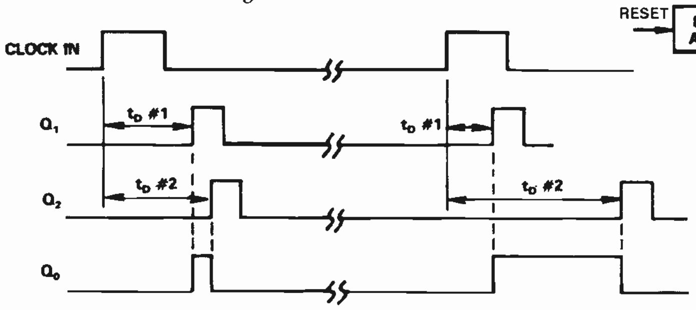


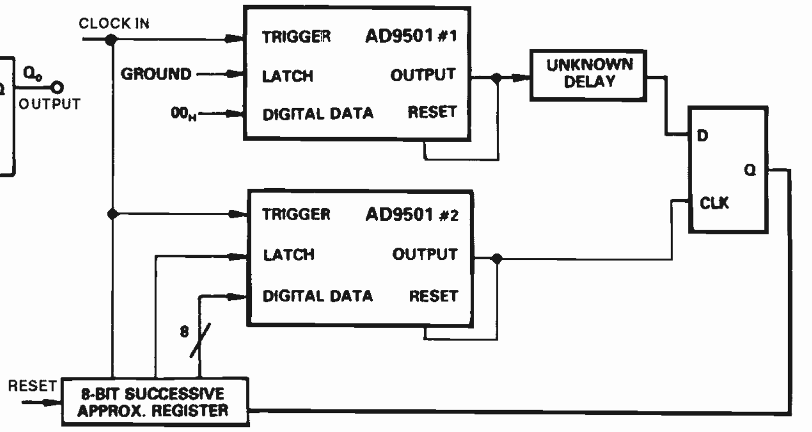


Fig. 6

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 theA-to-D's comparator. In calibration, short out the uknown 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 reinstsated. All this compensates for stray delays and setup times.

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

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 enouth in manyapplications.

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_1 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 $2\tau CR_1C_1$ (reducing noise bandwidth by, say, 100 reduces noise by10). The R_2D_2 arrangement maintains stability and R_2D_2 should equal $2R_1C_1$ 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_1 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_2 is almost nothing, giving rise to negligible leakage current. When driving large capactiive loads, $(C_{LOAD}+C_1)R_1$ must be less than $0.5R_2C_2$

■

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

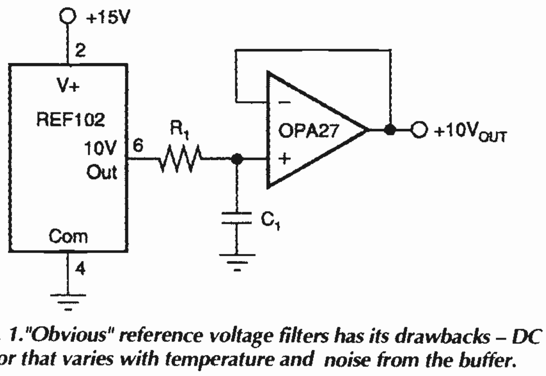


Fig. 1."Obvious" reference voltage filters has its drawbacks – DC error that varies with temperature and noise from the buffer.

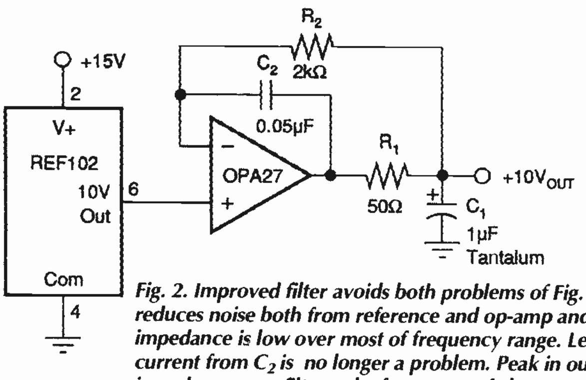


Fig. 2. Improved filter avoids both problems of Fig. 1. Filter reduces noise both from reference and op-amp and output impedance is low over most of frequency range. Leakage current from C_2 is no longer a problem. Peak in output impedance near filter pole frequency of about 35 is reduced by reducing R_1 and increasing C_1 – peak is $0.7R_1$.

