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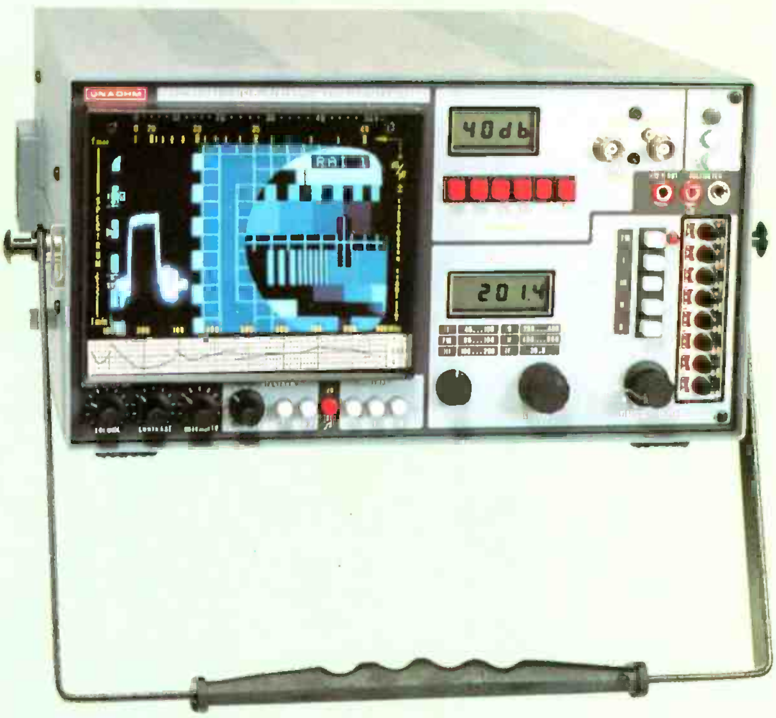

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Measurement Indication:	ANALOGUE: brightness stripe against calibrated scale superimposed on picture tube. The stripe length is proportional to the sync peak of the video signal.	
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Video Frequency Input:	Minimum Voltage: 1Vpp. Impedance: 75Ω or 10KΩ in case of a through-signal. Connector type: BNC
Teletext Input:	Voltage: 1Vpp/75Ω.
Teletext Clock Input:	Voltage: 1Vpp/75Ω. Measurement: Aperture of eye pattern: linear or Lissajous figures, selectable. Indication: directly on the picture tube. A calibrated scale shows percentage of eye pattern aperture. Error: the instrument introduces an error of less than or equal to 5% with video input and 20% with RF input. Jitter on regen'd clock: less than or equal to 25ns. Line selector: Selection of any TV line between the 2nd and the 625th scanning cycle by means of a 3 digit thumbwheel switch.
Oscilloscope:	VERTICAL CHANNEL: Sensitivity: 0.5 to 2Vpp/cm. Frequency Response: DC to 10MHz. Rise time: pre & overshoot less than or equal to 2%. Input Coupling: AC. Input Impedance: 75Ω/50pF. TIME BASE: Sweep Range: 20 to 10ms (1.1/2 frames); 32; 64/192us (1/2: 1: 3 lines). Linearity: +/-3%. Horizontal Width: 10 divisions; x5 magnification.
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AUGUST 1989 VOLUME 95 NUMBER 1642

REGULARS

Sound and vibration research is the subject of Research Profile, page 768.

Satellite tv received on a simple, efficient and cheap flat-plate antenna – page 784.

Pioneer Venus Orbiter was designed to let us see under the planet's could cover – Research Notes, page 749.

next issue

In next month's issue. PC based systems. The September issue of *Electronics & Wireless World* details how the IBM PC has escaped into the world of instrumentation and process control. As the machines now cost less than the most basic industrial rack computers, they threaten to take over the industrial and scientific scene.

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REED BUSINESS PUBLISHING

Electronics & Wireless World is published monthly. USPS687540. By post, current issue £2.25, back issues (if available) £2.50. Order and payments to 301 *Electronics and Wireless World*, Quadrant House, The Quadrant, Sutton, Surrey SM2 5AS. Cheques should be payable to Reed Business Publishing Ltd. **Editorial & Advertising offices:** *EWW*, Quadrant House, The Quadrant, Sutton, Surrey SM2 5AS. **Telephones:** Editorial 01-661 3614. Advertising 01-661 3130. 01-661 8469. **Telex:** 892084 REED BP G (EEP). **Facsimile:** 01-661 8939 (Groups II & III). **Beeline:** 01-661 8978 or 01-661 8986. 300 baud, 7 data bits, even parity, one stop bit. Send ctrl-Q, then EWW to start, NNNN to sign off. **Newstrade:** Quadrant Publishing Services No. 01-661 3240. **Subscription rates:** 1 year (normal rate) £30 UK and £35 outside UK. **Subscriptions:** Quadrant Subscription Services, Oakfield House, Perrymount Road, Haywards Heath, Sussex RH16 3DH. Telephone 0444 441212. Please notify a change of address. **USA:** $116.00 airmail. Reed Business Publishing (USA), Subscriptions Office, 205 E. 42nd Street, NY 10117. **Overseas advertising agents: France and Belgium:** Pierre Mussard, 18-20 Place de la Madeleine, Paris 75008. **United States of America:** Jay Feinman, Reed Business Publishing Ltd, 205 East 42nd Street, New York, NY 10017. Telephone (212) 867 2080. Telex 23827. **USA mailing agents:** Mercury Airfreight International Ltd Inc, 10(b) Englehard Ave, Avenel NJ 07001. 2nd class postage paid at Rahway NJ. Postmaster – send address to the above.

Making sense of the world

Most electronics developments are evolutionary and have little overall significance on their own. But there are two areas of emerging technology which will affect our quality of life profoundly.

Neural networks and digital signal processing either separately or taken together promise far greater change than was brought about by the coming of the microprocessor. The micro simply added more functions to essentially familiar objects, creating digital petrol pumps, pocket calculators, highly intelligent washing machines and the like. But the newer technology promises machines which speak and understand my language or yours, that can synthesize a worldly way of thinking, that can directly influence the world about us.

Lotus makes fine sports cars. They are fast, but suffer from a high internal noise level due largely to the use of lightweight composite body shells. Certain combinations of road surface, speed and engine revs can create unpleasant resonances which affect driver and passenger comfort. Lotus called in TI which co-operated with the car firm to design an adaptive sound feedback system based on the semiconductor company's best selling 32020 DSP chip. The system listens to the vehicle noise and then works out the sort of row it should create inside the car to cancel the unwanted row from outside and elsewhere. It kills the noise dead. It simply requires four strategically placed microphones, $50 of silicon arranged as a frequency and time adaptive filter, with the filter's audio output played through the car's own audio system.

GCHQ routinely monitors cable and satellite overseas telephone traffic using facilities at Cheltenham and Morwenstow on the north coast of Cornwall. It has automated the process. Banks of digital signal processors, programmed to respond to keywords of the sort expected in "sensitive" conversations, listen for their cue to turn on a tape recorder. The processors respond effectively to a variety of languages and accents. They are forever vigilant and never get bored.

Masses of DSP applications between these two extremes are already in existence. The world of test and measurement uses the technology as routinely as the Government. It has transformed spectrum analyser design to the point where it is not possible to conceive of a design without one. DSP can be used to recognize a voice, a thumb-print or a signature on a piece of paper. It simply demands the right signal and the right programming.

Electronic neural networks make use of massively parallel processors which link together according to the needs of, and stimuli from the outside world. This technology and the closely-related science of artificial intelligence have only just started down the development path and there are few useful things which they can do as yet. But that will change, everywhere.

Mimicking our own thought processes, their essence is to be able to do the sort of things which conventional computers can now do, but without the unnatural language of conventional computer programming. Machines are already in existence which will accept dictation. Robotics is looking to AI for a generation of obedient, amiable and tireless automata to staff our factories and offices. Before AI the willing robot was a pipedream. It is now becoming a reality.

It seems arguable that DSP and AI will make the world a better place. This new handful of sand will certainly make it a different one.

MHD – new hopes for clean power?

Magnetohydrodynamics – mercifully shortened to MHD – could well emerge as the Cinderella of the energy scene, at least if there's a positive outcome to a new research project being undertaken at BNF Metals Technology of Wantage in Oxfordshire. MHD generates electricity by the direct interaction of fast-flowing ionized gases with a magnetic field, thus bypassing the need for steam boilers and turbines. Also, because of the high temperature at which MHD generators operate, thermodynamic efficiency can be much greater.

What happens essentially is that coal is burnt in the usual way, except that the temperature is a lot higher and that a seed material (usually potassium carbonate) is added to ionize the gases. When the resulting plasma flows between the poles of magnet, it generates electricity in much the same way as any other generator. The difference is that whilst an ordinary generator relies on rotating copper wires, an MHD generator uses gases at sonic velocities.

The generator itself can come in a variety of different forms, but at its simplest is a channel in which the plasma encounters the all-important magnetic field. With a field of around 6 tesla and a gas velocity of around 600m/s, output power density from collector electrodes can be as high as 24MW per cubic metre of reactor space.

The benefits of an MHD generator are clearly considerable in theory: no moving mechanical parts, simple design, extremely compact and relatively efficient. The only reasons it hasn't replaced existing steam turbines are the formidable problems of designing a practical reactor. Experiments carried out twenty years ago by the CEGB ran into problems because of repeated component failure. Elsewhere difficulties have arisen from corrosion or from problems of operating at the necessarily high temperatures of around 2500°C. Yet in spite of those difficulties, a number of reasonably successful pilot plants have operated in Japan, Australia, the USSR, USA and Italy.

In all these systems, however, the magnets have placed a fundamental limit on efficiency. To create a field as high as 6T they've all used electromagnets which inevitably consume a significant proportion of the electricity that's generated. What BNF Metals Technology is now doing is researching the design of permanent magnets that would eliminate this problem.

It if succeeds then MHD generators are likely to have a practical future, not as stand-alone units, but as bolt-on devices to conventional power stations. The main reason for this is that the exhaust gases of an MHD generator are still hot enough to boil water for driving a conventional steam turbine.

But there's a second reason why it's considered a good idea to add MHD to conventional power stations: the elimination of pollution. The seeding process in which chemicals are added to make hot gases conduct electricity also necessitates a recovery system for those chemicals. Such a recovery system would simultaneously remove sulphurous materials from the coal and thus help to eliminate acid rain.

All that seems necessary to get MHD off the ground is – as BNF has rightly identified – a detailed look at some of the basic ingredients like permanent magnets. . . and of course some fairly hefty DC to AC inverters!

Lossless wave propagation

Photon torpedoes may not forever remain the preserve of Captain Kirk *et al.* aboard the Starship *Enterprise*. The notion of bundles of energy travelling through space without dissipation has recently taken a step towards reality as a result of work being undertaken at Houston University and at the Lawrence Livermore National Laboratory. Richard Ziolkowsky and his co-workers report (*Phys. Rev. Lett*, 9.1.89) that they have created ultrasonic pulses that can travel several metres in water without dispersing.

These acoustic energy-directed pulse trains (Adepts), which are thought likely to have their counterparts in electromagnetic radiation, are yet another hint that wave behaviour isn't only limited to what can be deduced from the laws of diffraction, refraction and reflection. Yet it is from these laws that the existence of Adepts was first hinted at. Ziolkowsky says that mathematical expressions describing the propagation of waves do actually allow for non-dispersive transmission of energy.

His latest work, supported interestingly enough by the Strategic Defence Initiative, uses an array of underwater piezo-electric transducers designed to produce a pencil-thin beam of radiated energy. But unlike even the finest laser beams, Adepts do not spread outwards at all, at least not within the range so far achieved. Thereafter they rapidly lose coherence and dissipate.

Although a few metres in water may seem undramatic, the eventual applications of transmitting energy without loss could be revolutionary.

. . .And for the record books

Journalists (*mea culpa*) can so easily get hooked on the numbers game. . . "95-year-old twice-married grandmother of 20 wins 26 mile marathon in 2hr 30min" and so forth. There are occasions, however, when it's worth at least taking note of some of the more significant milestones in modern electronics, especially for those of us who can remember marvelling at the prodigious speed of a 741 op-amp.

P.C. Chao *et al.* (*Electronics Letters* Vol. 25 No 8) describe the successful development of a high electron mobility transistor (HEMT) with a noise figure of 3dB and a gain of 5.1dB at 94 GHz. The team, from General Electric and Cornell University, claim that this device, with a gate width of 0.1μm, is the first ever low-noise transistor that would be practical for use in W-band receivers.

In the same issue of *Electronics Letters*, a West German team from Siemens and the Ruhr University at Bochum describe what they claim is the fastest silicon static divider IC capable of working on a standard 5V power rail. It's a four-stage 16:1 divider with an integral 50Ω output buffer stage and operates at up to 15GHz.

Faster (25GHz) dividers have been made, but they either need higher voltage supplies or require the use of III-V semiconductors or exotic fabrication technologies. Achieving 15 GHz with a (more-or-less) bog-standard silicon technology seems impressive enough to me!

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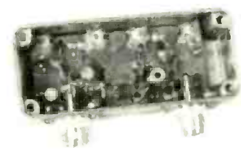
TYPE 9035. Mains power supply for above amplifiers £39
TYPE 9010. Masthead weatherproof unit for above amplifiers £12

TYPE 9006



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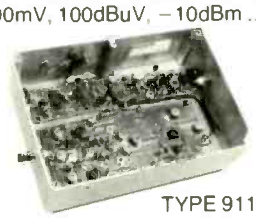
TYPE 9253. 40-860MHz. NF 6dB. Gain 30dB. Voltage output 100mV, 100dBuV, −10dBm £94



TYPE 9113

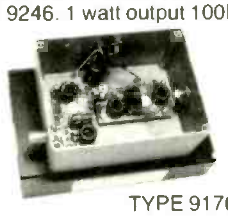
TYPE 9252

PHASE LOCKED LOOP FREQUENCY CONVERTERS

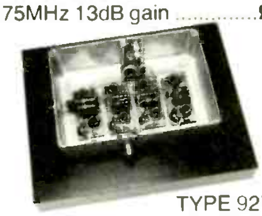
TYPE 9113 Transmitting. Converts your specified input channels in the range 20-1000MHz to your specified output channels in the range 20-1000MHz. 1mV input, 10mW output (+10dBm). AGC controlled. Gain 60dB adjustable −30dB. Will drive transmitting amplifiers directly £356
TYPE 9114 Receiving. Low noise Gasfet front-end. NF 0.7dB. Gain 25 dB variable £356

TMOS WIDEBAND LINEAR POWER AMPLIFIERS

TYPE 9246. 1 watt output 100KHz-175MHz 13dB gain £108



TYPE 9176



TYPE 9271

TYPE 9247. 4 watts output 1-50MHz 13dB gain £108
TYPE 9051. 4 watts output 20-200MHz 13dB gain £108
TYPE 9176. 4 watts output 1-50MHz 26dB gain £254
TYPE 9177. 4 watts output 20-200MHz 26dB gain £254
TYPE 9173. 20 watts output 1-50MHz 10dB gain £308
TYPE 9174. 20 watts output 20-200MHz 10dB gain £308
TYPE 9271. 40 watts output 1-50MHz 10dB gain £616
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TYPE 9235. Mains power supply unit for above amplifiers £164

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FM TRANSMITTERS 88-108MHz. 50 watts RF output

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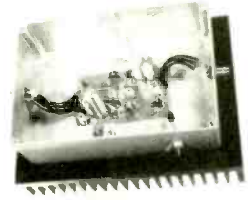
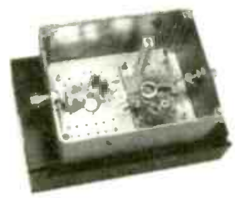
TYPE 9087. Includes integral mains power supply £1,110
TYPE 9182FM exciter ±75KHz deviation. Output 10mW £248



TYPE 9263



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TELEVISION LINEAR POWER AMPLIFIERS

Tuned to your specified channels in bands IV or V. 24V + DC supply.

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TYPE 9252. 10mW input, 500mW output £254
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TYPE 9263. 2-3 watts input, 15 watts output £400
TYPE 9266 10 watts input, 50 watts output £1,585

See below for Television Amplifiers in bands I & III.

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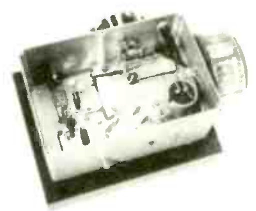

TYPE 9158/9235

TMOS RF LINEAR POWER AMPLIFIERS

Tuned to your specified frequency in the range 20-250MHz, or your specified channels in bands I or III. 24V +DC supply.

TYPE 9105. 10mW input, 1 watt output £230
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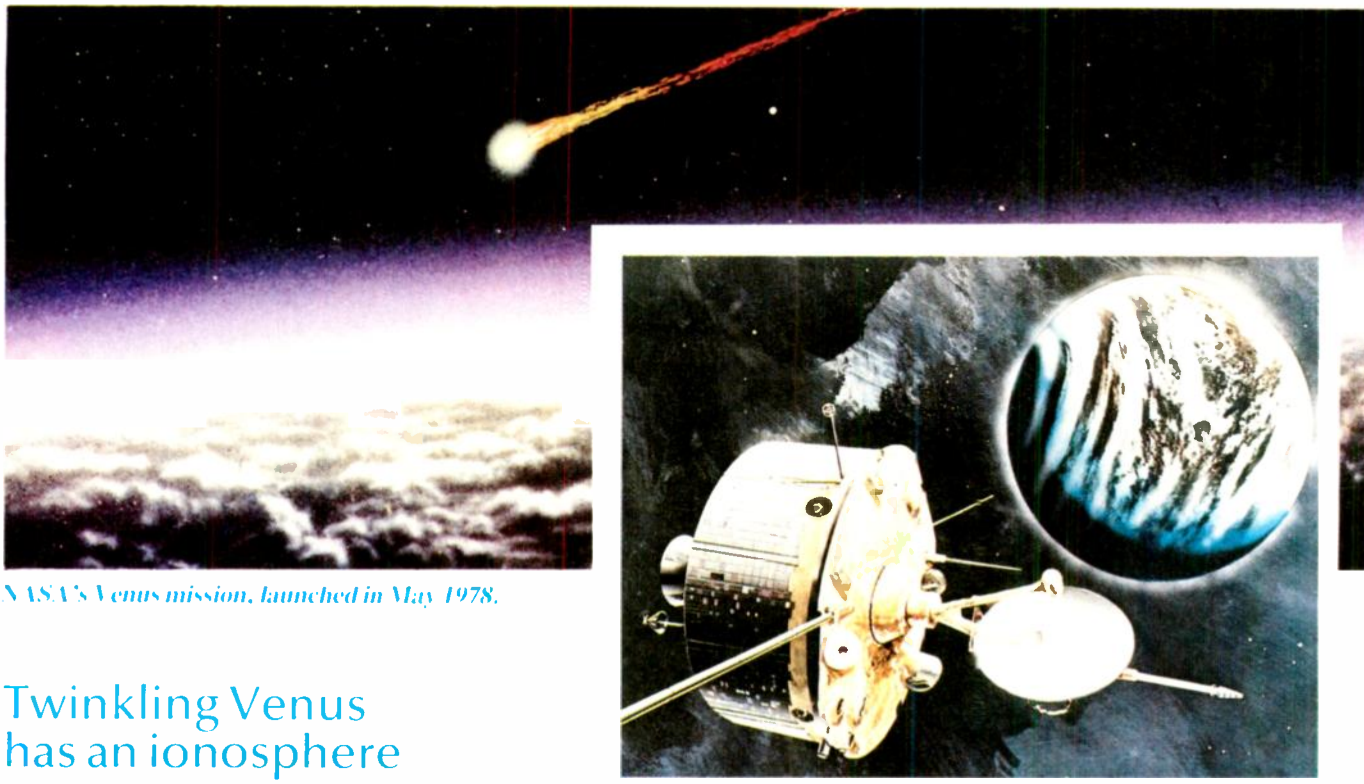

NASA's Venus mission, launched in May 1978.

Twinkling Venus has an ionosphere

Although most of us are aware that stars twinkle because of temperature-induced turbulence in the atmosphere, we're perhaps less familiar with the ionospheric irregularities that can cause twinkling at radio wavelengths. It happens essentially when variations in the solar wind interact with the Earth's magnetic field and it affects radio transmissions passing through the ionosphere.

It's not just our Earth, however, that has a twinkling ionosphere; the effect has been observed on signals from deep space probes such as Voyager-2 (Research Notes, July, page 648) as they've passed close to Jupiter, Saturn and Uranus. Each of these giant planets has a strong magnetic field which interacts with solar radiation to create its own ionosphere.

But until recently, no-one knew whether Venus, our nearest planet, had a magnetic field strong enough to support an ionosphere. Now, to judge from a re-analysis of data from the Pioneer Venus Orbiter spacecraft (*Science News* Vol. 135 No 8), Venus does indeed have an ionosphere. William L. Sjogren, Richard Woo *et al.* from the Jet Propulsion Laboratory, Pasadena, were mapping the Venusian gravitational field by measuring Doppler shifts on radio signals from the spacecraft when they observed irregularities coinciding with peaks in the solar wind. But the irregularities weren't always present, nor could they be attributed to instrument malfunction or direct interference from the Sun.