INSTRUMENTS TO BUY

FREQUENCY COUNTERS

MX1010F and **MX1100F** are 8-digit frequency counters offering a broad range of features.

MX1010F: 1Hz to 100MHz, sensitivity of 15mV and resolution to 0.1Hz, data auto set, 10:1 attenuator, high impedance input – £129.00 plus VAT (£151.58).

MX1100F: 1Hz to 1GHz, features as MX1010F except ranges 70MHz to 1GHz and 50Ω impedance. £160.00 plus VAT (£188.00).

SC-130 and SC-40 are full featured, microprocessor-based, hand held frequency counters providing portability and high performance. Both instruments provide measurement of frequency, period, count and RPM plus a view facility enabling min, max, av and difference readings.

SC-130: 5Hz to 1.3GHz, 8 digit readout, sensitivity typically 10mV, high impedance input, battery condition indicator. £109.00 plus VAT (£128.08).

SC-40: As SC-130 except 5Hz to 100MHz. £89.00 plus VAT (£104.58).

MULTIMETERS

The **180 series** of high performance multimeters provide advanced features and are supplied complete with probes, battery and rubber holster. The case is dust and splash proof making it ideal in most environments. Designed to meet IEC348 Class II safety standard.

183: $3^1/_2$ digit large LCD display, ACV, DCV, ACA, DCA, resistance, continuity buzzer, diode test, hold, basic accuracy 0.5%. £39.50 plus VAT (£46.41).

185: $3^1/_2$ digit LCD, bar graph, ACV, DCV, ACA, DCA, resistance, continuity buzzer, diode test, hold, temperature (–40°C to 1370°C), capacitance (1pF to 40μF), frequency (1Hz to 200kHz), max min, edit, %, compare, basic accuracy 0.3%. £74.50 plus VAT (£87.54).

187: As 185 except auto ranging. £99.50 plus VAT (£116.91).

285: As 185 except $4^1/_2$ digit true rms, basic accuracy 0.05%. £109.50 plus VAT (£128.66).

MULTIMETERS (2)

The **MX170B** and **MIC-6E** offer low cost measurement yet retain a large number of features. Supplied complete with probes.

MX170B: $3^1/_2$ digit LCD, compact size, ACV, DCV, DCA, resistance, diode test, low voltage battery test. £24.00 plus VAT (£28.20).

MIC-6E: $3^1/_2$ digit LCD, ACV, DCV, ACA, DCA, resistance, diode test, buzzer.
£33.50 plus VAT (£39.36).

20MHz 2-CH OSCILLOSCOPE

The CS4025 20MHz dual trace oscilloscope offers a comprehensive range of facilities including a high sensitivity vertical amplifier providing from 1mV to 5V/div in CH1, ALT, CHOP, ADD, CH2 modes with inverse polarity on CH2. The horizontal timebase offers a sweep range of 0.5s/div to 0.5μs/div plus ×10 sweep expansion and X-Y mode. Triggering can be auto or normal from vert, CH1, CH2, line or external sources with coupling provided for AC, TV-F and TV-L. The CS4025 is supplied complete with matching probes for £295.00 plus VAT (£346.62).

PROGRAMMABLE POWER SUPPLIES

The PPS series of GPIB programmable DC power supplies offer high performance yet are extremely competitively priced using a 16 x 2 backlit LCD and 14 button keypad. All functions and conditions are easily selected and displayed. Overvoltage and overcurrent are selectable as is output enable/disable. Terminals for output and sense are provided on the front and rear to allow easy rack mounting.