Eventually the team showed that the noise signals occurred only when the spacecraft was on the side of Venus facing the Sun and then only when it was orbiting low enough for its signals to pass through the Venusian ionosphere. The reason why the solar wind plays a part is that the magnetic field is too weak to hold it at bay and prevent it causing the variations in the ionospheric electron density that in turn are responsible for the 'twinkles'.

Another planet with a weak magnetic field that may have a twinkling ionosphere is Mars. Sadly, though, since the demise of the Russian Phobos probes we are unlikely to find any direct evidence, at least not till the next spacecraft – the US Mars Observer – heads out in that direction in 1992.

Technology transfer

Those of us who spent our younger days sticking transfers on model Spitfires will probably warm to an idea developed by Bell Communications Research in Red Bank, New Jersey. For all practical purposes it's a way to transfer ultra-thin gallium arsenide devices on to silicon substrate, thus obviating the problems of co-fabrication.

The new technique, called epitaxial liftoff, involves first constructing a gallium arsenide device such as a thin film laser on a conventional 0.5mm thick substrate. But the clever bit lies in first depositing a seven-molecule-thick layer of aluminium arsenide between the substrate and the active components. Finally a layer of mechanically-supportive wax is added to protect the device.

As it stands this looks much like any other gallium arsenide chip: more than 99% substrate. The active components and the aluminium arsenide are all contained in the top 0.001mm, a tiny part of a standard device. It's this bulky substrate that places constraints on conventional GaAs technology, being a poor thermal conductor and having different lattice constants from those of silicon.

Bell's ingenious idea (*Photonics Technology Lett.* February 1989) involves dunking the completed wafer, with its extra layer, into a bath of hydrofluoric acid. The acid eats its way through this aluminium arsenide layer, liberating the active components in much the same way as a dip in water liberates a paint transfer. Once freed, the thin-layer laser (or whatever) can then be deposited on anything from glass to silicon.

As a novel way of producing hybrid chips, this approach offers considerable possibilities. It also means that even if the new substrate is nothing more than a good heat-sink, the performance of many existing GaAs devices could usefully be improved.

Fusion: a damp squib

Left to right: Dr Stanley Pons; Marvin Hawkins, a graduate student; and Dr Martin Fleischmann.

My prediction that the bright star of cold fusion would rapidly fade (Research Notes, July, page 544) is not, as might be imagined, a source of smug satisfaction. Who in their right mind could be pleased when the scientific community indulges in an undignified scramble leading mainly to public ridicule and a considerable waste of resources? To recap on this saga could well add to the ridicule, though it's certainly a valuable insight into the pressures facing research scientists.

On March 23, two chemists, Martin Fleischmann from Southampton and Stanley Pons from Utah, announced at a press conference that they'd discovered how to make deuterium atoms fuse together in an electrochemical cell at relatively low temperatures. In making this claim they were effectively proposing a route to limitless energy at virtually zero cost. The pressures to publish – and hence claim proprietary rights – were enormous, especially in view of the parallel work being undertaken just along the road at Brigham Young University by Steven Jones and his colleagues.

Nature, due to publish papers by both groups simultaneously, decided not to publish that of Fleischmann and Pons because referees had called for additional information which the authors had been unable to provide in the time available. Jones's paper *did* appear (*Nature* Vol. 338, No 6218), though not without considerable editorial comment. John Maddox, *Nature's* editor, pointed out the vulnerability of the scientific community's reputation when two research teams make astounding claims for an experiment using heavy water when neither has undertaken even a simple control experiment with plain H_2O.

Despite cautious reservations from nearly all the world's responsible labs, it seemed as if everyone from North Korean universities to North Surrey housewives had jumped on the fusion bandwagon. Palladium futures soared. Yet in spite of the inability of prestigious organizations like Harwell and Caltech to detect any evidence of cold fusion, the subject refused to go away. Fleischmann has continued to argue his case, notably at the US Electrochemical Society, while other scientists have equally vehemently pointed out what they regard as gaping holes in his experimental procedures.

Were this a scene from some Brian Rix farce, we'd all be expecting a new character by now to emerge from the closet. True to its dramatic potential the fusion farce has presented us with just such a character in the shape of double Nobel prizewinner (and vitamin proponent) Linus Pauling – now a sprightly 88. In a letter to *Nature* (Vol. 339 No 6220) Pauling argues that the excess heat observed by Fleischmann *et al.* could be the consequence of a chemical reaction in which an unstable compound, palladium deuteride, decomposes spontaneously to its elements.

Cold fusion may well have died the death, and with it the hopes of those who might have made a fast buck. But will there be any real losers? And, as Philip Campbell asks in *Physics World* (Vol. 12 No 5), can there be anything wrong with science as a spectator sport so long as spectators are not put at risk? Even if they do turn out to be wrong, men like Martin Fleischmann could well earn their place in history by setting other minds racing productively. John Maddox observes, for example, that the fusion rush is already teaching us that metal hydrides are an even more interesting field of research than had previously been thought.

One is still entitled however, to ask why certain unproven concepts attract public attention while others, such as ball-bearing motors, don't. The answer must regrettably lie in the answer to one simple and unscientific question: will it make money?

Cod piece

I have frequently observed in these columns how far electronic measuring instruments have come since the first Avo appeared on the market. But if a digital Aids meter is still a few years off, the same is not true for an invention from the Department of Materials Science and Engineering at Nagasaki University. There they've developed a sensor likely to form the basis of every Japanese housewife's dream – a device that measures directly the freshness of raw fish!

Fish freshness factor, K (say that without your dentures) is defined by the percentage of two chemicals, inosine and hypoxanthine, present in the raw fish. Hitherto that has only been measurable by destructive testing (i.e. cooking) – not very popular in oriental cuisine. What the Nagasaki engineers have done is to develop an electronic sensor that measures, not inosine or hypoxanthine, but a smelly gas, trimethylamine, which is also given off in progressively greater quantity as fish becomes time-expired.

The sensor, (*Platinum Metals Review*, 1989, Vol. 33 No 1) was developed initially using stannic oxide doped with gold, palladium or ruthenium. When coated on to an alumina tube heated from a coil inside, this device will respond to a trimethylamine concentration of 300p.p.m. at a temperature of 555°C. Other sensors, equally sensitive, have been developed using zinc, tungsten or titanium oxides with small additions of ruthenium.

When practical sensors were built and tested on actual fish, the Japanese researchers found that the readings could be calibrated and reliably compared to the so-called K values determined by chemical analysis. It looks therefore as if it won't be long before we'll be able to challenge the "caught yesterday, guv" claim with the same digital precision we use to turn out a perfect cod mornay.

Research Notes are by John Wilson of the BBC World Service's science unit.

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ENTER 31 ON REPLY CARD

NICAM DIGITAL STEREO

Television sound moves into the digital age with the introduction of NICAM for terrestrial and DBS transmission. Paul Gardiner of the IBA explains how it works.

Studies by the European Broadcasting Union in the early 1980s indicated that an audio bandwidth of 15kHz was quite adequate for broadcasting to the home; indeed, subjective tests showed that increasing the bandwidth further offered no significant improvement in perceived quality. The process of converting an analogue signal into digital form involves sampling the analogue signal at a rate of at least twice the highest frequency. A 15kHz audio bandwidth can be obtained using a sampling frequency of 32kHz, and this rate was selected both for terrestrial NICAM 728 and for high-quality sound in the MAC/packet system.

An initial coding accuracy of at least 14 bits per sample is desirable to represent the analogue sound signals accurately. If fewer bits are used, the quantizing error can become audible in the form of a 'gritty' quality for low-level signals, an effect known as granular distortion.

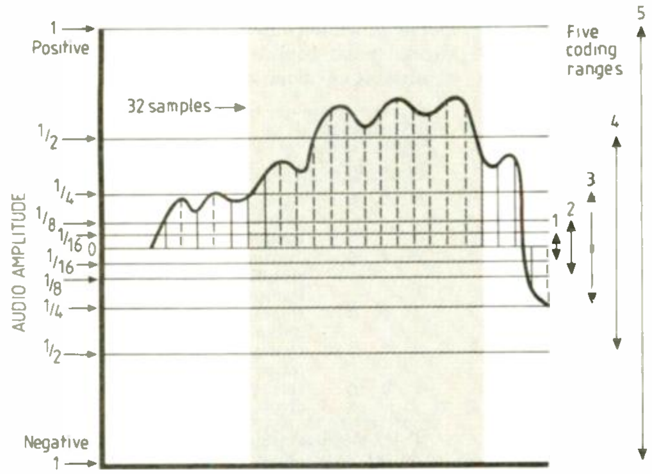

Fig. 1. Groups of 32 successive audio samples are treated as separate blocks and coded into one of five ranges, depending on the amplitude of the largest sample in the block.

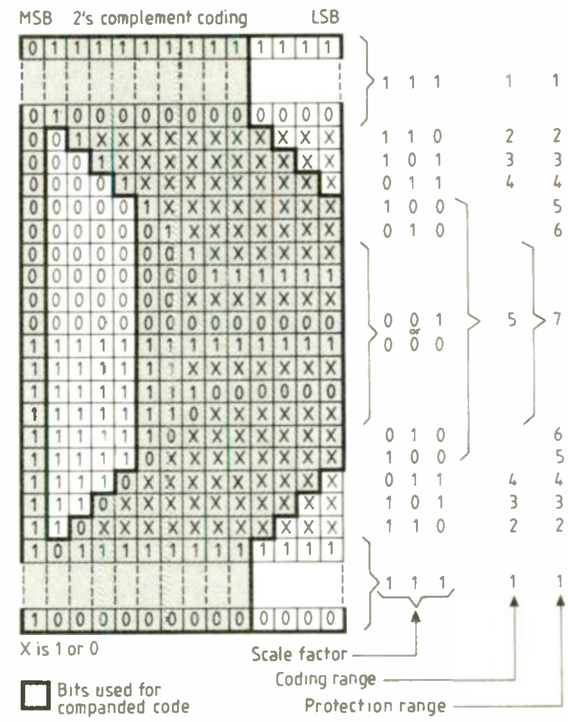

Fig. 2 (right). Coding of companded sound signals.

The use of 32kHz sampling with a coding accuracy of 14 bits per sample (i.e. 'linear' coding) would imply a transmission data rate (including parity bits for error detection) approaching 1Mbit/s for stereo. This rate is too high to be easily accommodated (in the form of a digitally modulated subcarrier) within the 8MHz channels of System I, bearing in mind the need to ensure that the introduction of any new signals must not cause interference to pictures or sound on existing receivers, on either co- or adjacent channels. Near-instantaneous digital companding enables the number of bits per sample to be reduced from 14 to 10, with virtually no perceptible degradation in sound quality. The minimum bit rate for two sound channels is reduced to about 640kbit/s, to which must be added the overheads needed for reliable transmission, such as framing words, parity bits and control information associated with the companding process. In its terrestrial application for dual-channel sound, the end result is a 728kbit/s data signal.

Near-instantaneous companding

With analogue companding systems (which generally have the objective of improving the signal-to-noise ratio), accurate matching of the expander characteristic in the decoder to that of the compressor in the coder is important in order to minimize audible 'pumping' effects. However, in a digital companding system (which has the objective of reducing the data rate), because all operations are performed in the digital domain, the coder and decoder can be matched precisely, avoiding any mistracking.

The companding technique adopted for both terrestrial television sound and for MAC/packet satellite transmission is that of near-instantaneous companding, based on the NICAM-3 system developed by the BBC in the 1970s. Essentially, sound signals coded initially with a resolution of 14 bits per sample are reduced to 10 bits per sample. However, for low-level signals, the receiver is able to re-create the original 14-bit samples. There is a loss of initial resolution only on high-amplitude signals. (NICAM techniques could, in theory, be used to reduce the number of bits from *any* given initial resolution; but throughout this article 'NICAM' refers to the 14-to-10 bit companding system specified in the NICAM 728 and MAC/packet standards.)

Companding is achieved by grouping successive audio samples (in two's complement form) into blocks of 32 samples (i.e. a duration of 1ms) and finding the largest sample in the block. The amplitude of this sample is then used to determine the way in which the entire block of 32 samples is treated. There are five coding ranges (**Fig.1**); range 1 represents a block where the largest sample falls between maximum amplitude and half maximum amplitude; range 2 from half to quarter maximum amplitude, and so on. Range 5 represents one-sixteenth maximum amplitude down to silence. The coding range for each 32-sample block is signalled by a three-bit scale factor word.

For a block of samples representing signals in the largest amplitude range (coding range 1), the four least significant bits of each sample are discarded. In the case of blocks falling in the second coding range, the next-to-most significant bit of each sample is discarded, along with the three least significant bits. (The most significant bit is always retained to indicate positive or negative signal excursions.) However, because the 32-sample block is 'labelled' as falling within coding range 2, the decoder can re-constitute the missing next-to-most significant bit, since this always has the same value as the most significant bit (**Fig.2**).

The effective resolution of the decoded signal for samples in range 2 is, therefore, 11 bits. In the third range the two least significant and two next-to-most significant bits are discarded prior to transmission, but the latter *two* bits are restored in the receiver (since they are identical to the most significant bit). The effective resolution is 12 bits. Similarly, lower amplitude signals in the fourth and fifth ranges are recovered in the decoder to 13-bit and 14-bit resolution respectively.

The companding process relies on the fact that the high-level signal itself masks programme-modulated noise. Prior to compression, pre-emphasis to CCITT Recommendation J.17 is applied, either by using an analogue pre-emphasis network (**Fig.3**) before digital conversion, or by using digital filters

WHAT IS NICAM?

NICAM is an acronym for Near-instantaneously companded audio multiplex, and is essentially a digital bit-rate reduction system designed specifically for high-quality sound. 'Companding' simply refers to a compression process applied to the audio signal prior to transmission, followed by an equal and opposite expansion process in the receiver. NICAM was devised by the BBC in the late 1970s with the aim of conveying sound programmes on digital circuits designed for telephony.

In the early 1980s there was considerable interest both in the UK and in Europe in the possibility of adding dual-channel sound to the existing terrestrial services, either for multi-language sound or for stereo. Various analogue schemes were considered, but it became clear that a digital system would offer considerably improved quality in terms of signal-to-noise ratio, lower distortion, and complete absence of crosstalk between channels. The BBC carried out intensive development work and field tests of a proposed new digital system, joined in the latter stages by the IBA (much of whose resources had been devoted to development of the MAC/packet specification for DBS).

This led to a joint BBC/IBA/BREMA specification of NICAM 728 (728 refers to the digital bit-rate of 728kbit/s). This system is now approved by the Department of Trade and Industry as the UK standard for two-channel digital sound with television on the terrestrial networks. Although it was originally designed for UK System I, it has attracted considerable interest overseas. Extensive tests have proved the ruggedness and compatibility of the system, which makes use of an additional low-level digitally modulated carrier. The European Broadcasting Union has recommended that those members wishing to introduce digital multi-channel television sound transmission should base their choice on NICAM 728, and the system is being considered as a possible international standard by the International Radio Consultative Committee (CCIR).

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with the digital signals. A corresponding de-emphasis network is used in the receiver decoder following the expansion process. This reduces significantly the audibility of programme modulated noise arising from the companding process.

Frame structure

For terrestrial transmission, the transmitted serial data stream is partitioned into 728-bit frames, transmitted continuously every 1ms; the overall bit-rate of 728 kbit/s is made up as follows:

8-bit frame alignment word	8kbit/s
5 bits for control information	5kbit/s
11 bits for 'additional data'	11kbit/s
704 bits for sound and parity	704kbit/s
total:	728kbit/s

This multiplex is used to modulate an additional low-level carrier (at 6.552MHz above the vision carrier for System I) which is added to the conventional broadcast signal.

The 704-bit sound/data block consists of 64 11-bit sound samples (a single parity bit is added to each sample to protect the six most significant bits), made up of a coding block for each of the two sound channels. In the case of stereo sound, the 32 samples for each of the A and B channels within each sound/data block are interleaved (odd-numbered samples convey the A channel, see **Fig.4**). If used for two independent mono sound channels, alternate frames convey data from each of the two audio channels. Alternatively, the system can provide a single mono sound channel and a separate data channel, or can be devoted entirely to data transmission.

In each frame, the first eight bits comprise a frame-alignment word (01001110) for receiver/decoder synchronization.

The first of the five control information bits (the frame flag bit C_0) is set to 1 for eight successive frames, and to 0 for the next eight frames. The first frame (Frame 1) is defined as the first of the eight frames in which $C_0=1$; the last frame (Frame 16) of the sequence is the last of the eight frames in which $C_0=0$. This 16-frame sequence is used to synchronize changes in the type of information being carried in the data channel.

The next three bits (the application control bits C_1, C_2 and C_3) indicate the type of information being carried by the current 704-bit data block. Four possible applications have been defined: stereo sound, two independent mono sound signals, mono sound and separate 352kbit/s data channel, and a single 704kbit/s transparent data channel (see **Table 1**). In the case of IBA transmissions, these bits will usually be set to 0, since for the time being it is planned to use NICAM 728 mainly for stereo sound. If it is required at any time to change the application, the forthcoming new application is signalled to the decoder by a change in the control bits on Frame 1 of the *last* 16-frame sequence of the current application.

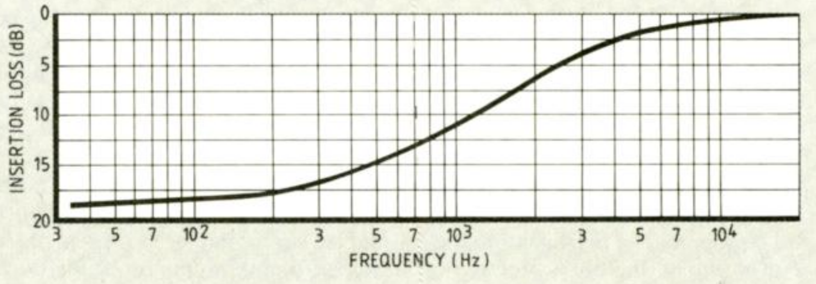

Fig.3. CCITT J-17 pre-emphasis is applied prior to the companding process.

Table 1: application of the 704-bit sound/data blocks of NICAM 728 to provide either stereo sound, mono sound, transparent data, or a combination of mono sound and data.

Application control bits C_1	C_2	C_3*	Contents of 704-bit sound/data block
0	0	0	Stereo signal comprising alternate A-channel and B-channel samples.
0	1	0	Two independent mono sound signals (designated M1 and M2) transmitted in alternate frames.
1	0	0	One mono signal and one 352kbit/s transparent data channel transmitted in alternate frames.
1	1	0	One 704kbit/s transparent data channel.

* $C_3 = 1$ provides for signalling additional sound or data coding options. When $C_3 = 1$, decoders not equipped for these additional options should provide no sound output.

The fifth control bit (the reserve sound switching flag, C_4) lets the receiver/decoder know whether the conventional analogue FM subcarrier is carrying the same programme sound as the digital signal. When the FM signal is *not* carrying the same information as the digital stereo (or primary digital mono) sound signal, C_4 should be set to 0. (Should there be two separate digital mono sound chanels rather than stereo, C_4 would relate to the content of the mono sound channel transmitted in odd-numbered frames). A use of this control bit can be to prevent a receiver switching to different FM sound should there be a 'drop-out' in reception of the digital signal. However, this control bit should not be relied on for the automatic selection of television programme sound in the receiver; the broadcasters recommend that receivers should be switchable manually to select either FM or digital sound.

Use of the 11 bits of additional data (i.e. an 11kbit/s data signal) has not yet been defined.

Bit interleaving is applied to the 704-bit sound/data frame in order to minimize the effect of multiple bit-errors. This results in adjacent bits in the transmitted serial data stream being separated by a 16-bit interval, so that provided an error burst spans 15 or fewer bits it will be spread as single bit-errors in the sound samples after de-interleaving in the decoder. Any errors affecting the six most significant bits of any sound sample will be detected by the parity check and concealed. Errors in the least significant bits will not be detected but, should they occur, will probably not be subjectively annoying to listeners. The decision to apply parity checking and error concealment to six MSBs was made after conducting listening tests during the development of the MAC/packet system. If more bits are protected, there is a greater chance of an undetectable multiple error; however, if fewer bits are protected, then the effects of unconcealed errors in the least-significant bits become more disturbing.

Scale factor information

The decoder needs to know the appropriate three-bit scale factor word used for each 32-bit sound-coding block. This information is conveyed without the use of dedicated bits by using a technique known as 'signalling-in-parity'.

Two sound coding blocks are contained within each frame so that, for each frame, six scale factor bits must be transmitted. The principle of operation is to take a group of nine samples within a coding block, and to signal one bit of the scale-factor by allocating either even or odd parity to each sample of the group of nine. If it is required to signal a scale factor bit of 0, then the group of nine bits is allocated even parity; odd parity is used in order to signal 1. This form of signalling is effective because, under normal reception conditions, it is most unlikely that there will be several errors within each group of nine. The receiver checks each sample for parity in the normal way, and compares the results with the transmitted parity bits for each group of nine samples.

For those groups in which the majority of samples have odd rather than even parity, the scale-factor bit signalled by the group is taken to be a 1. Error concealment can then be applied to any samples in the group as necessary. 'Signalling in parity' is very robust, since at least four samples in a group must be in error before a wrong decision on scale factor can be taken by the decoder.

The scale factor bits also provide a further form of protection against errors. In addition to signalling one of the five coding ranges, seven protection ranges are signalled **(Table 2)**. These allow the receiver to make certain deductions about the most-significant bits of the incoming samples of the group. For example, in protection range 6, the six MSBs should all be the same (Fig.2). This makes it possible to identify errors in the MSBs even if the parity check indicates that all is well (as a result of multiple bit-errors). Majority logic can then be used to correct these errors.

To retain compatibility with NICAM in its application with the MAC/packet system, there is no protection range 8. Scale factor 000 indicates protection range 7 (as does 001), but in the MAC/packet system is used to manage receiver buffer storage of the data packets which make up the sound channel. ■

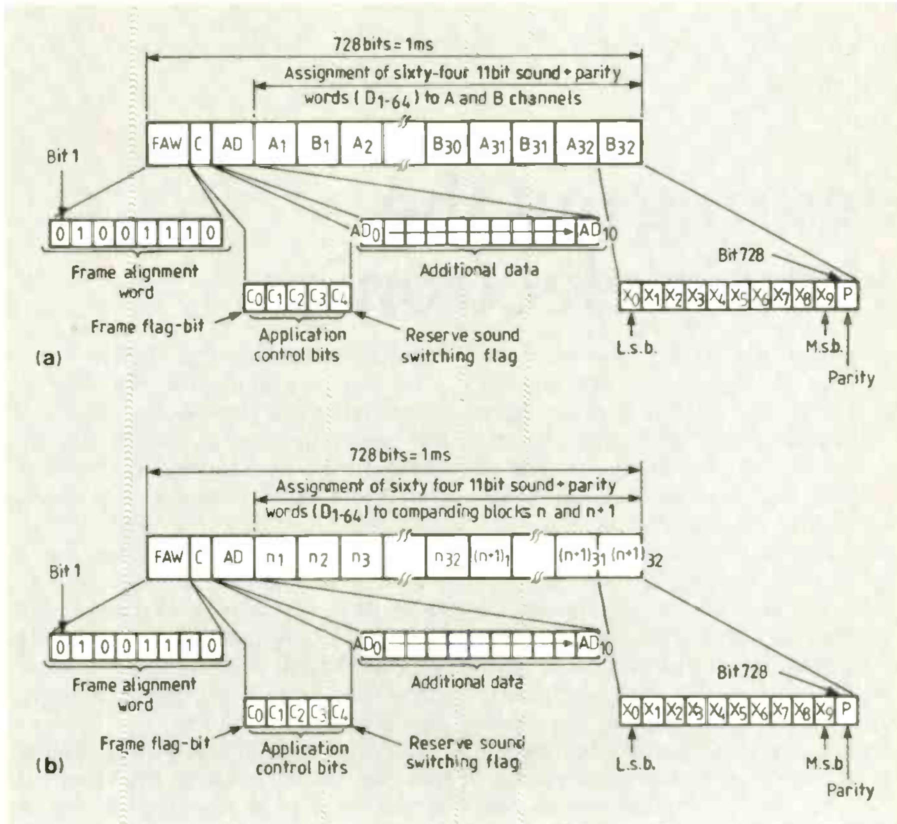

Fig.4. Structure of a 728-bit frame containing (above) stereo sound and (below) a single mono sound signal (before interleaving).

Table 2: coding and protection ranges associated with each three-bit scale-factor word. The five coding ranges indicate the degree of compression to which each block of samples has been subjected for the near-instantaneous companding process.

Coding ranges	Protection ranges	Scale factor value R_2	R_1	R_0
1st range	1st range	1	1	1
2nd range	2nd range	1	1	0
3rd range	3rd range	1	0	1
4th range	4th range	0	1	1
5th range	5th range	1	0	0
5th range	6th range	0	1	0
5th range	7th range	0	0	1
5th range	7th range	0	0	0

Pictured on page 754: the IBA's transmitting station at Emley Moor in Yorkshire, which, together with the London station at Crystal Palace, will launch a stereo sound service on ITV and Channel Four Television. Dependent relays will also carry the NICAM 728 signal.

To be continued.

Figuring out the trade in electronics

Sensitivity towards large negative trade figures in electronics and information technology also surfaced when Alan Clarke, Minister of State at the Department of Trade and Industry, confirmed the figure of £2.3 billion as the UK's trade deficit in high-technology for the year ending September 1988.

However, the ensuing exchange produced a veritable smokescreen of seemingly contradictory figures from the Minister, who asserted three propositions with respect to the deficit: (i) "France, Italy and Germany are all in deficit in sums vastly greater than ours"; (ii) the UK's exports in IT and electronics "are worth £11 billion and the deficit accounts for less than 5% of the total"; and (iii) "the Community has a deficit of £7 billion". All of this left observers consulting their Japanese calculators wondering how 5% of £11 billion makes £2.3 billion; and how, if four of the Community's major trading partners have large deficits, its total trade deficit can be only £7 billion.

By way of explanation, the Government argues that it is wiser not to spend too much time on a single isolated number such as international trade in electronics (since this gives a wholly misleading picture); far better is the consideration of the economy as a whole. For example, a single investment such as Fujitsu's IT component plant in County Durham will, when in full production, simultaneously reduce imports and improve exports, so altering the figures completely.

Unfortunately, this holistic argument has one major political drawback: it cannot be used in reverse. If the economy as a whole is in trouble, it will be very difficult to pinpoint one single feature (for example the UK's formidable record in the export of instrumental control systems) and then claim that single example as evidence of improvement in the overall economic outlook.

They're leaving school

The number of science graduates who re-enter or leave the profession is an important measure of the quality of skilled teachers who can teach electronics in schools. Although the figures show that most returning scientists are women in the 30-40 age bracket (returning after child rearing), the overall loss in science graduates is severe and progressive. For example, 2110 maths and science teachers left the profession in 1984, and by 1986 that number was 2590; by comparison in 1986 re-entrant graduates totalled 850.

These figures have been seized upon by Opposition MPs. Jack Straw, Education shadow minister, stated that shortfall in maths teachers alone by 1995 could be 12 000, whilst Simon Hughes, the Democrats' Education spokesman, accused the Government of recoiling from the 20% science content of the national curriculum only because it knew that 20% could not be delivered. According to Hughes, the decision to opt for a 12.5% science content arose directly from the shortage of teachers and equipment.

Challenged to disavow the figures, Kenneth Baker, Secretary of State for Education, said that he "hoped and believed" that the combined effects of all government proposals "will lead to adequate staffing in those subjects in the 1990s." Such statements will only reinforce those in the electronics industry who are worried that there is too much dependence on 'hopes and beliefs' to solve the ever-widening high technology skills crisis.

The missing Link

Several Labour MPs have suddenly become interested in LINK, a joint public-private r&d programme announced two years ago and supported with £210 million of public money. According to Government figures, more than "40 companies and over 20 science-based organisations have already indicated that they wish to be involved". So far, 14 Link schemes have been announced and a breakdown of Link projects shows many (e.g. molecular electronics, advanced semiconductors, opto-electronics) that have implications for the electronics of the 21st century.

Labour interest in the topic is two-fold. Firstly, only £70 million has been spent and Labour identifies this as evidence of a lack of commitment in Government towards r&d. However, the Government has a well-established defence to this allegation, saying that it will commit public money to r&d projects only if private sources meet 50% of the costs, that such an r&d project is not near-market, and that there should be a reasonable prospect of a tangible product at the end of the project which can be successfully developed. If these criteria cannot be met, the Government has always argued, why should the project be supported in the first place?

The second reason is politically far more important. According to Alice Mahon, Labour MP for Halifax, in 1987 "the PM boasted that £210 million would be spent on the Link scheme". By blaming the Prime Minister, Labour hopes to pin any r&d policy failures on the leader of Government, not some junior Minister who can be sacked when the going gets tough.

Rescued by drink?

Robert Banks, Conservative MP for the Yorkshire town of Harrogate, has discovered the stark reality surrounding the UK's trade with Japan. He was told that, against the background of a crude trading deficit of some £4.8 billion last year, the UK's top ten imports included computers, fax machines, printers, disk units and photocopier parts, making half the list electronic components or electronic equipment. By contrast, the UK's top ten exports to Japan included paints, blended whisky, malt whisky, antibiotics, antiques and postal packages. Nothing in the top ten was remotely high-tech in nature. Perhaps the Government hopes that when the Japanese drink too much whisky, they will be incapable of exporting!

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Saws and superhets

Surface acoustic wave filters can confer dual-conversion performance on single-conversion UHF receivers, claims Peter Johnson of Quantelec

The problems involved in designing a conventional radio receiver are twofold: the receiver must amplify a signal from a level of approximately one nanowatt to several watts (that is around 95dB gain) with a good signal-to-noise ratio, and it must reject all signals but the desired one from a spectrum of RF transmissions ranging from 100kHz to tens of gigahertz. Over the years, this problem has been tackled using homodyne, TRF (tuned radio frequency) receivers, superheterodyne and a number of other configurations with varying degrees of success. Apart from a recent revival of the homodyne principle in digitally-synthesized domestic receivers, the "superhet" has eclipsed all other designs.

Figure 1 shows the general layout of a single-conversion superhet receiver for a fixed signal channel with a carrier frequency of 200MHz. The basic principle is to convert any incoming signal's carrier to a fixed intermediate frequency (IF). The incoming signal is mixed with a local oscillator (LO) sinewave of 199MHz, producing an input to the IF stage of 1MHz, the frequency to which this stage is tuned. Unfortunately, with all practical designs of analogue mixer, there will also be a spectral component at 199MHz and another at 399MHz. The problem of removing these unwanted components is not as difficult as at first appears, since the LO contribution is of constant amplitude and will be removed by the demodulator. 399MHz is sufficiently far removed from the 200MHz carrier to be easy to filter out by the IF amplifier's tuned circuits.

A more serious problem is the possibility of receiving a spurious signal at 198MHz, the image channel, which also produces an output of 1MHz at the IF input. This component must be removed before the mixer by a preselector (a narrow band-pass filter). Since the image channel at 198MHz must be attenuated by at least 60dB compared with the desired 200MHz signal, a very sharply-tuned preselector filter would be needed.

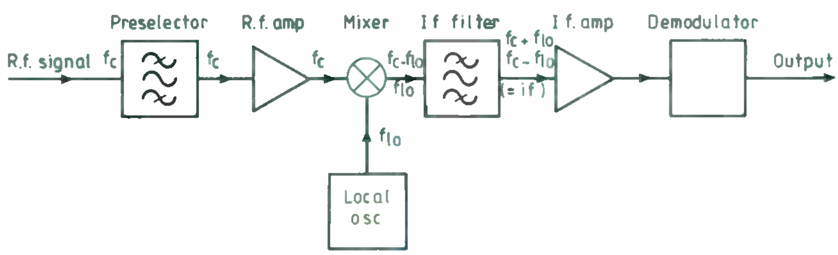

Fig. 1. Single-stage superheterodyne AM receiver block diagram.

One solution is to use a much higher intermediate frequency. In the above example, if an IF of 20MHz is used, the image frequency would be shifted to 160MHz and the preselector design is simple. However, now the IF amplifier is tricky to design and necessitates the use of crystal filters to achieve the required combination of gain and stability.

Dual-conversion superhets

The traditional approach here is to build a dual-conversion receiver. In this, the front end uses a high IF, but does not contribute significantly to the 90-odd dB of overall gain: unity-gain tuned circuits are quite easily implemented. A second superhet receiver then converts this high IF to something more manageable, usually the 455kHz used in commercial AM radios, where the necessary gain is achieved.

If the receiver in question must be continuously tunable over a frequency band, the dual-conversion design using a variable-frequency front-end filter is unavoidable, despite its principal drawback of using two complete sets of receiver components. Many receivers, however, are only required to tune to a few fixed channels, and can employ a series of switched, fixed-frequency preselector circuits; the problem then reverts to that of finding a set of filters of sufficiently high Q-factor to reject the image frequency from the channel information.

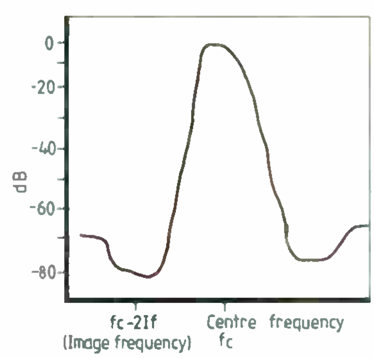

Fig. 2. Selective null response of saw coupled-resonator.

Saw filters

Saw filters are integrated passive devices with band-pass filter characteristics and consist of a quartz, synthetic ceramic, (e.g. lithium niobiate or tantalate) or other piezoelectric material as substrate, screen-printed with a layer of metal. Photo-etching techniques are used to produce a pattern of fine, interlinking finger-like electrodes (called interdigital transducers) which act as electrical input and output transducers when electrical energy is applied to the device.

The input transducer emits mechanical (or acoustic) surface waves which

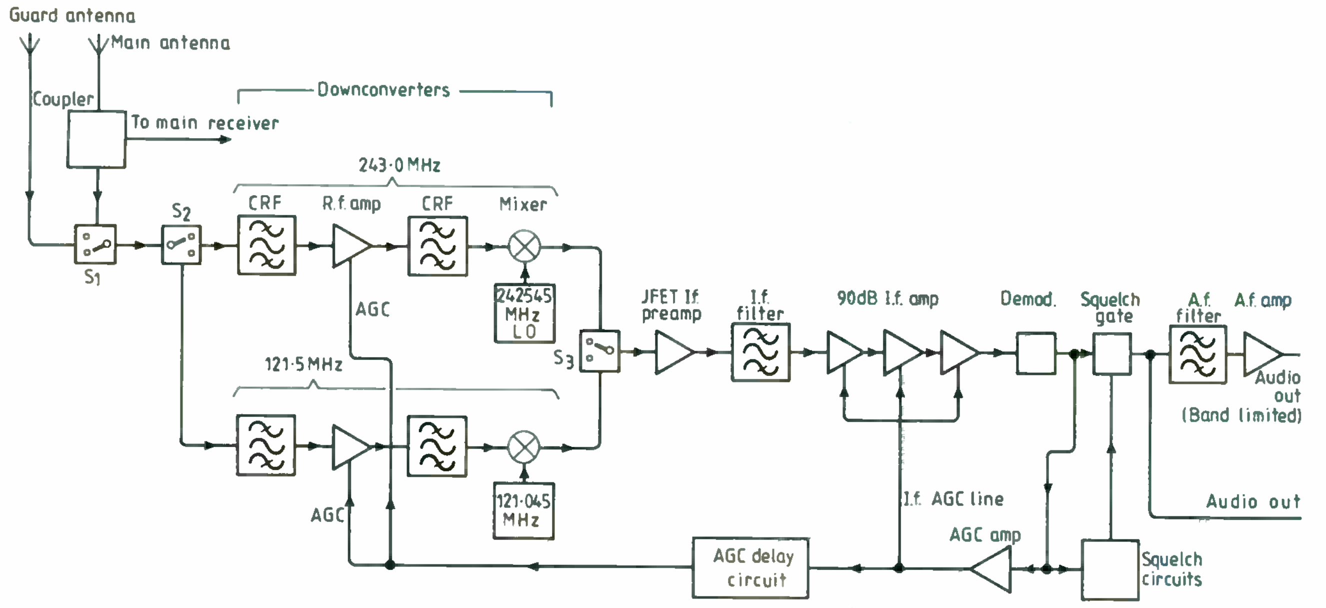

Fig. 3. ARC-182 Guard receiver, using saw CRF image-rejection circuit.

produce electrical energy in the output transducer, which can be "tapped off" at many points, making saw transversal filters and tapped delay lines possible.

The centre frequency, bandwidth curve and group delay are determined by the configuration of the interdigital transducers and a wide range of characteristics can be achieved by varying their arrangement.

Saw filters exhibit a high Q-factor, but the tuning is still not sharp enough for the type of application described above, i.e. that of separating a 198MHz signal from a 200MHz one. A much sharper cutoff of out-of-band signals can be achieved by arranging several stages of filter on the same substrate, each stage coupled to the next by the acoustic wave itself. This so-called coupled-resonator design can give an out-of-band rejection of 60dB and a Q factor of several thousand. Conventional filter response shapes such as Chebyshev and Butterworth can be readily synthesized and the device can be mass-produced at relatively low cost. **Figure 2** shows the frequency response of a typical coupled-resonator filter.

More important for a superhet design, it is possible to include a deep notch in the response at the image frequency (selective null placement). For example, saw coupled-resonator filters (CRFs) have been made with centre frequencies from 90MHz to 800MHz with a null response at the centre frequency minus 910kHz, enabling the designer to use a single-conversion receiver which will reject the image frequency of a standard 455kHz IF. At higher IFs (10.7MHz is widely used), image rejection can be arranged for carrier frequencies of up to 1.6GHz.

Guard receiver

As an example of a single-conversion receiver using a CRF image-rejection circuit, the US Navy's RF Monolithics ARC-182 is a transceiver which continuously monitors the two military emergency frequencies of 121.5MHz and 243MHz. A block diagram is shown in **Fig.3**.

Switch S_1 a low-loss GaAs fet switch, selects one of two antenna inputs and couples the RF signal to S_2, which is a combination band switch and AGC attenuator. It is implemented with pin diodes and is in a low insertion-loss state for the selected channel at low signal levels and exhibits high insertion loss when a large RF signal is present. The unused channel is in the high-loss state. The front ends consist of two low-loss saw bandpass CRFs, with centre frequencies of 121.50MHz and 243.00MHz and deep selective notches at the image channel frequencies for a 455kHz IF (i.e. at 121.059MHz and 242.090MHz), separated by a low-noise GaAs fet RF amplifier. The filters achieve an image channel rejection of better than 80dB.

The local oscillators are crystal based and generate 121.045 MHz for the 121.5 MHz channel and 242.545 MHz (using a 121.2725 MHz doubler) for the 243 MHz channel. These frequencies are combined in the mixer to produce the 455kHz IF, together with easily-filtered 121.045 and 224.545MHz components for the 121.5MHz channel and corresponding signals for the 243.00MHz receiver.

The IF signal is amplified by a JFET preamp and then filtered by a high-selectivity ceramic filter; the IF bandwidth is a minimum of 28kHz. Approximately 90dB of IF gain is supplied by the IF amplifier/detector IC; this chip, a Plessey SL6700A, also provides IF AGC and delayed-AGC functions. The detected audio signal is coupled to noise-squelch circuitry to generate a squelch output for control of the audio. There are two audio outputs, an unfiltered detected audio and a 3kHz band-limited output.

The ARC-182 Guard Receiver achieves a 10dB S+N/N sensitivity of −110.5dBm, coupled with a dynamic range of over 120dB, for a power consumption of around 1W. The unit is now standard equipment in USN patrol aircraft.

Using multi-stage selective-null coupled resonator saw filters, a single conversion superhet receiver can be built with a very large difference between IF and carrier frequency; in the above example, the carrier-to-IF ratio is 534:1, compared with 10:1 for a domestic VHF receiver or 3.5:1 for a medium-wave AM radio. Obviously, it is necessary to know the exact value of the image frequency, which implies a receiver with a number of fixed channels rather than a continuously-tuned set. The saw filters are sufficiently low in cost to make the use of a number of fixed channels a viable option, and are available in most standard UHF/VHF channel frequencies. For other frequencies, custom designs are feasible, since the design process is well-established and involves mainly changes to the metallization pattern printed onto the substrate. ■

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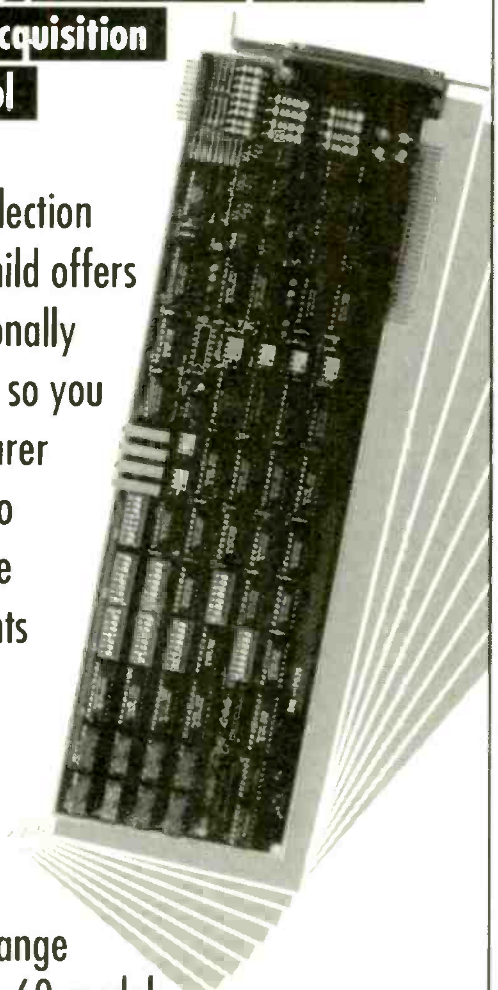

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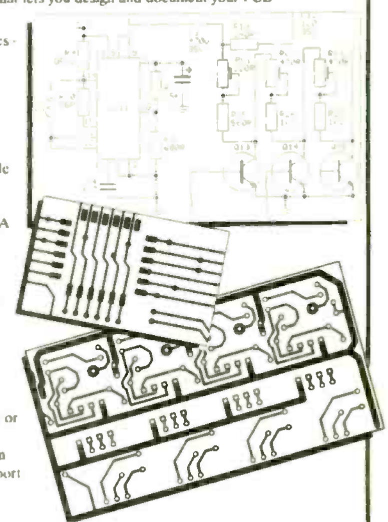

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Two-channel PWM with gap synchronization

Although not as power efficient in AM systems as PPM, pulse-width modulation offers self-clocked demultiplexing at the receiver and easy data recovery using a simple low-pass filter.