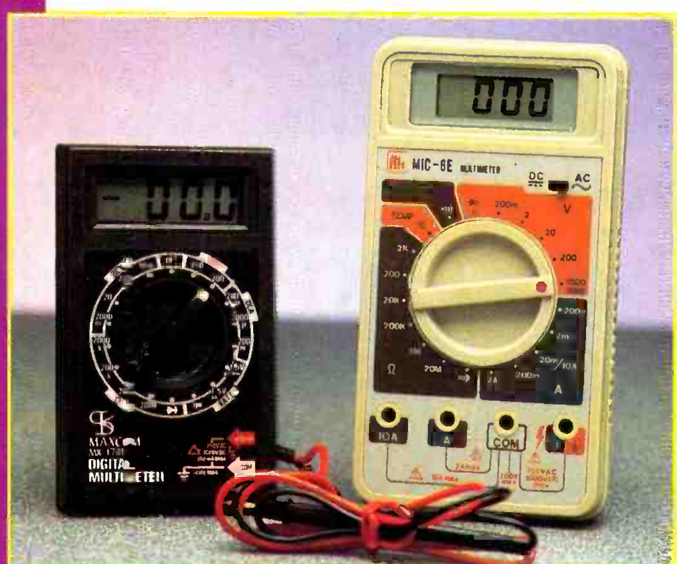
PPS-1322: 0-32V 2A (GPIB) £375.00 plus VAT (£440.63)

PPS-2322: Dual 0-32V 2A (GPIB) £555.00 plus VAT (£652.13)

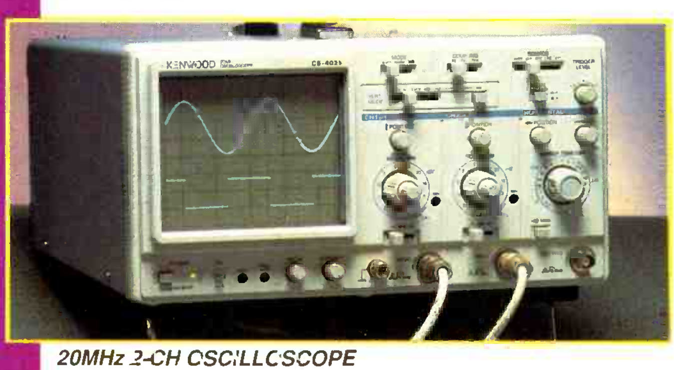
MX1010F MX1100F

FG506 FG513

THE 180 SERIES



MX170B MIC-6E



20MHz 2-CH OSCILLOSCOPE

PROGRAMMABLE POWER SUPPLIES

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FUNCTION GENERATOR

The **MX2020** 0.02Hz – 2MHz sweep function generator with LED digital display offers a broad range of features. Output waveforms include sine, square, triangle, skewed sine, pulse and TTL. Lin and log sweeps are standard as is symmetry, DC offset and switchable output impedance from 50Ω to 600Ω. The digital display provides readout of the generators' frequency or can operate as separate 10MHz frequency counter. £199.00 plus VAT (£233.83).

LCR METER

The **MIC-4070D** LCD digital LCR meter provides capacitance, inductance, resistance and dissipation measurement. Capacitance ranges are from 0.1pF to 20,000μF plus dissipation. Inductance ranges from 0.1μH to 200H plus a digital readout of dissipation. Resistance ranges from 1mΩ to 20MΩ. Housed in a rugged ABS case with integral stand it is supplied complete with battery and probes at £85.00 plus VAT (£99.88).

FOUR INSTRUMENTS IN ONE

The **MX9000** combines four instruments to suit a broad range of applications in both education and industrial markets including development work stations where space is at a premium.

The instruments include:

1. A triple output power supply with LCD display offering 0-50V 0.5A, 15V 1A, 5V 2A with full overcurrent protection;
2. An 8-digit LED display 1Hz - 100MHz frequency counter with gating rates of 0.1Hz, 1Hz, 10Hz and 100Hz providing resolution to 0.1Hz plus attenuation inputs and data hold;
3. A 0.02Hz to 2MHz full featured sweep/function generator producing sine, square, triangle, skewed sine, pulse and a TTL output and linear or logarithmic sweep. Outputs of 50Ω and 600Ω impedance are standard features;
4. An auto/manual $3^1/_2$ digit LCD multimeter reading DCV, DCA, ACV, ACA, resistance, and relative measurement with data hold functions.

The MX9000 represents exceptionally good value at only £399.00 plus VAT (£468.83).

FG SERIES FUNCTION GENERATORS

The **FG500 series** sweep/function generators provide two powerful instruments in one package, a 6MHz or 13MHz sweep/function generator and an intelligent 100MHz frequency counter. The microprocessor based instruments offer sophisticated facilities yet remain extremely competitively priced. A menu driven display allows easy set up and operation. A 16 character by 2-line LCD display provides clear and unambiguous readout of generator output and frequency measurement.

FG-506: 2Hz to 6MHz sweep/function with 100MHz counter £325.00 plus VAT (£381.88)

FG-513: 2Hz to 13MHz sweep/function with 100MHz counter £482.00 plus VAT (£566.35)

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EWWO/93

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 ($R_d = Q\omega L$, assuming the capacitor is perfect) replaced by an equivalent series loss component r associated with the coil ($r = \omega L/Q$). From

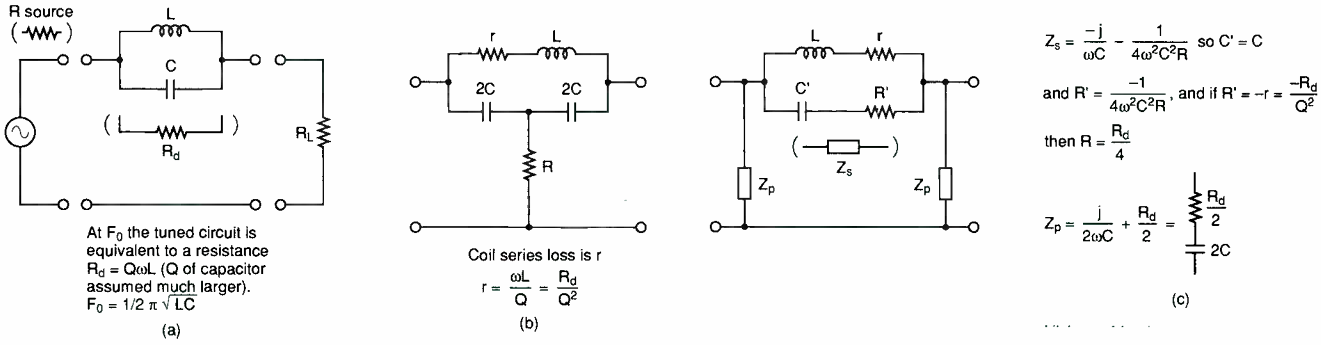


Fig. 1a). A parallel tuned circuit used as a rejector. The notch depth is set by the ratio of the tuned circuit's dynamic resistance R_d and the load resistance R_L. At F_0 the tuner circuit is equivalent to a resistance $R_d = Q\omega L$ (Q of capacitor assumed much larger). $F_0 = 1/2\pi\sqrt{LC}$. b) The circuit modified to provide a deep notch, tuned frequency unchanged. Coil series losses $r = \omega L/Q = R_d/Q^2$ c) As b), but with the star network transformed to the equivalent delta network. $Z_S = (-j/\omega C)-1/(4\omega^2C^2R)$. So $C'=C$ and $R' = -1/4\omega^2C^2R$, and if $R'= -r = -Rd/Q^2$ then $R = R_d/4$. $Z_p=(j/2\omega C)+(R_d/2)$

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 formulae. The result is as in **Fig. 1c** and the circuit can now provide a deep notch even if R_L is infinite, owing to the presence of the shunt impedance Z_p 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 $-R_d$. Now R' turns out to be $-1/(4\omega C R)$ and equating this to $-r$ gives $R = R_d/4$.