Overall system efficiency is enhanced if there is no need to output power when transmitting a synchronization pulse; this is the principle of the two-channel modulator/multiplexer shown here.

A square-wave clock driving a modulo-3 counter results in two channel-select wave-forms for analogue switches $S_{2,3}$ so the input channels are sampled at ⅓ the clock frequency. Resetting of the sawtooth generator is also provided by the input clock: comparing the sawtooth's positive-going ramp with a variable analogue input gives the desired PWM output.

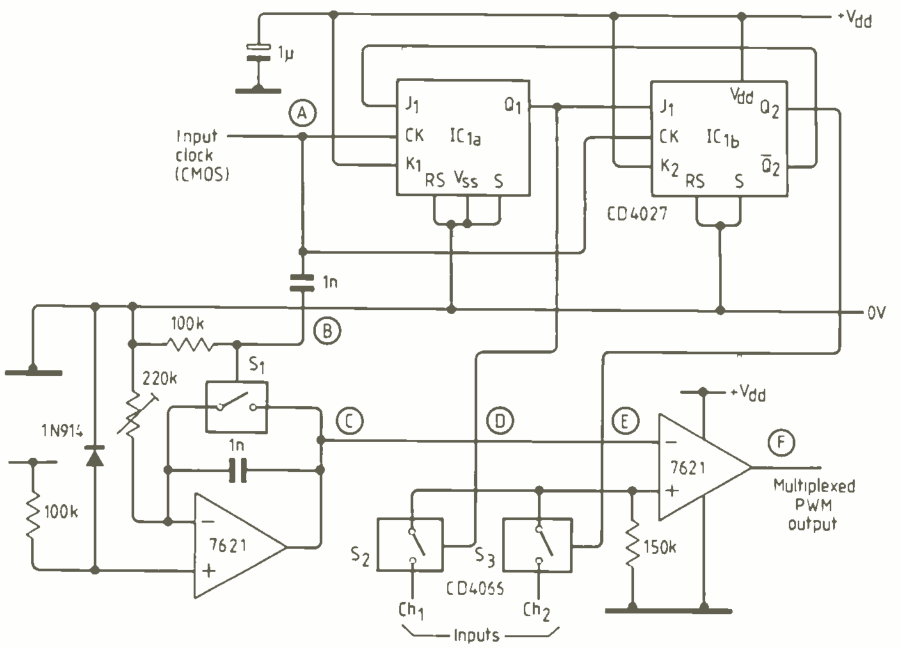
The 1:2 mark-to-space ratio of the counter results in a multiplexed output sequence of channel 1, channel 2, sync., gap, channel 1, etc.

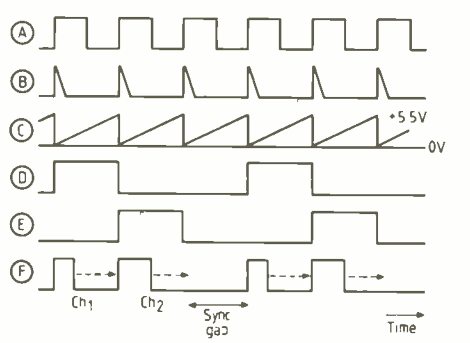
A demultiplexer for the PWM signal is also shown. Once again, a modulo-3 counter addresses c-mos analogue switches which route the serial pulsed input waveform to suitable filters at the appropriate times, thereby separating the two channels.

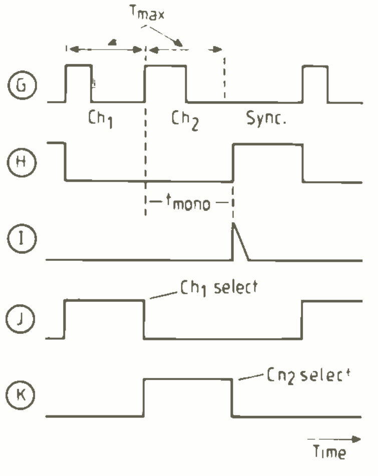
Resetting of the counter, caused by the gap in the pulse train, is done by retriggerable monostable device IC_3, output Q_2 remains low for slightly longer than one clock period when this device is triggered. A positive-going reset pulse is produced when a gap occurs.

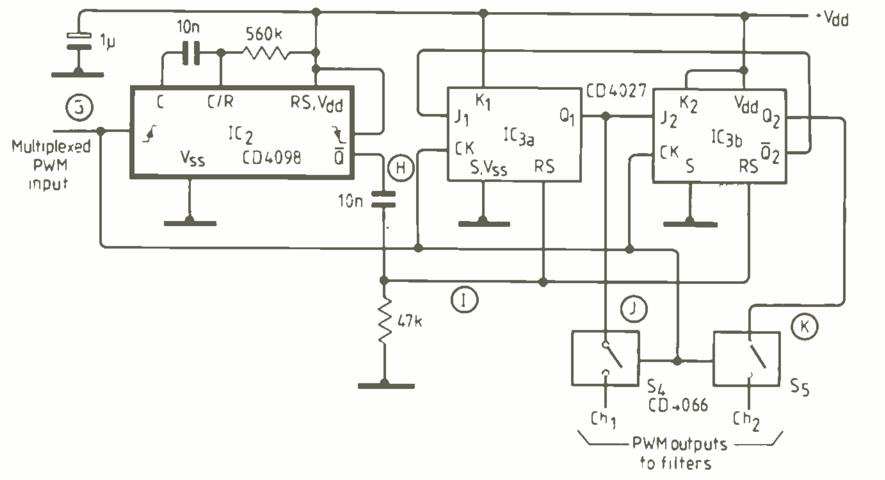
Component values shown are for system clock of approximately 600MHz.

N.E. Evans, University of Ulster







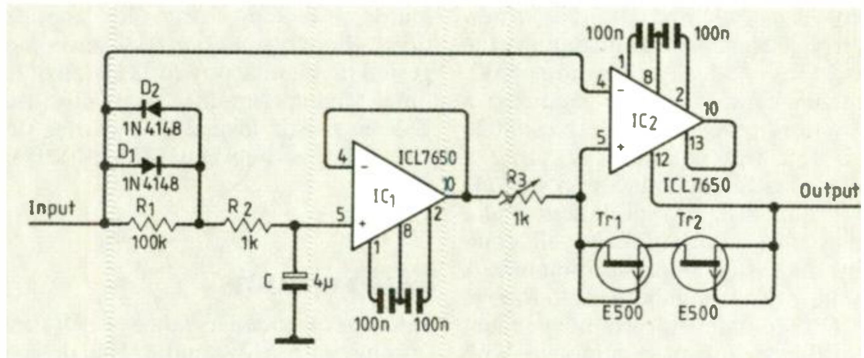


Fast DC-coupled trigger

This broadband trigger is immune to DC offset and base-line wandering. It permits high-quality performance over a wide range of input signals applied to the ICL7650 amplifier, IC_1, through the charge-storage network R_1, R_2, D_1, D_2 and C.

Diodes D_1 and D_2 quickly charge capacitor C to the input voltage. Output of IC_1 is then compared with the original signal at IC_2. Here, hysteresis is set by two p-channel field-effect transistors, $Tr_{1,2}$, wired as diodes in the IC_2 positive feedback loop; R_3, Tr_1, and Tr_2 ensure that the level of hysteresis is maintained for any DC offset at the input. Resistor R_1 provides a zero level for input signals whose amplitude is smaller than one diode drop and R_2 protects the input signal source.

In the circuit as shown, the output of IC_2, a chopper-stabilized operational amplifier, is fed back to the internal oscillator which helps square the output pulses. That is, the output at pin 10 is directed to the oscillator's input at pin 13 and a new output taken from pin 12 which is the amplifier's clock output. Because the oscillator has a divide-by-two counter the output will be one-half the input frequency.



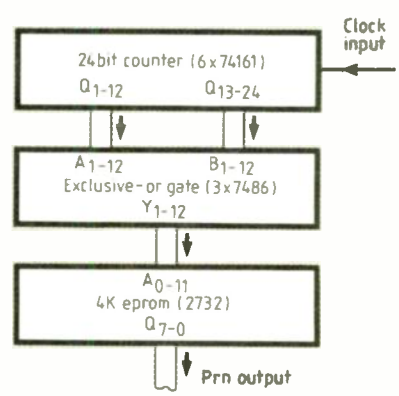
The trigger has a reliable triggering level and responds fast, its speed being limited only by the time constant $(R_{D1,D2}+R_2)$ C and can be adjusted to meet the requirements of practically any biomedical application.

Kamil Kraus
Czechoslovakia

Extending random number sequences

In this method of generating longer pseudo-random-number sequences from smaller ones, it is assumed that sequentially stored samples of a PRN sequence in a memory when read out randomly give another sequence.

In the circuit shown, samples corresponding to a 4Kpoint PRN sequence are stored in eprom and addressed by a 12bit counter. For larger sequencers, the most significant bit of the counter (13th bit onwards) are Exor-ed with the corresponding least significant bits and given as the address to the rom. This way the 4K PRN data is read out in many combinations and output as a different sequence for each possible combination of the most significant bits of the counter; as a result the sequence length is increased. This circuit has a maximum sequence length of 2^{24} clocks.



A. Dhurkadas
Naval Physical & Oceanographic Laboratory
Cochin
India

Ramp generator with wide frequency range

Only two integrated circuits and a few passive components are used in this ramp generator for a signal with adjustable level and frequency.

Negative current through R_1 produces the ramp's positive slope and causes the output of op-amp IC_1 to increase linearly toward +15V. Because the amplifier's output becomes the comparator's (IC_2) negative input, the comparator's output transistor switches on when the negative-input voltage exceeds the positive-input voltage.

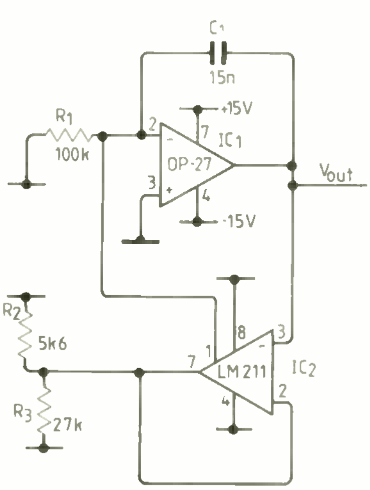
Switch voltage of the comparator is determined by the R_2R_3 divider. Switching IC_2's output transistor on forces the junction of R_2 and R_3 to 0V (value on negative input of IC_1). Current from R_2 decreases the discharge time of C_1 and allows IC_1's output to fall rapidly toward −15V. The comparator remains on until its negative-input voltage drops below 0V.

Output frequency can be expressed as

$$T_1 X \frac{1}{(V_{ON}-V_{OFF})/15}$$

where $T_1=R_1.C_1$; $V_{ON}=30$

$$\left(\frac{R_3}{R_2+R_3}\right)-15V;\ V_{OFF}=0V$$



Thus R_2 and R_3 provide adjustments for variations in both output frequency and peak-to-peak value of output voltage, then R_1 and C_1 for variation of output frequency only. This circuit works well in output frequency range from approx. 0.1Hz to over 100kHz.

Michele Frantisek
Barvicova
Czechoslovakia

Complex microcontroller

Motorola has released details of its new 32-bit 68300 microcontroller family, the first device of which will be the 68332. The chip consists of five modules; a CPU (compatible with the 68020 MPU), a timer-processor unit, a queued serial communications module (QSM), a 1.5Kbyte ram module and a system integration module, all communicating via an internal 16-bit bus. A version with a rom module is to follow. The TPU contains its own independent microcoded processor complete with ALU and is capable of operating almost independently from the CPU, once initialized. It can execute several standard timing functions on up to 16 channels simultaneously. The QSM contains a synchronous and an asynchronous serial port plus a queue to reduce CPU interrupt rates. The 68332, designed in conjunction with General Motors/Hughes Electronics subsidiary Delco Electronics for automotive uses, is suitable for many embedded controller applications, though some analysts have suggested that it may be a little ahead of its time. Manufactured in 1μm c-mos, the 132-pin plastic leaded chip carrier device will cost between $25 and $50 in volume.

Flash eproms

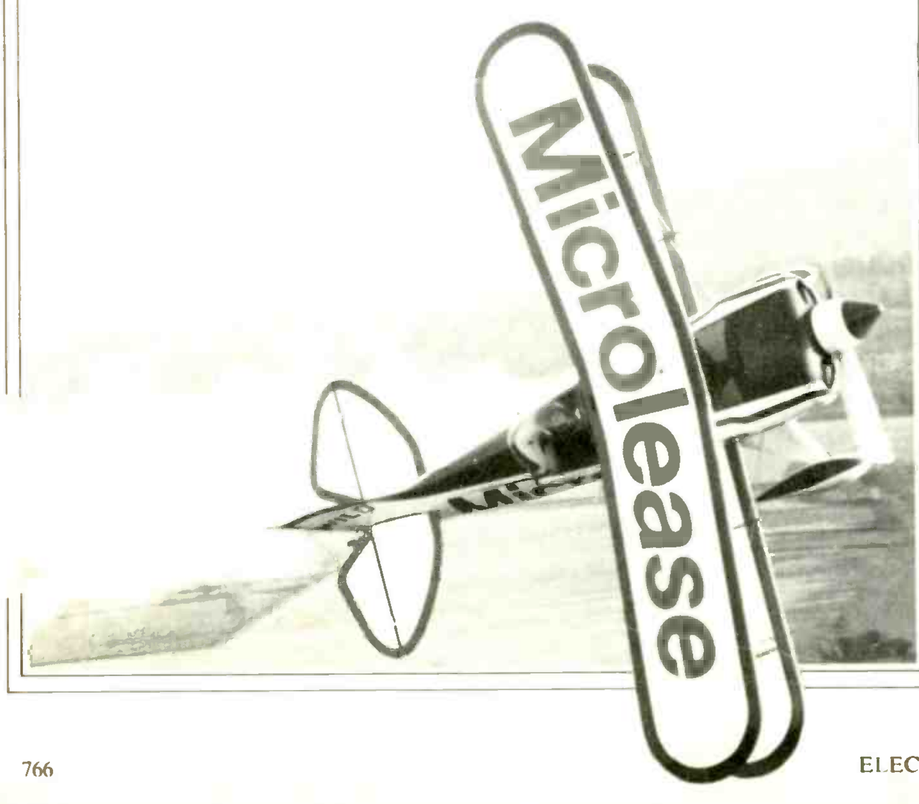
Intel has extended its range of 12V flash eproms with 512K and 1Mbit devices with a read access time of 120ns. The 28F512 and 28F010 are arranged as 64K×8 and 128K×8 respectively and can be bulk-erased in under one second. Programming can be executed in less than 1s for the 512K device, less than 2s for the 1M version. Applications include data acquisition in portable equipment and code storage in embedded systems which may require periodic updating in the field.

Qualified rental

Microlease has become the first UK equipment rental company to qualify under the BS 5750 pt2/BS 5781 approvals standard. Under the terms of the accreditation, all equipment supplied by the company has been calibrated to BS 5781 with traceability to national standards.

BS 5750 allows accredited companies to deliver components, equipment and services without infringing client assessed quality programmes. All the major hire organizations currently offer NAMAS calibration by special request; Microlease now offers QA equipment as a matter of course.

Brian Lecomber's Pitts special is sponsored by Microlease



New IT investment

The DTI and the Science Research Council are to organize a £22 million investment programme for information technology. Lord Young said that a total of £15 million would be put into opto-electronics with the balance going to the Information Engineering Advanced Technology Programme (IEATP).

Spending in opto-electronics will focus on research into the integration of opto-electronic devices and techniques within systems. Specifically, it will concentrate on optical communications systems and subsystems together with optical information processing.

Slowing electronics

The world electronics market will increase by 3.3% in the current year compared with 9% growth in 1988, says the latest edition of the Elsevier yearbook of world electronics. It predicts that future growth will stabilize in the region of 4%.

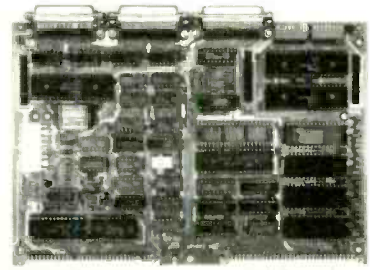
Prospects are brighter for optical discs. The research analysts Frost & Sullivan say that the European market for the technology will be worth some $900m by 1993. Source: Optical memory market in Western Europe.

Dominant Unix

Mr Sal Garcia, chief executive of the US CAE house VIEWlogic, claims "Unix will emerge in the next year to become the dominant PC platform". Speaking at the launch of the company's Workview Series II workstation software, he supported his statement by saying that "dos with its 640K limit represents a major deficiency allowing Unix to come into the market.

"Every day that goes by, OS/2 is losing ground. There is already a large number of technical applications designed to run on Unix and it is starting to pick up a few business applications", he said. He added that IBM itself had earmarked electronics design as a computing growth area based on Aix, the IBM-DEC version of Unix.

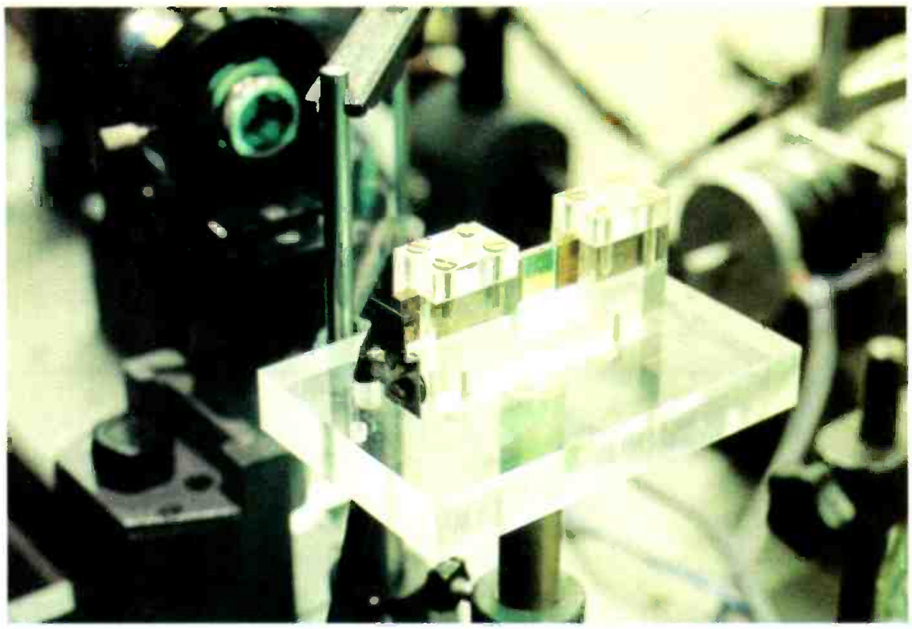
VIEWlogic's latest CAE software release, aimed firmly at customers who spend more than $1 million on CAE, makes use of the Unix X-windows interface and the VHSIC hardware description language. It employs open architecture meeting the EDIF standard in contrast to the Daisy/Mentor/Valid approach to CAE.



Research profile – Southampton University

MARTIN ECCLES

When photographer Ted Wright and I visited Southampton, we expected to cover the whole University in a day; but our look around the Institute of Sound and Vibration Research took up all of our allocated time. At some time in the future we hope to return to see the rest of the University.