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 $B = \omega\sqrt{(LC)}$ 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 1Ω 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 jacked 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

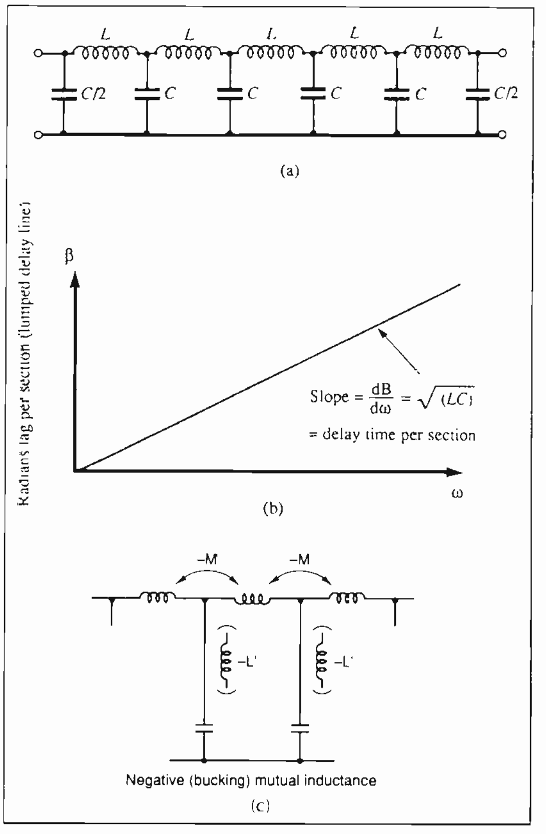


Fig. 2a). Basic delay line – b) providing a delay of √(LC) seconds per section at DC and low frequencies. c) Connection of negative inductance in the shunt arms to linearise the group delay over a larger proportion of the filter's passband. Not a physical component, it is implemented by negative mutual inductance (bucking coupling) between sections of series inductance.

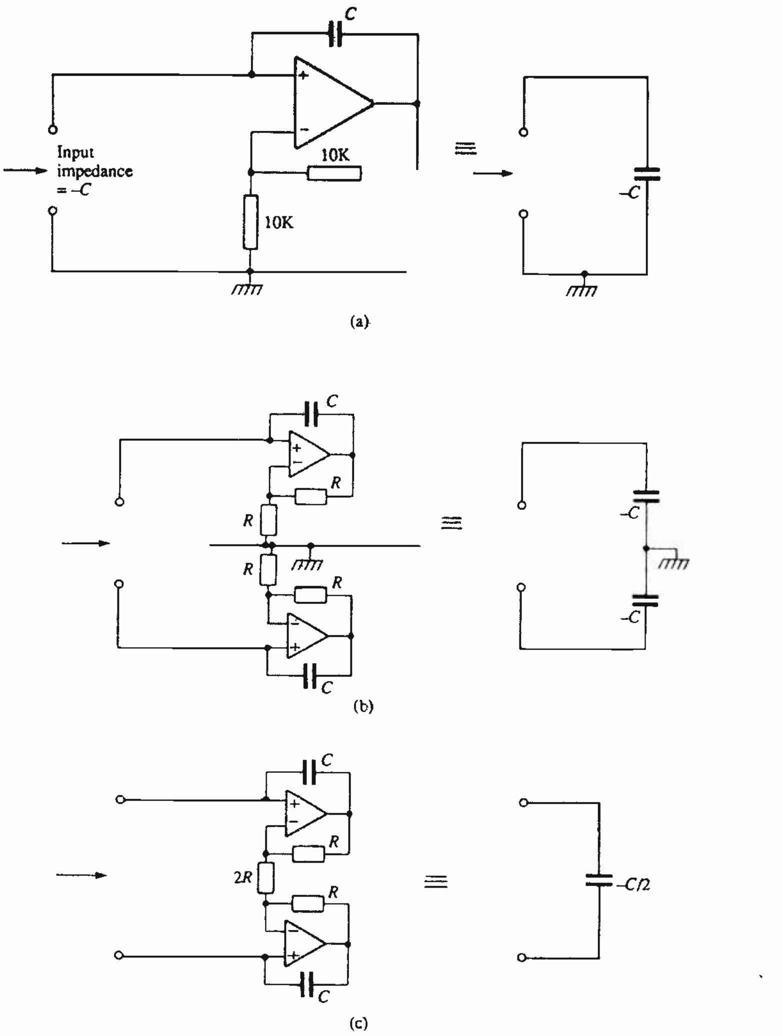
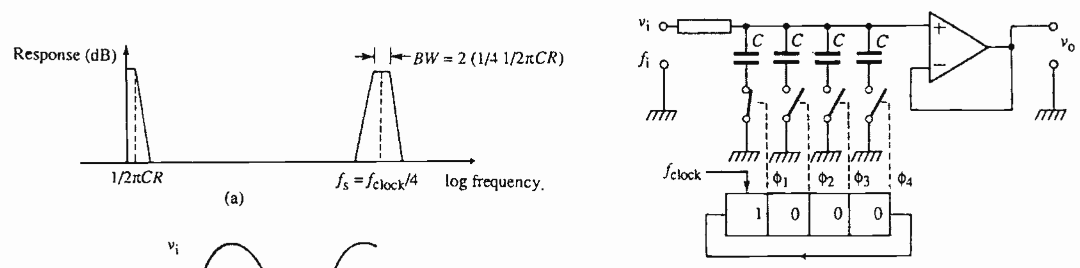
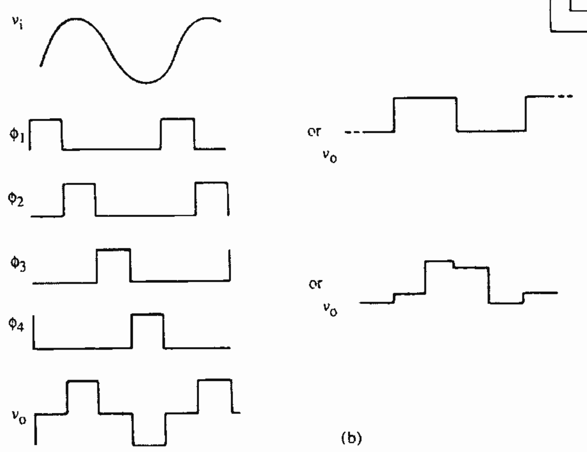


Fig. 3a) Unbalanced negative capacitor (one end grounded). b). Balanced, centre grounded negative capacitor. c). Floating negative capacitor.





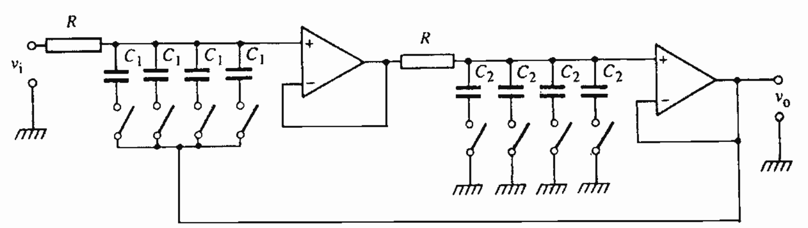
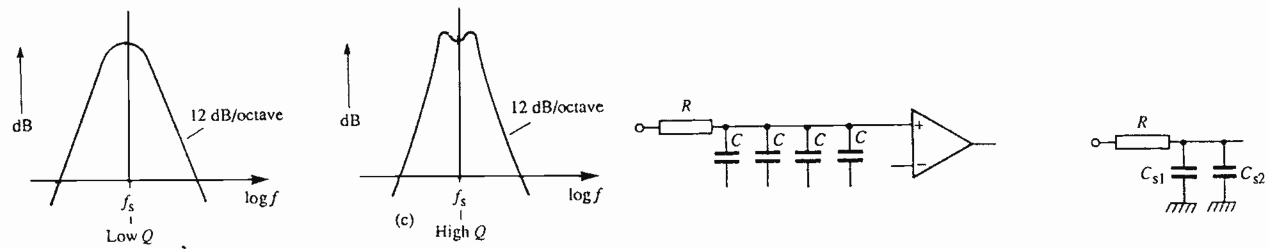