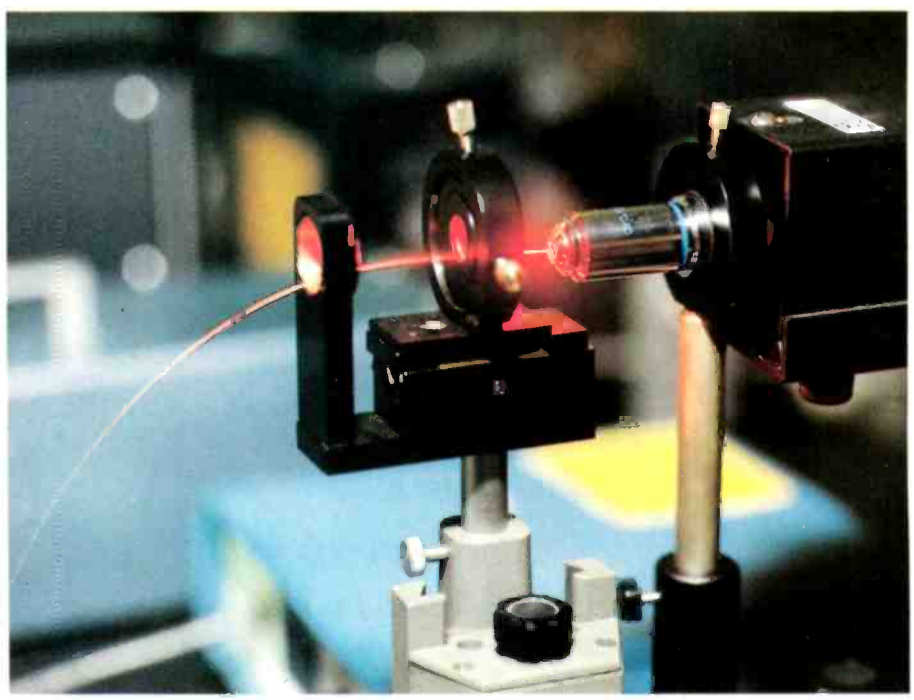
To find out the effects of, say, waves constantly beating against an oil-rig, detailed analyses of flow structures must be made.

This holographic crystal (bismuth silicon oxide) is part of a particle-image velocimetry analyser that measures speeds of particles in flow fields using high-speed photography. It is particularly useful for varying-flow fields (turbulence).

The autocorrelation required for the measurement process can be performed by a computer but this optical system, in which a convex lens performs spatial Fourier transforms, does the job twenty times faster.

The system is an optical implementation of the Wiener-Khin-Chin theorem.

Measuring out-of-plane motion in loudspeakers by laser interferometry is not new. This set-up is not to test the loudspeaker; it is an experiment to investigate enhancements to the measurement technique. Researchers at Southampton have developed new optical and electronic methods for reducing the problems caused by speckling of the laser beam on the reflective surface.



Laser vibrometers relying on the observation of Doppler shift to detect movement need optical heterodyning with a frequency-shifted reference beam if information about the direction of motion is to be obtained. This set-up is part of an experiment involving a new technique that provides directional information using only one laser beam and without a separate frequency shifter.

To obtain frequency shifting, a pseudo-heterodyning scheme is used. A sinusoidal signal at 1.05MHz drives a laser diode to provide optical phase modulation without the linearity problems associated with normal ramped-phased modulation. This technique gives a frequency shift capable of large vibration measurements of the order of 10^{-1}ms^{-1}.

Vibrometers benefiting from this research are insensitive to external vibrations, lightweight, inexpensive and robust.

High fuel bills are causing aircraft companies to consider replacing jet-engine planes with more efficient propeller-driven versions. One of the main problems with propeller-driven planes is the irritating cabin noise that they produce.

Digital signal processing combined with audio techniques can significantly reduce cabin noise not only in aircraft but in cars, lorries, buses, etc. Sound cancellation is only suitable for frequencies up to a few hundred hertz, but it is low-frequency noise that is most difficult to remove by conventional sound-damping methods. The objects you can see here hanging from the ceiling are microphones that feed adaptable-coefficient digital filters. Outputs from the digital processor feed the sound cancellation loudspeakers positioned around the room.

It is difficult to quantify briefly the results obtainable, but improvements from relatively simple sound-cancellation systems fitted in cars have been capable of reducing noise by more than 10dB. Perhaps in the future sound cancellation will be extended to reduce noise where it really hurts – in the houses situated under the flight paths of internally-quiet aircraft!

Inside this reverberation room with asymmetrical walls and ceiling is a horn, also with asymmetrical walls, that is subjecting an 'unspecified component' to 140dB of sound pressure at a range of frequencies. To see how the component reacts to high-intensity noise, it is fitted with 131 three-element strain gauges connected to computers and pen recorders.

In this experiment, readings from the test are being compared with results from a finite-element analysis program. According to researcher Tony Rogerson, the experiment is being done out of curiosity as much as anything. Soon, the horn will receive further padding in an attempt to take the level up to 165dB. Tony intimated that he is not absolutely certain what will happen at that level since it will be the first time that they have attempted 165dB with this set up and such a large sample!

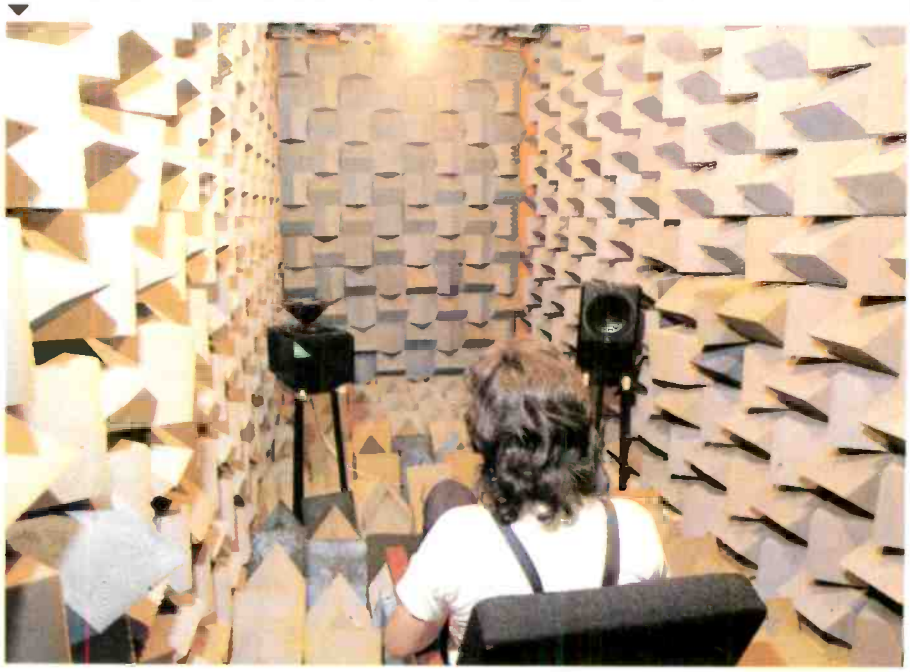
The horn – driven by chopped compressed air – is normally used by ISVR Consultancy Services to test aerospace parts; recently it has been vibrating electronics equipment housings that are to form part of a satellite for example.

Part of ISVR's consultancy work currently involves looking into new loudspeaker designs for a Japanese company wanting to break into the audio market with an innovative product.

In this room, one of the smaller anechoic chambers at the Institute, researcher Andrew McKenzie is listening to loudspeakers with conical reflectors. His work involves determining whether it is possible to achieve sound directionality without the undesirable associated effect of coloured off-axis response.

Conical sound reflectors have long been used to reduce the stereo image loss experienced when the listener is sitting non-centrally between the speakers, but in Andrew's experiments the conical reflectors are given an offset in an attempt to combine a degree of directionality with a good off-axis image.

Whether or not this experiment has produced the desired effect remains a secret; but Andrew did tell me that conical reflectors have opened up some interesting areas of stereo reproduction – namely uncoloured frontal and lateral reflections creating improved subjective sound.



World's fastest d-ram?

IBM claims to have produced a 1Mbit d-ram which can access any bit in just 22ns. This is about twice the speed of the best currently available 1Mbit products. The part has been produced at the company's Kyoto centre using what was claimed to be a standard d-ram production line.

IBM has employed a double polysilicon, double-layer metal c-mos process in the manufacture for the best combination of speed and density. It dissipates 500mW maximum on continuous cycling.

New, high-speed microprocessors are the driving force behind fast d-ram development. Slower generations of memory parts have forced system designers to use wait states where the processor twiddles its thumbs waiting for memory to catch up, or to use complex and expensive cacheing systems where fast static ram provides a buffer to slower memory.

- Hitachi says that the first two members of its 4Mbit d-ram family are now in mass production in speed ranges down to 80ns.

Fax on the move

A fax modem system which will tolerate drop-outs lasting up to two minutes virtually guarantees that fax links can be reliably set up with mobile users.

The device, manufactured by the UK company Intertec of Wimborne, Dorset, converts Group 3 fax signals into a handshaking protocol similar to the wire-based X.25 system. In the event of dropouts, fading or interference, the system retransmits lost packets until acknowledgement of their safe receipt is obtained by the receiving station. This provides virtually 100% assurance of successful transmission.

The built-in fax interface stores up to 10 pages of data for onward transmission on the radio link and provides buffers for received documents for downloading to the receiving fax machine. There is a guaranteed call-and-answerback protocol before any transmission takes place. The MR826 modem also has a built-in selective calling and addressing facility.

Go-faster stripes

Not so long ago I heard again that well-known kids' song, 'My Dad's Bigger Than Your Dad', which includes several verses with similar 'mine versus yours' content. Now the PC business is adding another verse – my PC's faster than your PC.

Not only does every company in the business (and some that aren't) now produce a PC based on Intel's 80386, 32-bit processor, but some are starting to jump up and down claiming to have the fastest versions, machines that use the latest 33MHz incarnation of the processor.

Compaq is the most recent combatant in this game of go-faster striped, machismo-laden, real-man machines, having just launched a 33MHz version of the Deskpro.

It does beg an interesting and important question, of course: is such speed either warranted or sensible? Let's face it, most 386-based machines are still running MS-DOS in single-tasking mode. In this context there comes a point where the advantages of speed stop being relevant. For the majority of users of 33MHz boxes there will be no real gain, except to the ego.

In fact, there could be several disadvantages. There are already indications that some DOS applications have a habit of falling over if raced too fast. Add-in boards can suffer the same fate as well.

The simple fact is, until a suitable operating system is alive, well and living in the hearts and minds of everyone, these go-faster systems have little real value to anyone except the manufacturers' bank managers.

And on the subject of operating systems, the main contender for working with 386 boxes is still Microsoft's (and IBM's) OS/2. But even the availability (just about) of this system is no great help for users of these raunchy machines. Again, the reason is simple. While MS-DOS straps them down to being go-faster 8086 processors, OS/2 straps them down to being go-faster 80286 processors. Not too much of an advantage, really.

In many ways, it would be sensible to wait until the 386-specific version of OS/2 appears, towards the back end of next year, before parting with good money on a 33MHz, 386 machine.

Real muscle

It would, of course, be a poor month without some news of IBM, and Big Blue has not let us down. Two new machines have been announced, one using the cut-down version of the 386 processor, known as the 386SX, and the other a 386-based portable.

The new boxes are additions to IBM's PS/2 line, with the SX-powered machine coming in as the Model 55SX, just above the 286-powered Model 50, and the 386 machine appearing as the company's latest attempt at producing a portable computer that someone wants to buy.

The PS/2 Model P70 has an impressive specification, coming with 60Mbyte hard disk minimum, 4Mbyte of memory and a 10-inch gas plasma display, as well as all the other bits you'd expect in a PS/2 machine.

It also has an impressive weight. it tips the scales at the best part of 10kg, which is no mean feat for a machine designated as a portable. Then again, as 386-based boxes seemed destined to be bought by 'real men', the biceps brigade will probably take to the machine with alacrity.

Cad on the Beeb

PCs are now used for most industrial cad applications but there are still many areas – particularly in education and r&d – where, as Martin Eccles reports, cad on the BBC microcomputer is still an attractive alternative.

You would not expect to see a BBC Microcomputer performing serious computer-aided design in a large electronics manufacturing company; but in many smaller companies, universities and r&d departments, where economy outweighs the need for a powerful industry-standard computer, there are still quite a few Model Bs and the like in use.

For designing the odd circuit or PCB, the BBC computer has one major advantage over the PC: it is relatively cheap, considering that it has high-resolution colour graphics. No doubt in departments on low budgets, just the fact that a BBC computer is there and an IBM PC is not will determine which cad software is viable.

```
Order of Filter       4b (2.69)
Maximum Passband loss dB = [illegible]
Reflection Coefficient %    [illegible]
Passband 25dB    4.1[illegible]
Minimum stopband Loss dB = [illegible]1.64
Cut off Frequency    10.00kHz
Stopband Edge Frequency    11.00kHz
Angle    67.[illegible]
Source resistance = 10.00k

                                  CC-04 61 69b

        Cap             Ind             Feal

 1      954.7pF
 2      2.582nF         74.91mH         11.44kH.
 3      2.675nF
 4        -             192.8mH         infinity
 5

Ro =  10.00k                    R6 =    zero
                                  CC-04 61 69b

        Cap             Ind             Feal

 1      1.205nF
 2      1.420nF         136.2mH         11.44kHz
 3      1.472nF
 4        -             226.9mH         infinity
 5        -

Ro = infinity                   R6 =   10.00k
                                  CC-04-61-69b

Frequencies of maximum passband loss:-
   zero          8.202kHz         10.000kHz

Frequencies of minimum passband loss:-
 4.935kHz        9.640kHz

Frequencies of maximum stopband loss:-
 infinity           11.44kHz

Frequencies of minimum stopband loss:-
 13.95kHz           11.00kHz
                                  CC-04-61-69b
```

One module in David Markie's elliptical-filter design program gives actual values for a singly-terminated filter as shown. It goes on to list normalized values and pole-zero locations. David is currently working on routines that output values for MF10 switched-capacitor filters.

The software discussed here – namely Markie's elliptic-filter design package, Miteyspice, Diagram and PCB – takes you through all the cad steps from implementing an idea to making PCB artwork, all on a 32Kbyte computer.

Elliptic filter design

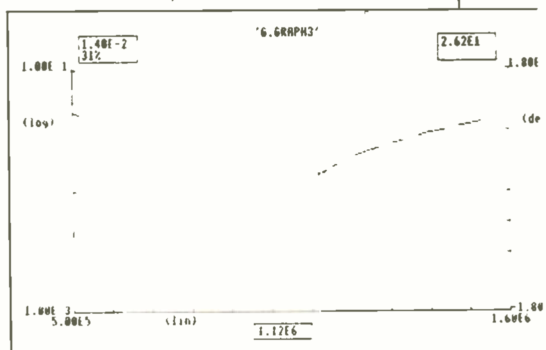
David Markie, who already sells a linear circuit analysis program for AC, has just completed a suite of routines for designing elliptical filters.

Low-pass, high-pass, band-pass and band-stop functions with variations for unterminated situations are all covered in the suite, which is based on the work of Amsutz, Daniels, Saal and Zverev. The programs are essentially for designing passive inductor/capacitor filters but the results for pole-zero locations and pass-band/stop-band specifications are equally applicable to other implementations including active and digital filters.

Initially, these programs will only be available for the BBC Model B computer but, if there is sufficient interest, David intends to adapt them for other computers. Their price will be £25.

Circuit simulation

For more general circuit simulation there is Miteyspice from Those Engineers. You may have heard of Microspice; more likely you will have heard of the Fortran Spice program on which the much more compact Microspice is based. Miteyspice is an enhanced version of Microspice, its main difference being that it can produce a graphical display as well as the numerical values output by its earlier counterpart.



Unlike its precessor, Miteyspice has graphical output.

Miteyspice for the Model B performs DC, small-signal AC and noise analyses for up to 25 nodes, while the Archimedes version runs much faster with a 200-node ceiling. You can use simple 'ideal' components or model devices including transistors, op-amps and transformers.

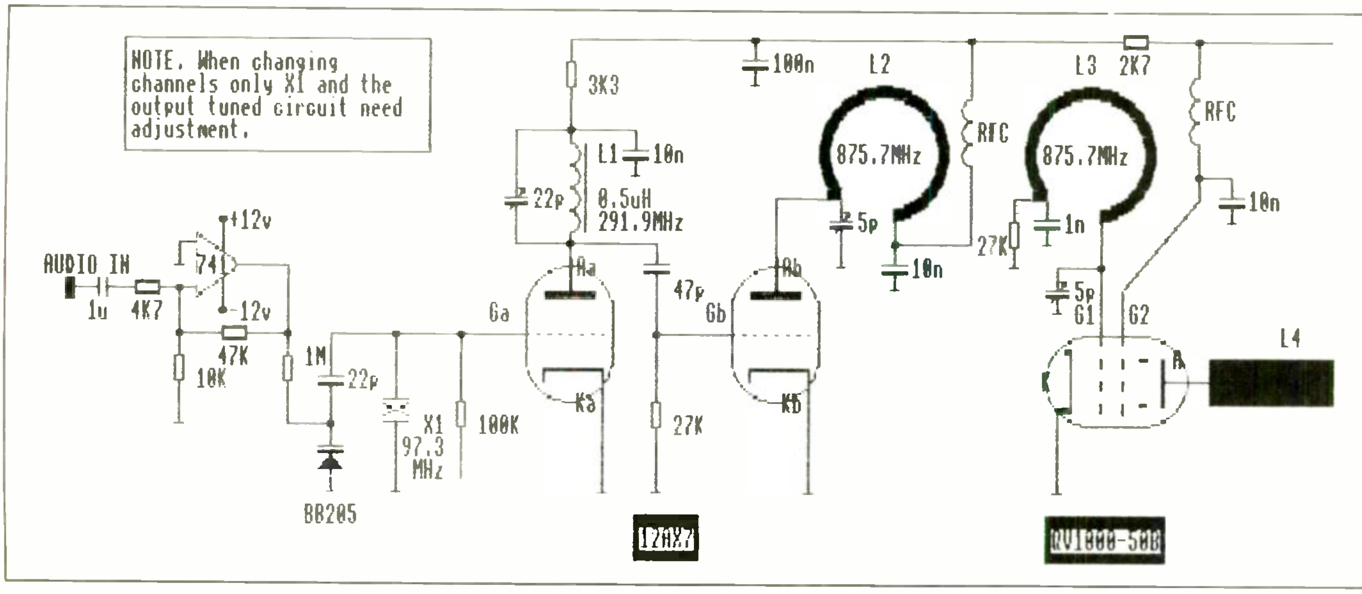
At its simplest, the software can be used to look at static operating points of nodes in a circuit and at the other extreme it can sweep selected component values or a range of input frequencies; with a user-written 'EXEC' file it can even do both together. Noise analysis may be used with either AC or DC sweep analyses to calculate total output noise and referred input noise at each step. A breakdown of noise sources in the circuit can also be selected to show the contribution of each to the total.

Prices of Miteyspice are the same for both the Model B and Archimedes computers at £119 excluding VAT and educational departments can buy a multi-user licence at £238.

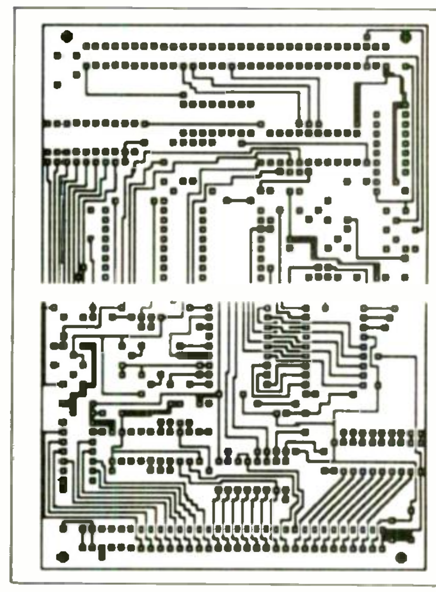
Making a drawing

There have been a number of drawing packages for the BBC computer and although most of them are fun to use, there is only one that I have ever found to actually save time when drawing circuits – Diagram from Pineapple.

Rather than using zooming and pan-



Artworks from Diagram, above, and PCB, right. The manual PCB layout has one or two blemishes, some of which could be ironed out with a bit of juggling.



ning to increase the amount of detail that a drawing can have, Diagram uses the CRT screen as a window on to a drawing with a fixed resolution. Because of this, and because of the way that the drawing is constructed in memory as pixel blocks, the drawing's size is limited only by disk capacity.

Diagram can be used to draw anything (in monochrome) but since it gives you the facility to define characters such as transistors and capacitors so that they can be inserted on the diagram with a couple of key-presses, it is particularly suited to circuit diagram draughting.

Connecting components is simple too; the cursor keys trace horizontal and vertical lines. There are also rubber-band and pixel-drawing facilities, but once you start using these the time taken to make a drawing goes up.

Even though scrolling of the window is quite quick, finding particular areas on a large diagram can be time-consuming; so any point on the diagram can be given an index name that puts the window over the area with a few key-strokes. You can save and copy areas of the diagram too.

One limitation of Diagram is that it can accommodate only 1024 different 8-by-8-pixel blocks. I have never reached the block limit even when drawing things other than circuits but if you are thinking of becoming an artist, buy a box of crayons instead. Diagram outputs scaled drawings to a dot-matrix printer with graphics mode and although the detail is adequate to convey all the necessary information, the presentation might be a limitation in some applications. The package costs around £155 excluding VAT.

Producing PCB artwork

When someone told me that PCB artwork could be produced on a BBC computer with a dot-matrix printer I did not believe it. Since then I have made at least 50 boards using Pineapple's PCB program and I could not go back to producing artworks manually.

The PCB program can handle double-sided boards up to 8-by-5.6in with tracks between pins, and it can produce 1:1 and 2:1 artworks. Although the package is designed for placing components on a 0.1in grid, it is possible to place pads on a 0.025in grid with some limitations; line drawing steps are 0.025in.

Recently, an autorouteing option for PCB has become available. It did not take me long to get used to using PCB but the autorouteing extension asks me whether it should choose north, south east or west biasing and whether or not I want it to route tracks in order of slope. When I give it random answers it does not make a very good job of the routeing.

Biasing and sloping are explained in the manual but I doubt whether I shall ever have the time to train myself to use them. But I would certainly make the effort if I produced boards on a regular basis. Although the autorouteing program would not let me out of it once I had entered it, it did at least do its best when I gave it a random input.

Using the basic PCB package, placing components like 40-pin ICs and large pads on to the layout takes only a few key-strokes and you can equally easily insert text on the layout. Deleting tracks is as simple as laying them, which is important if you make as many modifications as I do. With a little planning you can produce artwork for a fairly dense PCB; the main factors that prevent you from producing a *very* dense PCB are that you cannot pass tracks between pads that aren't on the main 0.1in grid and you cannot pass tracks at right angles between four pads.

Originally, PCB was designed for producing artwork on a dot-matrix printer with high-resolution graphics mode, but there is now an optional plotter-output routine. The main PCB routine is £85 and the auto-routeing extension costs £185, both excluding VAT. ■

Addresses

Markie Enterprises, 60 Park Drive, Sunningdale, Ascot, Berkshire SL5 0BE.
Those Engineers, 106a Fortune Green Road, London NW6 1DS, tel. 01-435 2771.
Pineapple Software, 39 Brownlea Gardens, Seven Kings, Ilford, Essex IG3 9NL tel. 01-599 1476.

ACTIVE

A-to-D and D-to-A converters

12bit 1MHz A-to-D converter. Two low-cost, 12-bit, sampling A-to-D converters are capable of digitizing sinusoidal input signals at up to 1MHz sampling rate. ADS-105 and ADS-106 combine a high-speed sample-and-hold amplifier and a 12-bit flash-derived A-to-D to overcome layout problems. Datel, 0256 469085.

Low power triple DAC. The HD49304 triple video c-mos D-to-A converter from Hitachi contains three independent 8-bit converters with a typical maximum conversion rate of 50 MHz and a maximum differential linearity error of ±0.2%FS. Features include a power dissipation of 300mW (typical) and a 42-pin shrink-dip package. Hitachi Europe, 0923 246488.

8-bit video A-to-D converter. The Plessey 8-bit A-to-D converter (SP94308) is designed for use with NTSC or PAL. A 1V video signal is AC-coupled to the device input, where it is DC clamped, amplified and fed through a buffer which drives the A-to-D converter input capacitance. Also within the SP94308 is an internal clock amplifier and driver. RR Electronics, 0234 270272.

Active hybrid circuits

ECL oscillators. A range of ECL-compatible oscillators offers a choice of positive or negative supply operation. The oscillators are available with or without built-in pulldown resistors, and also offer a choice of single or complementary outputs. They are available for frequencies up to 160MHz. Magna Frequency Management, 0223 892015.

Development and evaluation

Mac II-based development board. A Macintosh II-based development system board for Integrated Device Technology's IDT79R3000 risc processor is now available. The NuBus card, the IDT7RS201 with the R3000 microprocessor, plugs directly into the Mac II. The card is offered with base-level software, optional R3000 support software and optional Macintosh design support software. Micro Call, 0844 215405.

General microprocessors

T425 transputer. The IMS T425 is a pin-compatible upgrade for the IMS T414 32-bit transputer which offers additional features, including faster serial links and new pin functions. It offers an additional set of instructions, including 2-D block move and breakpointing. There are two new pin functions: a refresh pending pin used for holding a DMA request whilst refresh occurs, and an event waiting pin which allows the user to control external logic. INMOS International, 0454 616616.

Risc processor. Acorn's VL86C020 is a 20MHz, third-generation risc processor with a 4Kbyte on-chip data and instruction cache. It is compatible with the earlier VL86C010; its 64-way set-associative cache allows a threefold increase over the 010. The 020 works with standard 80ns DRAMs without wait states. VLSI Technology, 0908 667595.

Interfaces

Logic analyser card for PCs. From Flight Electronics is a PC expansion card which transforms an IBM PC (or compatible) into a low-cost logic analyser. The CLK-2450 system comprises the PC control card, software disc, three data input pods with signal line clips and a target-IC test clip. The analyser has 24 channels and is capable of sample rates of up to 50MHz. Flight Electronics, 0703 227721.

RS-232 driver/receivers. Two dual RS-232 driver/receiver chips reduce the size of charge-pump capacitors. LT1180 is a dual 5V RS-232 driver/receiver with a shutdown feature that puts the driver and receiver outputs in a high-impedance state, allowing RS-232 line sharing. LT1181 is supplied without the shutdown feature. Linear Technology (UK), 0932 765688.

IEEE-488 bus controller. The Prism 2020 can store up to eight bus control programs in a battery-backed non-volatile memory and can control two IEEE-488 bus systems. Control programs are generated on a microcomputer and down-loaded via the RS-232 port. STC Instrument Services, 0279 641641.

Linear integrated circuits

BiFET op-amps. The TL050 and TL030 families are BiFET single, dual and quad operational amplifiers from Texas Instruments, which possess maximum offset voltages down to 800 microvolts (compared to the 6mV of the TL071A). Minimum slew rate of the TL051/A is 15V/μs. The TL030 family form a low-offset alternative to the TL060 series, needing a supply current of 0.25nA to 1mA. Hawke Components, 01-979 7799.

Precision op-amp. LT1012A is a universal, precision operational amplifier with an offset voltage of 25 microvolts maximum and a drift over temperature of 0.6μV/°C maximum. Maximum supply current is 500μA, and bias and offset current are both 100pA maximum. All versions of the LT1012 will operate on ±1.2V supply rails. Linear Technology (UK), 0932 765688.

Fast, low-power buffer. Elantec's EL2001 monolithic 60MHz buffer has a slew rate of 1000V/μs and a quiescent current of 1mA. It will boost the output of ordinary amplifiers up to 100mA. Microelectronics Technology, 08446 8781.

Memory chips

Static rams. A range of fast, static rams for use in graphics and digital signal processing. Micron Technology SRAMs are available in various size/speed combinations, including 256K×1 down to 25ns and 16 or 64K down to 15ns. The SRAMs are fabricated in a c-mos double-metal process. Abacus Electronics, 0635 36222.

Ram/rom PC disc card. The PCL-790 range of PC ram and rom disc-emulator cards is designed as a replacement for mechanical disc drives. The PCL-790 disc card is a fully software-compatible, solid-state emulator of PC floppy discs. On-board firmware emulates a standard disc control card and allows power-on auto-booting as well as read/write modes of operation. Integrated Measurement Systems, 0703 771143.

Optical devices

Photodetectors. Now available from Centronic is a comprehensive range of photodetectors for visible, UV and IR radiation detection and measurement for military, space or commercial use. Although the standard range is extensive, the company will manufacture virtually any shape or size of silicon photodetector. Centronic, 0689 42121.

Photodiodes. The C30644 and C30645 avalanche photodiodes series APDs from RCA offer fast response and good quantum efficiency in the spectral range between about 1000 and 1600nm. They are optimized for use in optical-fibre communication systems at 1300 and 1550nm. Pacer Components, 0491 873077.

Standard logic circuits

12-bit A-to-D converter. The AD7870 converter is less than half the cost of earlier 12-bit designs and does not need a separate T/H amplifier or reference. It combines a 2μs T/H amplifier, a 3V buried Zener reference, clock and bus-interface logic on one BiMOS chip. Analog Devices, 0932 253320.

EMC feedthrough filters by Tesch

PASSIVE EQUIPMENT

Passive components

Film capacitors. Aerovox's SuperMet fluid-filled film capacitors have a design life of 60 000 hours at full rating. The metal-can units are protected against can explosion by a pressure-sensitive internal circuit breaker and the sealed case contains a corona-suppressing inert fluid. AVX, 0252 333851.

DC cooling fans. The range of Papst Multifan DC equipment-cooling fans offers adjustable noise level, low EMI and RFI and extreme tolerance of power supply variation. Five variants cover air movements of 19 to 47 cubic feet per minute. Dialogue Distribution, 0276 682001.

Chip electrolytics. Aluminium electrolytic capacitors from Nippon Chemi Con are designed to withstand cleaning by immersion, vapour or ultrasonics in Freon TE or TES for up to five minutes. Capacitance ranges from 0.1 to 68μF to a tolerance of ±20% at voltages from 6.3V to 50V. EEC Electronics, 0628 810727.

Ceramic capacitors. The RPL range of ceramic leaded capacitors from Taiyo Yuden are available with a choice of temperature characteristics and offer capacitance values from 100 to 47 000pF. They are rated at 50V DC and withstand 125V DC. ECC Electronics, 0628 810727.

Surface-mount Zeners. BZX 84C surface-mounting Zener diodes by Mistral are housed in industry-standard SOT-23 packages with tin-plated terminals. The devices exhibit breakdown voltages from 5.2 to 28.9V and maximum power dissipation of 350mW. Surtech Interconnection, 0256 51221.

Miniature tantalums. The Kemet series of resin-dipped solid tantalum capacitors is available to IECQ300201/US003 requirements in an extended range of 0.1 to 220μF, at 20% tolerance, from 5.6V to 50V. Unitel, 0438 312393.

Precision film resistors. The GC65 metal-film resistor, with a maximum body length of 10mm and diameter of 2.5mm, has a resistance range of 1MΩ to 1GΩ. Tolerances are available down to 0.1% with TCRs of 100ppm/°C. Welwyn Electronics, 0670 822181.

Connectors and cabling

Marine connector. The MK22 range of circular marine connectors features an extra-long coupling nut to permit easy coupling and uncoupling. The knurled surface allows connection and disconnection to be carried out even with a gloved hand. Gold-plated copper alloy is used for the contacts; working voltages are from 500V to 2.8kV at −55 to 125°C. AB Connectors (Northampton), 0604 712000.

Ribbon cable. Flat, jacketed cable type FBLDT is constructed from 28AWG (7×0.127mm) tinned-copper conductors with extruded PVC insulation. With a pitch of 1.27mm, the cable is suitable for IDC termination. Hayden Laboratories, 0753 88844.

Power filters. The Littelfuse Tracor range of power line filters is designed to protect equipment against mains transients. The range includes plastic-cased devices for board and chassis mounting. All are for use with 250V AC or DC at 0.6A to 30A. Highland Electronics, 04446 45021.

Displays

Hexadecimal displays. Hewlett-Packard HTIL-311A 4×7 dot-matrix hexadecimal displays include an IC to implement all decoder, driver and memory functions required to accept, store and display 4-bit binary data, off-loading display management tasks from the host processor. The binary data is displayed as the characters 0-9, A-F. Jermyn Distribution, 0732 450144.

LCD driver. The Teledyne Semiconductor TSC828 flexible, multi-digit LCD driver can be connected directly to the TSC827 A-to-D converter and simultaneously displays the A-to-D conversion result and high/low set-

NEW PRODUCTS CLASSIFIED

point values. Annunciator inputs, alarm logic outputs, a buzzer driver and all display decoder/drivers are included. Trident Microsystems, 0737 765900.

Filters

Digitally programmable anti-alias filters. Frequency Devices Series 848P8E are digitally programmable, low-pass, active filters intended for anti-alias applications. They are tunable over a 256:1 frequency range and provide an 80dB/0.03dB shape factor of 1.77, an 82dB stop-band floor and a 2-pole monotonic roll-off. Lyons Instruments, 0992 467161.

EMC feedthrough filters. The four ranges of Tesch panel-mounting, feedthrough RFI filters handle currents from 60mA to 15A. Two of these, A11X17 and A14X11, are primarily low-voltage types for telecomms applications up to 35GHz; the other two, designated A14X21 and A14X22, are mostly rated for mains use. Schaffner (EMC), 0734 697179.

Instrumentation

Smart oscilloscope. Philips' PM3070 Smart series oscilloscope is a 100MHz delayed-timebase instrument with automatic cursor-controlled measurement and an illuminated liquid-crystal display for readout of status and settings. Accelerating voltage is 16kV and an "autoset" facility selects the channel, sensitivity, timebase and triggering for any signal. IR Group, 0753 580000.

Track-current meter. This is an instrument for the measurement of current flowing in printed-board tracks, which do not need to be broken. Two probes inject a secondary direct current along the track, thereby causing a change in the potential gradient, which is measured and used to calculate the original current. The result is displayed digitally. Laplace Instruments, 061 440 9579.

Voltage and current calibrator. Model 521 from Electronic Development Corporation is a microprocessor-based, IEEE-488 (GP-IB)-controlled, voltage and current DC calibrator. It features current-mode outputs from 10nA to 110mA with compliance of 100V DC, and voltage outputs from 100nV to 100V DC and optional to 1100V DC. Compliance current is 100mA. The one-year voltage accuracy is ±0.002% of setting. Lyons Instruments, 0992 467161.

Temperature measurement. An optical-fibre temperature measurement and control system by Accufiber has a response time claimed to be 1000 times faster than an equivalent thermocouple system, with resolution of 0.01°C. The system is immune to EMI or RFI and can use contact or non-contact sensors. System configurations available include single and multi-channel versions with bidirectional RS-232 interfaces. Megatech, 0705 472868.

Portable DSO. The Tektronix 2211 portable oscilloscope offers both analogue and digital capability, with screen cursors, CRT readout and a hard-copy serial interface. It also provides 50MHz bandwidth, a 20M sample/s sampling rate, 8-bit vertical resolution and 4K record length. Tektronix UK, 06284 6000.

40MHz oscilloscope. The Trio-Kenwood CS5135 two-channel, four-trace oscilloscope has delayed timebase and allows examination of complex video signals using a clamp function. For complex signals, trigger hold-off is made variable. Thurlby Electronics, 0480 63570.

lGHz programmable attenuator. In the dB 125 attenuator, a total of 125dB may be switched in 1dB steps using a combination of 1,2,4,8,10 and 20dB stages. High-frequency relays direct the signal either to a pad or straight through. Accuracy at 500MHz is better than 0.1dB and input/output impedances are 50 ohms. VSWR is 1.15:1. Quartzlock, 080426 282.

Power supplies

Switched-mode regulator. The SPG dual-output switched-mode regulator from Schroff has a temperature-controlled current-limitation circuit which adjusts the maximum output current as a function of the ambient temperature. The unit is compatible with VME, Multibus II and STE systems and provides 110W at 50°C, with a soft-start feature. Schroff UK, 0442 40471.

DC/DC converters. The AA series of DC/DC converters are now available in triple-output versions. They operate from 28V or 48V inputs and are available with isolated output combinations of 5V, ±12V or ±15V. They can be paralleled to provide more current. The modules produce a maximum power output of 25W at up to 80% efficiency. Astec Europe, 0384 440044.

Power supply designers' kit. BICC-Citec's help kit is a means of enabling designers to produce a regulated power supply. The kit includes all the components necessary to produce the power supply: a PCB, the 3T regulator, input and output capacitors, inductor, a mosfet, schottky diode and heat sinks. It also includes a PC-compatible floppy-disc guide. BICC-CITEC, 0793 611666.

DC/DC converter. A 15W converter, known as Model 12S5.300XC, will accept any input voltage from 9V to 29V DC and provides an output of 5V at 3A to power digital circuit components. There is a minimum of eight hours short-circuit protection using pulse-by-pulse monitoring. The unit shuts down if the temperature exceeds specified limits. Calex Electronics, 0525 373178.

200W switching PSU. Astec's AS200 PSU, which operates from inputs of 80-135V AC or 180-270V AC, provides outputs of 5V, 12V, 24V or −5V/−12V. All outputs are overload protected and the unit is approved to international safety and RFI standards. Powerline Electronics, 0734 868567.

Production test equipment

PCB fault diagnosis. Designed to test digital devices with up to 40 pins, including surface-mount packages, the DDS-40 is a stand-alone system with a host of features for rapidly testing digital ICs in circuit. TTL, c-mos, LSI, static and dynamic memory and interface devices are among the test programs in the system library. ABI Electronics, 0226 350145.

Radio communications products

Low-noise microwave amplifiers. Two narrow-band, low-noise amplifiers from KDI/Triangle, the AN-72-1 and AN-72-5, provide a minimum gain of 10dB over the range 7.25 to 7.75GHz. Noise figure is better than 1.4dB at 25°C. Maximum input VSWR is 1.25:1 and output VSWR 2:1 for the AN-72-1, which gives a power output figure of 3dBm for 1dBm compression. Anglia Microwaves, 0227 630000.

11-14GHz orthomode transducers. ERA's range of OMTs enables a satellite antenna to operate with both vertical and horizontal polarization channels, either dual-receive or transmit-receive. Up to 40dB isolation between channels ensures interference-free operation and insertion loss is only 0.15dB. Other bands are in the 3-40GHz range. ERA Technology, 0372 374151.

EMC receiver. The Schwarzbeck FMLK 1518C is a precision radio receiver suitable for measuring levels of RF interference as laid down in CISPR, VDE and other standards. Tunable manually or digitally from 9kHz to 30MHz, the receiver will perform peak and quasi-peak level detection as well as average level. A direct output for a pen recorder is fitted. Schaffner EMC, 0734 697179.

Switches and relays

High-speed video switch. The HI-222 monolithic dual SPST switch has a bandwidth of over 200MHz with an on impedance of 35Ω and switching speeds of 100ns. Its true T-switch design provides high-frequency off isolation and crosstalk protection. Harris Semiconductor, 0276 685911.

Transducers and sensors

Digital input devices. A low-cost digital contacting encoder – the DP16 – is suitable for applications where microprocessors are built into equipment, such as compact disc machines, cameras, test and measuring instruments and cars. It has 360° continuous rotation. BICC-CITEC Ltd, 0793 611666.

Six photo-electric sensors. The MTE RK10 range of modular photoheads has been extended to include solid-state output for load switching up to 75V AC/DC. Light-on/dark-on mode selection is also provided, plus an optional timing facility. Thirteen types are now available to cover retro, object-sensing and through-beam applications up to 12m. MTE, 0702 421124.

COMPUTER

Computer board level products

Single-board computer for VME bus. The HK68/V2E 32-bit single-board computer from Heurikon is based on the Motorola 68020 CPU. It features 4 or 16Mbyte of on-board DRAM, up to 2Mbyte of eprom, and an extensive range of I/O and interprocessor communications facilities. Software support includes VxWorks from Wind River System, a real-time Unix-compatible operating system. GMT Electronic Systems, 0372 373603.

Data communications products

RS-232 to RS-422/485 converter. This is a general-purpose serial communications converter which provides signal conversion and transmitter controls to enable equipment with an RS-422 or RS-485 standard communications interface to be connected to RS-232 ports. Klippon, 0732 460066.

Mass storage devices

Rewriteable optical disc system. The rack-mount Ricoh RO-5030 can store more than 650Mbyte of data on an optical disc. Measuring only 5¾×8×3¼ inches, it features an ISO-approved standard 5¼ inch rewriteable optical-disc cartridge. Data Peripherals (UK), 0785 57050.

Programming hardware

Universal device programmer. The Stag System 3000 is capable of programming the whole spectrum of devices using manufacturers' approved algorithms and proven pin-driver technology on a single universal station. It incorporates comprehensive diagnostics. IR Group, 0753 580000.

PC-based eprom programmer. STRATOS is a programmer designed for use with any PC (IBM PC, XT, AT or compatible), and catering for all the most common eprom device types. Stag Electronic Designs, 0707 332148.

Software

Electronics design package. EE Designer IIIE is a new version of the EE Designer PC-based CAE package which incorporates a number of totally new features, including thermal analysis and an additional logic simulator. The thermal analysis feature in EE Designer IIIE calculates component and junction temperatures as well as a board surface-temperature gradient. Betronex UK, 0920 69131.

Temperature map of a PCB, produced by EE Designer III E software

JULY 1989

INDUSTRY INSIGHT

A glance at the future of microwave engineering • MVDS – multi-television takes the air • lowest manufacturing costs for a new type of flat-plate antenna? • designer's guide for microstrip patch antennas • advances in microwave spectrum analysis • report on a new Fresnel microwave antenna • progress survey – antennas combined with mmics • 10GHz GaAs op-amps made with microwave technology • James Clerk Maxwell telescope (above): a refined example of near-millimetre-wave engineering

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Microwave engineering – the future

Hugh McPherson of Heriot-Watt University briefly reviews the current state of the microwave industry and speculates about the future.

In terms of market areas for microwave technology, four main categories may be identified, as shown in the diagram. Traditionally, the microwave industry has always been driven by the demands of the military sector, which has constantly provided designers with the challenge of achieving higher frequencies, higher power levels, wider bandwidths, lower noise and more compact hardware – normally with cost as a secondary consideration.

In recent years, the driving force behind most activity in the military microwave field has been the ever-increasing capability of digital signal processing, and this is expected to continue in the future. As processing capacity continues to expand, more and more complex systems become feasible. Since this is not usually accompanied by a corresponding growth in the space available to house the equipment, miniaturization of the microwave circuitry becomes a crucial factor.