Fig. 4a) One-pole low-pass equivalent (LPE) N-path band-pass filter section. A solitary 1 circulating in a shift register is only one of the many ways of producing the four-phase drive waveform shown in b).
b). Waveforms associated with a). The exact shape of v_o, when $f_1 = fc_{lock}/4$ exactly will depend on the relative phasing of v_i and the clock waveform. For very small difference between f_1 and $f_{clock}/4$, the output waveform will continuously cycle between the forms shown, and all intermediate shapes.
c). Second order N-path filter, showing circuit frequency response. $Q=1/2\sqrt{(C_1/C_2)}$, exactly as for the low-pass case.
d). Stray capacitance. Showing the stray capacitance to ground, consisting of op-amp input capacitance C_{s2} plus circuit and component capacitance to ground with all switches open at C_{s1}.



the current always of the wrong sign, Ohm's law will still hold if the value of the resistor is labelled as -1Ω. **Figure 3** a shows the scheme, this time put to use to provide a capacitance of $-C\mu$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 4** a illustrates the basic scheme, using a first-order section.

Apply at V_i 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 $4CR$, 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)

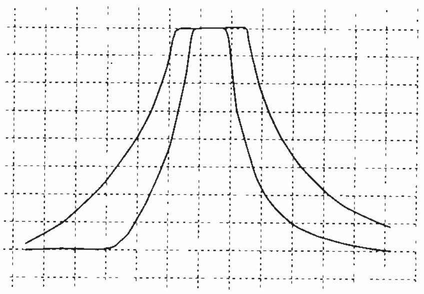


Fig 5 a

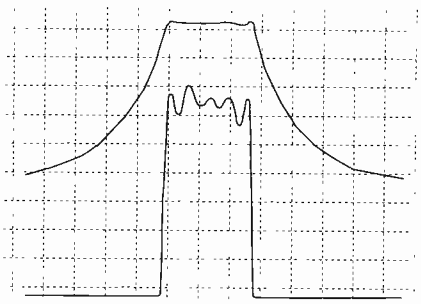


Fig 5 b

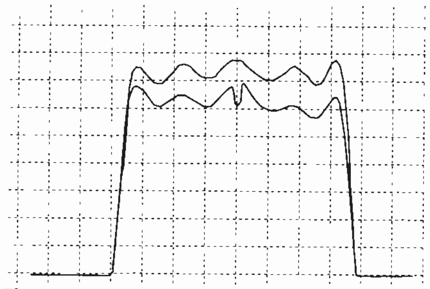


Fig 5 c

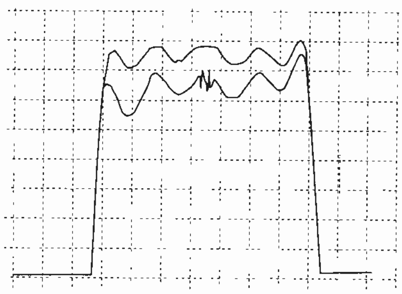


Fig 5 d

Fig. 5a). 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 and 100Hz/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; 20Hz/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.

by an amount less than $1/(\pi 4CR)$, 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_{clock} 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 – low pass equivalent) section, **Figs 4c** and **4d**. **Figure 5** a 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 this 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_1 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 of Fig. 5b, correcting the slope. **Figure 5d** shows the corrected passband (upper trace) and the effect of increasing the negative capacitance to –100pF (lower trace), resulting in overcompensation. These, and other examples which could be cited, show the usefulness of negative components to the professional circuit designer. While they may not be called for every day, they should certainly be regarded as a standard part of an armoury of useful techniques.