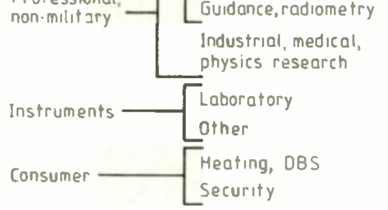
A good example of this increasing complexity and demand for miniaturization may be found in airborne radar where the goal is the development of an all solid-state, multi-mode pulsed-Doppler system employing an active phase-scanned flat aperture array antenna. To meet the expected demand for large volumes of transmit/receive modules for such systems, GaAs technology has had to advance toward the point of monolithic integration.

This is now leading to a new generation of digital hardware, which will allow even faster processors to be built, and will consequently place even greater demands on the microwave and other high speed analogue parts of the system. In radar, this new technology will open the door to further advances such as conformal antenna arrays and digital beam-forming, increasing further the need for high levels of integration.

Advances in signal processing techniques will also cause attention to become much more sharply focused on the spectral purity of the microwave and RF signals generated and handled by such systems. The two most critical items in this respect are the microwave signal source and the analogue-to-digital converter, where present technology often barely meets the needs even of current systems. It seems likely that further advances will depend very much on the appearance of devices employing new physical principles; for example, high-temperature superconductivity may bring sources of very much lower phase noise than we have at present, and some device using quantum effects may eventually be discovered for digital conversion, avoiding the need to carry out the process purely by circuit techniques.

Device technology

Much current activity is centred on research into transistors capable of operating at millimetre and sub-millimetre wavelengths. Devices under investigation at present include the high electron mobility transistor (HEMT), the permeable base transistor (PBT), the heterojunction bipolar transistor (HBT), and others. It is impossible to say which device will come out on top, but HEMTs are receiving the most publicity and there are devices available. When the most suitable technology has been established, and has matured to the point of monolithic integration, the electronically scanned, multi-beam, millimetre phased-array antenna may become a reality, enabling a completely new range of systems to emerge.



Main market areas for microwave technology.

Although great strides will undoubtedly continue to be made in solid-state device technology, the microwave tube seems likely to remain in existence for a very long time to come.

Commercial applications

So far we have been essentially concerned with defence applications of microwave technology. Recent restrictions on defence budgets together with peace initiatives have suggested that there is likely to be a levelling-off, if not an actual decrease in this area of application in the future. Many microwave companies, especially those involved in MMICs, are therefore anxiously seeking outlets for their capabilities in areas other than defence. In the consumer field, the size of the market is causing serious consideration to be given to products such as millimetre wave vehicle radars and electronically beam-steerable flat antennas for DBS.

One other potentially large market has not yet been exploited to any significant extent. This is the industrial sector, where microwaves could offer solutions to a wide range of measurement, monitoring, sensing, and heating problems. Millimetre-wave systems could be especially useful in many of these applications, having the advantages over optical methods of being able to operate in more hostile environments, and allowing Doppler information to be easily extracted. ■

Microwave television

Peter Hope of Marconi Communication Systems looks at the options for local broadcast services on microwave frequencies.

The idea that a large number (greater than 12) television channels can be transmitted directly to homes using frequencies higher than those currently in use in the UHF television band is not a new one. Microwave television systems have existed in the USA since the late 1950s.

Initially the service in the USA was restricted to one channel operating at 2.1GHz. This was called the Instructional Television Fixed Service or ITFS: it was used on university campuses to provide additional lecture material during the day and entertainment at night, and in city areas to relay educational programmes to local schools.

In the late 1970s, many of these systems incorporated transmission of a satellite channel: 'Home Box Office' used the system for distribution during the evenings when educational programmes were not being transmitted. From these beginnings the Multichannel Microwave Distribution Systems in the USA and Canada, authorized in 1984, came into being.

The transmitter for an MVDS system (**Fig. 1**) obtains its programming from a source which in this example is a satellite receiver. Signals from the programme source are fed to a modulator which processes the 1V pk-pk video and 600Ω audio signals into a PAL signal output at an intermediate frequency which is normally in the VHF band used by cable systems. Each channel to be transmitted uses a separate up-converter and amplifier. This is necessary because intermodulation between the individual programme channels would mean that very high power amplifiers operating at considerable back-off would be needed if all the channels were fed through one amplifier and up-converter.

Programme channels are then combined at the microwave frequency for transmission from a single antenna. The combining technique uses frequency-selective filtering. This method, although more expensive than couplers, has an overall loss for, say, a 12 channel system of typically less than 2dB, compared to 12dB for a network of 3dB couplers.

Most of the transmitter power is therefore available at the transmitting antenna – an important cost and performance consideration. A typical station broadcasting nine channels and giving combined effective isotropic radiated power (EIRP) in excess of 1kW will cost approximately £20 000 per channel and would cover a radius of approximately 30km.

At the receiving household (**Fig. 2**) a simple antenna and low-noise up-converter costing approximately £50 is all that is needed: the AM signal can be down-converted into existing television sets. If the programmes are encrypted, a standard cable-system set-top box can be used.

Microwave TV in the UK

A lobby for a system of microwave television in the UK was started in 1986 by a number of cable franchise holders and applicants. They saw the technology as a means of financing the expansion of cable networks by having a revenue-earning service up and running early in the life of the system.

As a result of this lobby the British Government commissioned the consultants Touche Ross to investigate the possibility of providing an MMDS (or Microwave Video Distribution Service, MVDS, as it has now been rechristened) in the UK. From the Touche Ross report the Government formulated ideas which are incorporated in the White Paper "Broadcasting in the '90s" currently before Parliament. The cable companies' wish to have MVDS distribution technology available to them is now reflected in the White Paper, which discusses franchising local delivery services which are 'technology neutral' after 1991.

By 'technology neutral' the Government means that a company franchised to deliver local television and telecommunications services will be able to use a mix of cable, microwave and any other authorized means of delivery available, rather than being restricted to the use of cable as is now the case.

The British Government, through the Touche Ross report, has considered three possible frequency bands for MVDS. These are 2.5GHz, using amplitude modulation, 12GHz using AM or FM and 40GHz using FM.

For an illustration, consider the performance of some practical systems:

Band	Mode	Power	Range
2.5GHz	AM	20W	30km
12GHz	AM	2W	2km
12GHz	FM	2W	25km
40GHz	FM	0.1W	2km

Tests carried out by Marconi at 2.5GHz and 12GHz suggest that these figures present a reasonable picture of the situation. However, the 40GHz predictions are based on the 29GHz trials conducted by British Telecom at Saxmundham.

Typical transmitter and receiver parameters are listed in Table 1. The carrier-to-noise (C/N) values stated are those which achieve a video signal-to-noise ratio of 40dB weighted, corresponding to a picture quality of 4.5 on the CCIR five-point scale. For simplicity, it is assumed here that this grade 4.5 picture should be available over the entire path with light rain (2.5mm/hour) between transmitter and receiver, and that the picture should remain with C/N above threshold during heavy rain. This is taken to be 25mm/hour over 1km of the path, with 2.5mm/hour over the remainder.

Table 1: MVDS link parameters

	2.5GHz AM	12GHz FM	38GHz FM
Power at Tx antenna, dBW	+9	0	−10
Tx antenna gain, dBi	15	15	15
Rx antenna gain dBi	20	27	27
Rx noise figure (dB)	3	2	12
Rx bandwidth (MHz)	5.5	33	33
Operating C/N at Rx (dB)	43	16	16
Carrier/intermod (dB)	53	26	26
Resulting permissible path loss (dB)	135	153	133
Path length (km)	30	25	2

Although MVDS on 2.5GHz has been rejected for the UK, such systems will soon be operating in Ireland and are already in use in the USA and Canada.

In Ireland it is planned to provide a nationwide service of 11 channels using the frequency band 2.5 – 2.65GHz. This is possible by spacing each channel of 8MHz bandwidth at 16MHz, and reusing frequencies on an 11-channel, 16MHz grid with 8MHz offset. With vertical and horizontal polarizations, this gives four alternative frequency plans (Table 2).

MVDS at 2.5GHz

Figures 3 and 4 show the basic equipment used in a 2.5GHz MVDS system. Programmes are obtained either locally or from satellite and fed to a PAL modulator in the same way as an existing UHF broadcast transmitter operates. The IF signal at VHF (100MHz) is up-converted to the microwave band, amplified to a suitable power and fed to

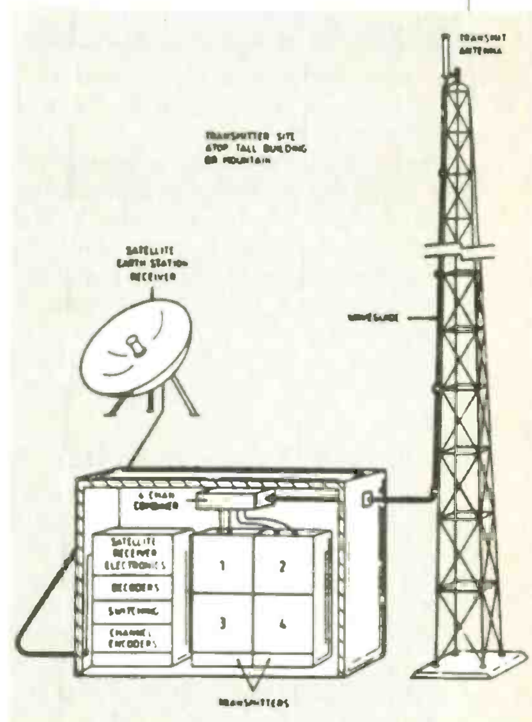


Fig. 1. Transmitting station of a satellite-fed multi-channel microwave video distribution system.

Fig. 2. Multi-channel video distribution system.

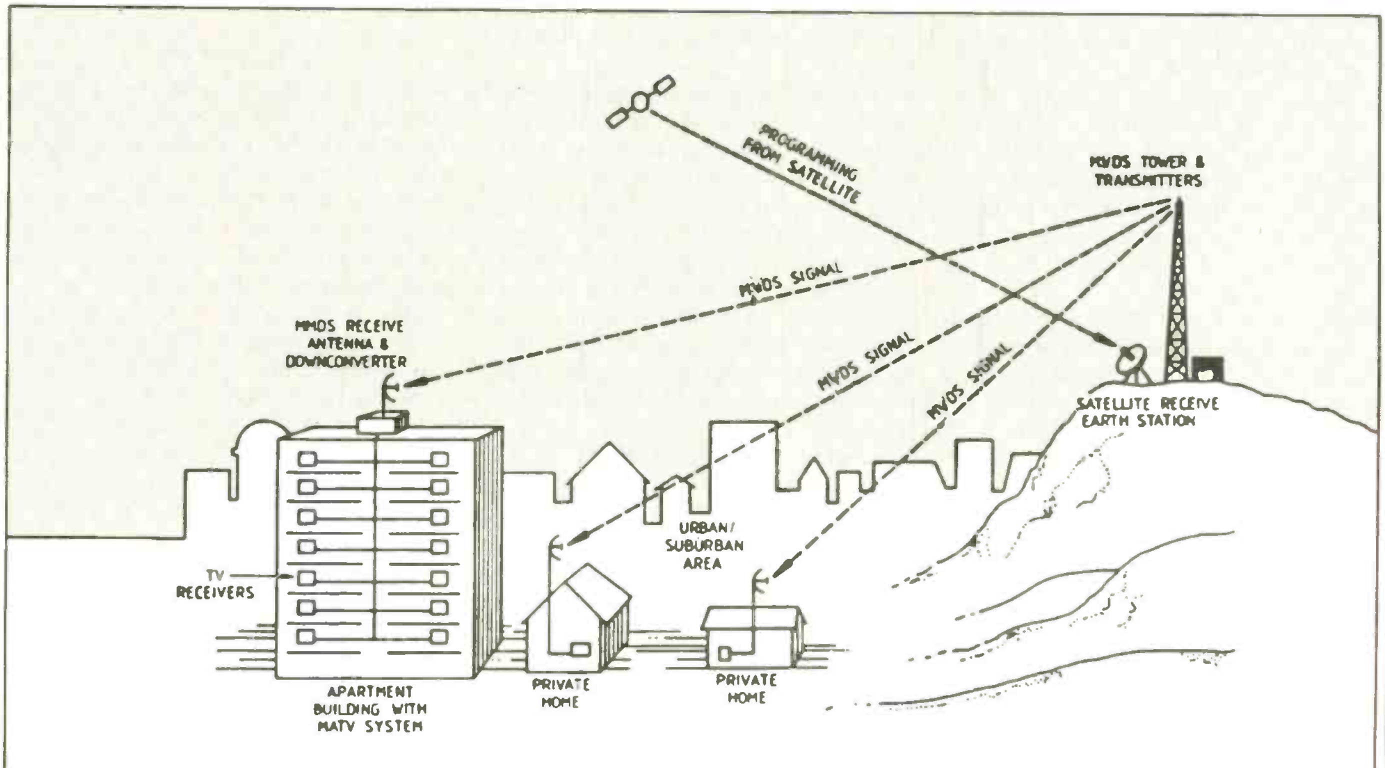


Table 2: Ireland's MVDS frequency plan

Channel number	Vision carrier (MHz)	Sound carrier (MHz)
1 A	2501.25	2507.25
2 B	2509.25	2515.25
3 A	2517.25	2523.25
....	......	......
22 B	2669.25	2675.25

an omni-directional or cardioid-patterned antenna to provide the required coverage.

In Ireland the systems are expected to be encrypted so that revenue can be collected from the users, although it is likely that a minimum package of five or six channels (comprising the four UK networks plus a new privately-owned Irish channel) will be provided at low cost to the household. Premium satellite channels would be charged at higher rates.

MVDS at 12.1 – 12.5GHz

In the UK the Home Office has recently announced that only two frequency bands are now under consideration for MVDS: 12GHz and 40GHz. A recent DTI statement called for interested organizations to make submissions on the viability of these two bands.

The reason for the interest and uncertainty over the 12GHz band is that currently this band is allocated to direct broadcasts by satellite (DBS). If part of the band were to be committed to MVDS then it would not be used for DBS in the future.

The main attraction of 12GHz is that the low-cost receiver technology developed for DBS is fully compatible with an MVDS delivery service. Furthermore, the advances towards high-definition television (HDTV) using the MAC system demonstrated at IBC at Brighton in 1988 are continuing to be developed through the European-funded Eureka programme, and are compatible with this method of programme delivery.

There appears to be a good technical case for allowing MVDS a chance to prove itself and this could be done immediately by establishing some large-scale system trials in this band. Since the technology for 12GHz MVDS is identical to that of a satellite TV system (but with the signal being broadcast terrestially) the transmitter equipment could be based on that used to transmit signals to the satellite (**Fig.5**). It would cost approximately £25 000 per channel, including the cost of combining.

A typical 12GHz single-channel transmitter comprises an FM modulator and double up-converter giving an output of between 1 and 10 watts depending on the power requirements of the system. Shown in **Fig. 6** is a multichannel network used to combine the signals on to a single omni-directional antenna.

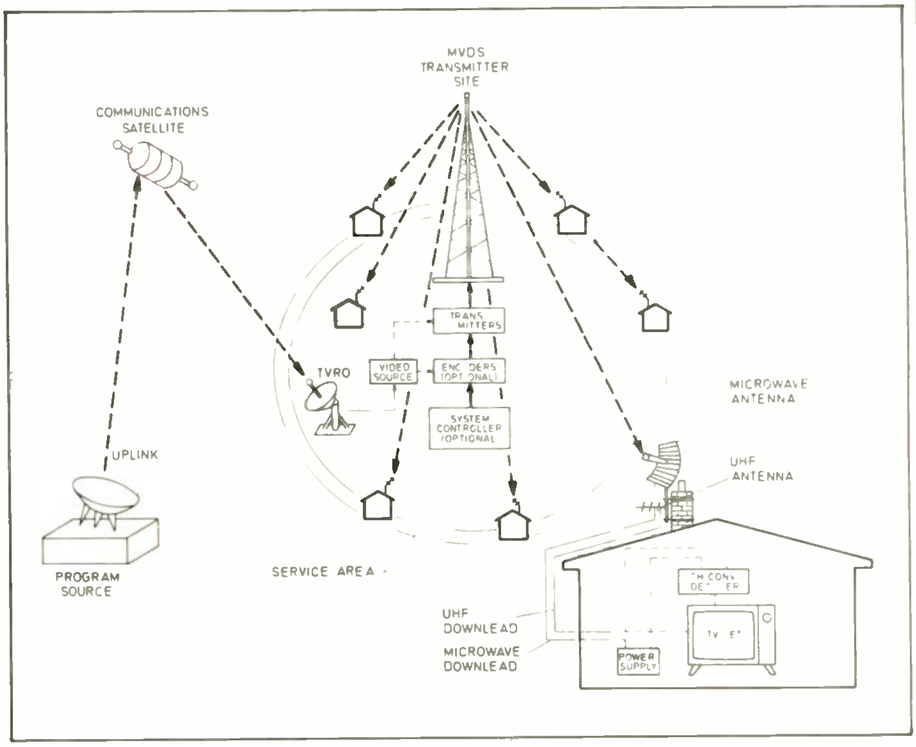


Fig.3. Complete microwave video distribution system.

MVDS at 40GHz

Millimetric MVDS has been on trial by British Telecom's Research Laboratories. The advantages of a millimetric system are

- the availability of spectrum means that at least 30 channels are achievable, with no disruption to planned services;
- this frequency may suit cable system operators who will need to use MVDS largely as an extension of their systems into the fringe areas of a town or city, or small market towns which are uneconomic to cable. This also overcomes the objections made against MVDS that, once in use, it will discourage the cable franchise holder from installing a cable system.

However, the limitations of millimetric systems should not be overlooked.

1. Range is short, 3km at best.
2. Technology is currently expensive. It is not certain when the technology will mature enough to make the domestic RF components available cheaply to the consumer, although one UK research organization has made optimistic statements made in this respect.
3. Penetration within a given coverage area will be less than that of the lower frequency options.

Equipment for millimetric systems using FM would be much the same as that for 12GHz systems. Apart from the domestic low-noise converter, the low-cost indoor satellite receiver units now on sale in the high street could be used. Several manufacturing organizations are fully confident that a 40GHz low-noise converter can be produced at a similar price to 12GHz DBS LNCs.

What next?

With a DBS system already established in the UK, local delivery systems including cable and MVDS will have to offer more to the customer than satellite alone can provide. These services must therefore include local low-budget programming and minority interest programming on subscription.

The choice in the UK is between 12GHz with available DBS technology and longer range but with a limited frequency allocation, and 40GHz. The latter, however, is, as yet unproven and will possibly be more expensive and with a shorter range. It has, however, potentially a larger frequency allocation.

The success of local radio should be a good indication that many people prefer to listen to local-interest programmes, news and current affairs. If economic programming can be established and market demand is sufficient to drive down the cost of equipment, local television will find a place. ■

Flat plate satellite antenna

Anthony E. Sale describes a multiple flat plate reflector design suitable for microwave television reception

The simplest form of satellite antenna system is a concave reflector, a dish, with a small horn aerial, the feed horn, located at the reflector's focus. If such a system is properly designed and constructed, about 50% of the intercepted signal energy is delivered to the receiver. The intercepted energy is equal to the area of the aperture of the reflector multiplied by the signal strength at the aperture of the reflector.

Slightly higher efficiencies can be obtained by shaping the reflecting surface so that the feed horn can be offset from the line-of-sight to the satellite and still be at the focus. Efficiency is increased mainly because the feed horn and its support no longer cast a shadow on the dish.

For both the axial and offset feed horn positions, the concave surface

View of the flat plate antenna. Outer plates can be set back in increments of half a wavelength, to reduce the physical depth of the reflector.

Signals from Radio-Télé Luxembourg and SAT-1, received from Astra and Eutelsat 1-F1 by the flat-plate antenna.

must be figured to a high degree of accuracy: ±0.5mm is often quoted. This accuracy must be maintained under wind loading, and the feed horn position must be held precisely.

To receive signals from different satellites, the complete reflector and feed horn must be rotated. Because of the large aperture, the beam angle is small – 2° is typical. This means that any positioning system must have a repeatability of less than half a degree and be sufficiently robust to withstand wind.

Antenna systems for multi-satellite reception are therefore expensive, rather unreliable unless heavily engineered, difficult to install, and very bulky and obtrusive.

There is however, an alternative approach. This reflector design described here allows cheap and easy production through greatly reduced tolerance requirements, easier selection of different satellites by moving the feed horn while keeping the multiple plate reflector fixed, and a far less obtrusive installation by virtue of the lesser depth of its reflecting surface. These improvements are achieved without loss in efficiency by constructing the reflecting surface from a number of flat plates.

The design relies on the focusing effect of the diffraction pattern resulting from reflecting a plane wave from a flat surface which is a small number of wavelengths across. This diffraction pattern is the same as that obtained by illuminating a slit by a plane wave and observing the field strength at different positions on the far side of the slit. The field intensity patterns differ according to the slit width, the distance and the wavelength (see panel). The far field, at distances greater than $2a^2/\lambda$, approximates to the $(\sin x)/x$ curve. Towards the end of the collimated region at a distance of $a^2/(2\lambda)$, the intensity of the axis is greater than the intensity without the slit. Close to the slit at $a^2/(30\lambda)$, the intensity approximates to a rectangular function equal to the incident plane wave intensity.