Acknowledgments

Figures 2a, 2b, 3 and 4 reproduced with permission from Analog Electronics, Ian Hickman, Heinemann Newnes, 1990 ISBN 0 434 907 3 5

Optoisolator correction

In the Design Brief "Bringing the optoisolator into line" (EW + WW December 1992 pp. 1050 - 105), 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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Dataman's new S4 programmer costs £495
You could have one tomorrow on approval*

If you've been waiting for S4 we have some good news. It's available now. S4 is the 1992 successor to Dataman's S3 programmer, which was launched in 1987. The range goes back through S2, in 1982, to the original Softy created in 1978. Like its predecessors, Softy4 is a practical and versatile tool with emulation and product development features. S4 is portable, powerful and self-contained. Design and manufacture are State of the Art. S4 holds a huge library of EPROMS, EEPROMS, FLASH and One Time Programmables. Software upgrades to the Library are free for the life of the product, and may be installed from a PROM by pressing a key. *S4 makes other programmers seem oversized, slow and outdated.* S4 is now the preferred tool for engineers working on microsystem development.

Battery Powered

S4 has a rechargeable NICAD battery. On average, you can do a week's work without recharging. On a single charge, up to a thousand PROMS can be programmed – and charging is fast: it only takes an hour. Normal operation can continue during the charging process.

Continuous Memory

Continuous Memory means never losing your Data, Configuration or Device Library. You can pick up S4 and carry on where you left off, even after a year on the shelf. If the NICAD battery loses all of its charge, RAM contents are preserved by the LITHIUM backup battery.

Remote Control

S4 can be operated via it's RS232 Serial Port. The standard D25 socket connects to your computer. Using batch files or a terminal program, all functions are available from your PC keyboard and screen.

Free Terminal Program

You could use any communications software to talk to S4. But the Terminal Driver program, which we include free, is the best choice. It has Help Screens to explain S4's functions and it sends and receives at up to 115200 baud – that's twelve times as fast as 9600 baud. At this speed a 64 kilobyte file downloads in 9 seconds. There is a memory resident (TSR) option too, which uses only 6k of your precious memory, and lets you "hot key" a file to S4. Standard *upload* and *download* formats include: ASCII, BINARY, INTELHEX, MOTOROLA and TEKHEX.

S4 loads its Library of programmables from a PROM in its socket, like a computer loads data from disk. Software upgrades are available free. Download the latest Device Library from our Bulletin Board.

Microsystem Development

With S4 you can develop and debug microsystems using Memory Emulation. This is an extension of ROM emulation, used for prototype development, especially useful for single-chip "piggy back" micros. When you unpack your S4 you will find an Emulator Lead with a 24/28/32 pin DIL plug and a Write Lead with a microhook. Plug the EMULead in place of your ROM. Hook the Write-Lead to your microprocessor's write-line. Download your assembled code into S4. Press the EMULate key and your prototype runs the program. S4 can look like ROM or RAM, up to 512K bytes, to your target system. Access-time depends on S4's RAM. We are currently shipping 85ns parts.

CIRCLE NO. 101 ON REPLY CARD

Your microprocessor can *write* to S4 as well as *read*. If you put your *variables* and *stack* in S4's memory space, you can inspect and edit them. You can write a short monitor program to show your *internal registers*.
S4's memory emulation is an inexpensive alternative to a full MDS and it works with any microprocessor. Many engineers prefer it because their prototype runs the same code that their product will run in the real world.

Dimensions & Options

S4 measures 18 x 11 x 4 cm and weighs 520 grams. 128k x 8 (1MB) of user memory is standard, but upgrading to 512k x 8 is as easy as plugging in a 4MB low-power static CMOS RAM.
The stated price includes Charger, EMUlead, Write Lead, Library ROM, Terminal Driver Software with Utilities and carriage in U.K. but not VAT.

**Money-back Guarantee*

We want you to buy an S4 and use it for up to 30 days. If it doesn't meet with your complete approval you will get your money back, immediately, no questions asked.

Call us with your credit card details. Stock permitting, we are willing send goods on 30 days sale-or-return to established U.K. companies on sight of a legitimate order.

Customer Support

Dataman's customer list reads like Who's Who In Electronics. Dataman provides support, information interchange, utilities and latest software for S4, S3, Omni-Pro and SDE Editor-Assembler on our Bulletin Board which can be reached at any time, day or night.

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