In the flat plate design, the width of the flat plates and the focal length, i.e. the distance from the plate to the feed horn, are chosen so that the diffraction effect causes a maximum signal to be delivered to the feed horn. The feed horn aperture must be large enough to capture all the energy in the diffraction

Reflections from a flat plate

This analysis deals with reflections from a flat plate illuminated by a distant radiating source. Since the source distance is very large compared to the plate dimensions, the illumination can be considered as a plane wave of uniform phase across the wave front.

A flat square plate can first be analysed by considering only two dimensions, i.e. treating the plate as a reflecting strip of infinite length and of width equal to that of the plate. This is mathematically precisely the same as an infinite slit in an infinite conducting plane, illuminated from behind. This well-known problem was first investigated and solved by Fresnel using Huygens' principle, which treats a wavefront as an infinite number of radiating sources, each with a $(1+\cos\alpha)$ intensity distribution.

If the z axis is normal to the centre of the slit of width a, and x is the distance off of the z axis to the point P (Fig.1), then the resultant intensity at P is obtained by integrating the vector contributions from each point in the slit. This involves the Fresnel integrals whose values are tabulated in terms of a normalized parameter, v. These values can also be presented graphically as Cornu's spiral, shown in Fig.2. The parameter v is evaluated as

$$v = x\sqrt{2/(z\times\lambda)} \qquad (1)$$

when λ is the wavelength of the illuminating radiation, and is marked on the Cornu spiral. This normalized form of the spiral makes the distance between the two infinity points equal to the free-field intensity, i.e. without the slit present. If just one edge of the slit is considered, with the other edge moved to infinity, then for different distances for the point P off the z axis, the resultant intensity is obtained from the Cornu spiral as the distance from one infinity point to the v point in the spiral corresponding to (1).

For a given distance on the z axis, this looks like Fig.3. If both sides of the slit are now included, then for a given point P two values of v are required – one relative to each side of the slit. The resultant intensity is then the distance between these two v positions on the spiral. This is obviously at a maximum for both v values equal to 1.25 and this corresponds to P being on the z axis and at a distance

$$z = a\times a/(2\times\lambda\times v\times v)$$

from the slit. The intensity at this point is nearly 3dB greater than the intensity without the slit. If circular symmetry is used this also corresponds to the focusing effect, due to diffraction, of the first Fresnel zone.

The shape of the variation in intensity as P moves off the axis is also important since this determines the aperture of the receiving horn aerial. This is shown in Fig.4 for various distances away from a slit. Close to the slit at about 5λ, the intensity is nearly rectangular and matches the slit dimensions. In the Fresnel focused region, the curve has a pronounced central hump, with most of the energy being within 70% of the slit width. At distances greater than the Rayleigh limit, $2a\times a/\lambda$, the curve approximates to $\sin(x)/x$.

For a plate width of 20cm and λ=2.64cm, maximum focusing occurs at 50cm, but is in fact only slightly less at 40 and 70cm.

Shapes of the diffraction patterns can be investigated with a computer program performing the integration. The inner loop calculates the phase and amplitude at P for each element in the slit aperture. The cosine and sine resolved amplitudes are then summed to give the resultant phase and amplitude at P. Two outer loops then vary the P offset from the z axis and the distance from P to the slit.

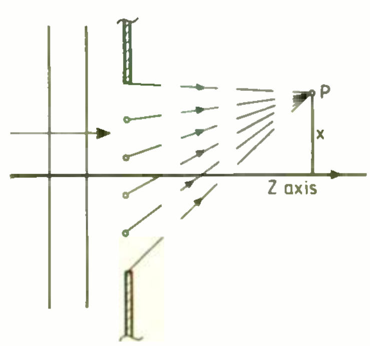


Fig.1. Intensity at P is obtained by integrating the vector contributions from each point in the slit.

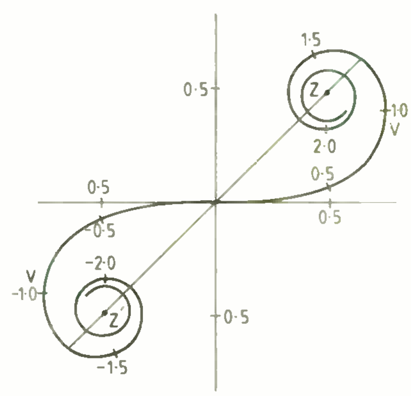


Fig.2. Fresnel integrals plotted as Cornu's spiral.

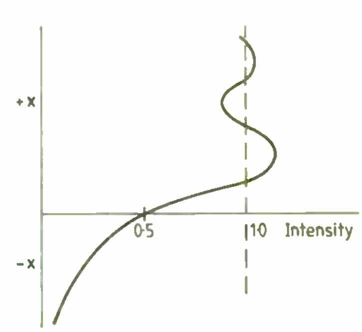


Fig.3. Resultant intensity, obtained from the Cornu spiral.

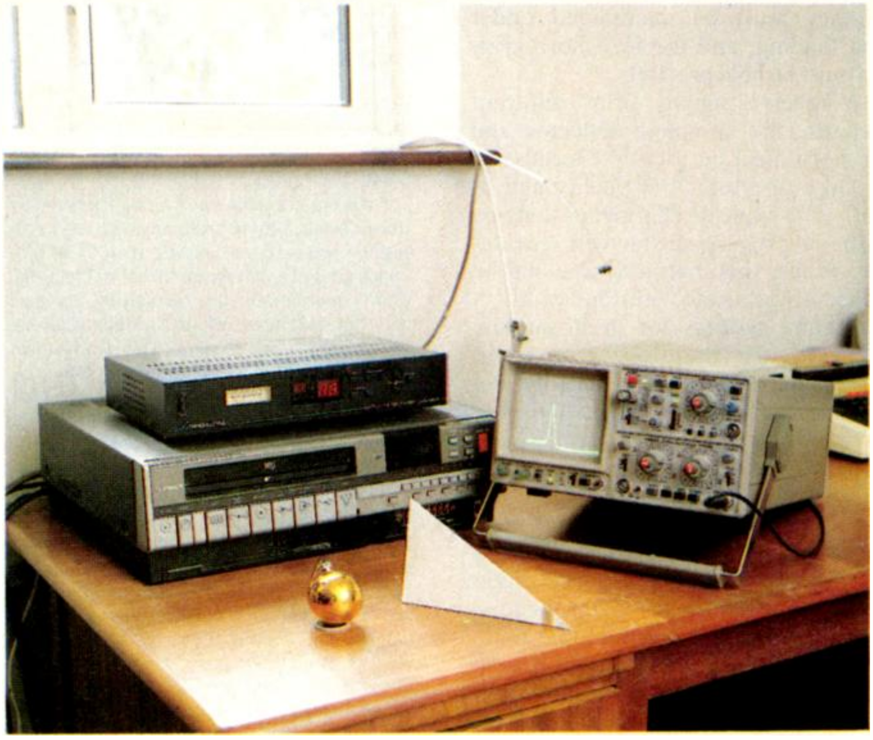
The reflecting antenna

The reflector consists of a matrix of flat metal plates (picture, page 784), each angled to reflect the signal which it intercepts into the feed horn at the focus. It may be made shallower by setting plates back by a half-wavelength, about half an inch, whenever possible. This is known as zoning.

Signals from the Astra satellite are on frequencies in the region of 11.5GHz, corresponding to a wavelength of about 2.6cm. Since the penetration depth of such high frequencies is very small in aluminium, quite thin foil can be used as the reflecting surface. Aluminium kitchen foil is completely adequate: it can be glued to squares of hardboard or plywood for support. The feed horn should have an aperture roughly equal to the size of a plate and a flare angle such that it "sees" the whole set of plates.

Because the satellite signal is weak, it is not practicable to use it for aligning the reflecting plates directly. An alternative, optical method is to place the multi-plate reflector at one end of a long, darkened room and to hang a small spherical reflector (such as a Christmas tree glass ball or a well-polished soup spoon) at its intended focus. Using a distant point-source of light positioned such that the sphere's shadow falls on the central plate, adjust each plate in turn so that its image appears, evenly illuminated, in the sphere.

However, if a sensitive swept frequency display is available, it is possible to align the plates against a signal from one of the high power satellites. With a horn aperture of 15cm and a flare angle of 60°, a direct signal can just be seen from Astra. It is also possible to see the reflected signal from a single plate 20cm square.



pattern. In practice, an aperture width approximately equal to the width of a plate is adequate.

The photograph of the multiple plate antenna system shows how some plates are set back, away from the feed horn, to reduce the depth of the aerial away from the supporting structure. The reflecting surface is mounted rigidly against a suitable vertical surface, such as the wall of a building. In the northern hemisphere, this wall should face the south, since this is where most of the satellites are situated in the sky. The feed horn is then moved across the front of the multiple plate aerial in order to receive signals from different satellites. An important characteristic of this reflector is that it maintains its efficiency for wide angles (±30°) off the normal to the reflector. This contrasts sharply with a normal, fully figured parabola, whose efficiency falls off quite rapidly as the feed horn is moved away from the design position. This improved off-axis performance results from two factors: firstly the flat plates themselves cause degradation only in proportion to their projected area, i.e. as the cosine of the angle off the normal; secondly, the setting back of plates gives a flatter reflecting surface which causes far less variation in phase for off-axis working.

To receive different satellites, the feed horn has to move in front of the multiple flat plate reflecting surface. Its path must pass through the images of the satellites, tracing a reflection of the geostationary orbit path. One simple way of constraining the horn to move on this image path is to suspend it on a single cord passing over pulleys mounted one at each top corner of the reflector, with a compression strut pivoted at the bottom centre of the reflector. By choice of the exact positions of the pulleys, the length of the cord and the length of the strut, the horn movement can be made to match the image path closely.

A reflecting surface of 16 plates, each 20cm square, appears to perform as well as a good quality 80cm dish. With 25 plates, each 20cm square, it is about the same as a 1.2m dish. In both cases, the multiple plate aerial was working at 30° offset in the vertical plane and showed less than 2dB degradation at 20° off the normal in the horizontal plane. ■

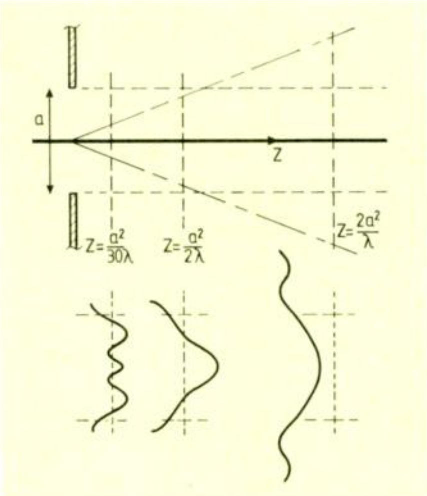


Fig. 4. Variation in intensity as P moves off the axis: in the Fresnel-focused region (centre), most of the energy is within 70% of the slit width.

Danger from sunlight

The flat plate antenna is a very good optical reflector, by virtue of its design. When mounted outdoors it focuses the Sun's rays and produces a very hot spot at its focus. Do not allow pets or children to get near the focus. Once the reflector is in position it should be covered in opaque plastics sheet or painted with a suitable non-reflecting paint.

The author has filed an application for a patent on the flat plate antenna design (A 8908759, April 18, 1989).

Further reading
Diffraction theory and antennas, by R. H. Clarke and John Brown, Ellis Horwood, 1980.
Radio wave propagation and antennas, by John Griffiths, Prentice Hall, 1987.
Antennas, Kraus, McGraw-Hill, 1950.

Designing patch antennas

Tim Forrester presents a designer's guide to one of the basic types of MMIC-compatible antenna.

Although the phenomenon of microstrip radiation has always been present in conventional microstrip transmission line design it was long accepted as almost an unavoidable loss in the system.

Most radiation from a microstrip line occurs where there is a discontinuity in the line, that is to say a mismatch of one sort or another. By enhancing this radiation it is possible to design a compact aerial with reasonable efficiency and a low physical profile.

Although this technique is suitable mainly for frequencies above 1GHz, it can be used down to 200MHz or even lower; however, the surface area tends then to become unwieldy.

A patch aerial can take many forms, from a simple rectangle **(Fig.1)** to a complex multi-patch array **(Fig.2)**. The more complex the aerial, the greater the design problems; but it is relatively simple to design a single patch aerial which will have a predictable performance.

To mass-produce a simple patch aerial is relatively simple once the design has been proved; however, there are pitfalls which can cause problems even with a simple single-patch design.

Problems with patches

Perhaps the easiest method of analysing the simple patch antenna of Fig.1 is to consider it as either a radiating slot or cavity. To analyse a patch aerial thoroughly involves using a full modal expansion technique, which can handle any arbitrary shape of patch aerial. This technique involves very difficult theory and extremely complex calculations. For this reason this article is concerned only with the design of simple patch aerials.

The type of dielectric used has a dramatic effect on the size and radiation properties of any patch aerial. Therefore the consistency of any dielectric material is of critical importance in any design which is going to be duplicated.

Also the actual choice of dielectric material will be compromise, since the requirements conflict **(Table 1)**. For

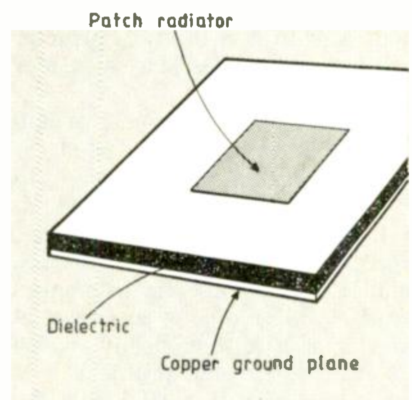


Fig.1. Simple patch antenna. Note that patches do not have to be rectangular.

Fig.2. Complex patch array.

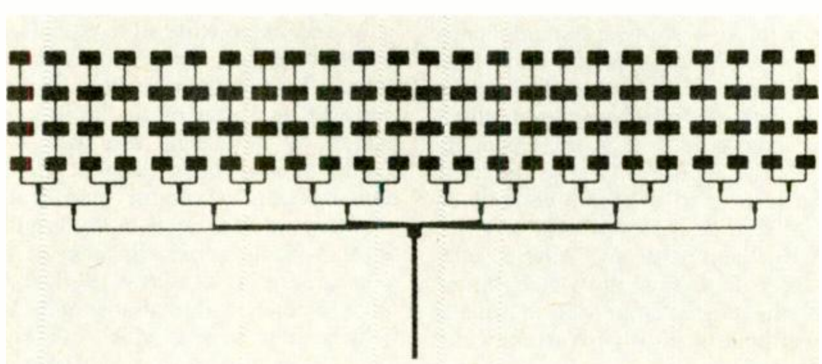


Table 1. Substrate choice for patch arrays.

Requirement	Dielectric constant	Substrate thickness
Low feed radiation	high	thin
Low surface waves	low	thin
Good tolerance control	low	thin
Low mutual coupling	low	thin
Low array losses	high	thin
Wide bandwidth	low	thick

example, a thick material of low relative permittivity gives a wider bandwidth but a lower efficiency; a thin high-dielectric material will produce a more efficient patch, but a narrower bandwidth.

Another important point to be considered is the method of feeding the patch. Since a microstrip transmission line radiates when the signal encounters a mismatch, if a particular radiation pattern is required then radiation from the feed line must be minimized.

The simplest solution is to feed the patch from behind, as shown in **Fig.3**.

The feeder transmission line could be conventional coaxial feeder, or in the form of a microstrip transmission line. There are other methods of feeding a patch from the rear besides making a direct electrical connection, such as having a small iris in the back plane to permit the patch to couple to the feeder network.

With all compact antennas, overall bandwidth is inevitably reduced. A typical single patch aerial is likely to have a bandwidth of only 1-3% at 10GHz (even less at lower frequencies), compared with 15% for a free space dipole. However there are techniques to broaden the frequency response, at the cost of increased complexity – which may in itself lead to other problems.

A factor limiting the patch bandwidth is the frequency range over which the feed impedance is within defined limits. One method of increasing the effective bandwidth is to use a technique known as feed inductance compensation. Since the patch is fed from the rear, there is a certain amount of unwanted inductance between the patch and the end of the coaxial feeder. By simply incorporating a series capacitor this unwanted inductance can be tuned out, so increasing the bandwidth. **Figure 4** shows the most common method. Alternatively, by using a dielectric material with low permittivity, bandwidth can be increased at the cost of making the patch larger. Feed inductance compensation can also be used simultaneously to further widen the bandwidth.

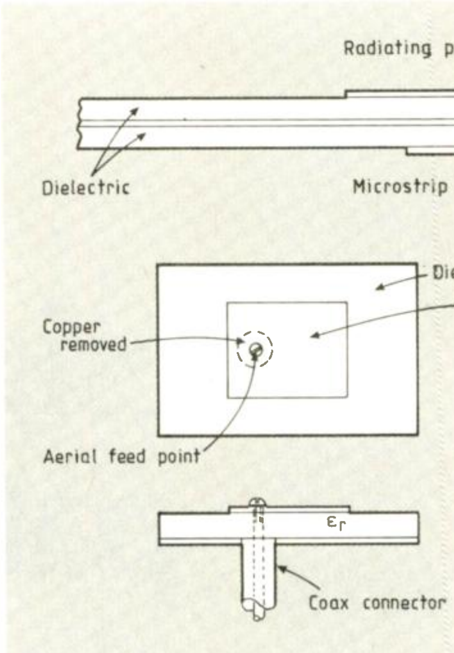


Fig.3. Microstrip antenna fed by coaxial transmission line.

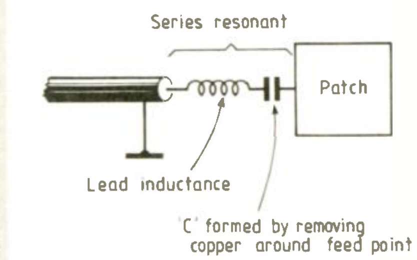


Fig.4. Compensating for feed inductance. The value of C must be chosen for resonance with the intrinsic lead inductance.

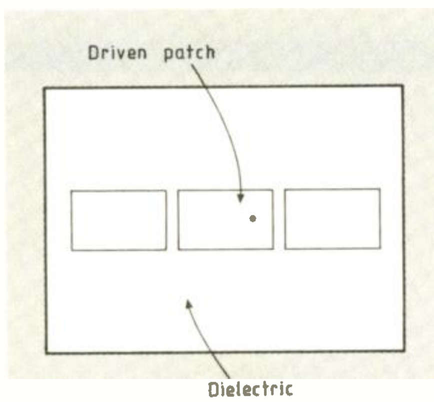
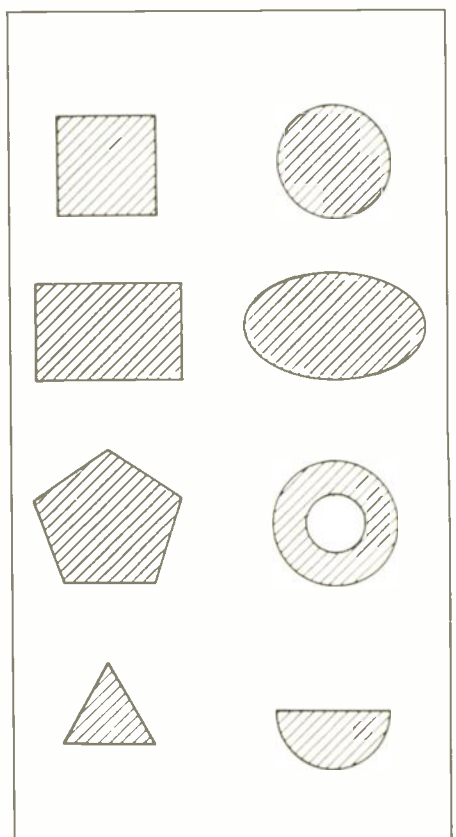


Fig.5 (left). Multiple patch, parasitically coupled.

Fig.6. Some practical microstrip antenna configurations.



A more complex method of increasing overall bandwidth is to use a number of patches, each resonant at a slightly different frequency and parasitically coupled to the next adjacent patch (**Fig.5**). A further benefit of this technique is increased aerial gain; the drawback is that the design is more much involved.

As the complexity of a patch aerial array increases, losses inevitably increase. Unfortunately there comes a point where little or no advantage is gained by increasing the size of the array. It has been shown that the approximate maximum gain from a complex patch array is in the region of 35dB, with an aerial efficiency of 10% when operating at about 10GHz. This low efficiency is due almost entirely to resistive line losses, which can be reduced by using a low-loss feed network.

This inherent loss in the aerial system also means that the overall noise figure of a receiving system is degraded. It is difficult to quantify the degradation because the loss is dispersed throughout the receive aerial system – unlike a conventional system where an aerial has a known gain, followed by a certain feeder loss into a receiver of known overall noise figure.

Design of a rectangular patch

For a given frequency there is a family of patch sizes which will be resonant at that frequency. There is, however, an optimum size which will exhibit a good radiation pattern and reasonable efficiency. The actual size of the patch will depend on a number of factors, such as the relative permittivity of the dielectric and its thickness, the shape of the patch (usually round or rectangular), and the overall bandwidth required.

The required bandwidth largely determines the thickness of the dielectric, but because the dielectric is usually available only in certain thicknesses, then the width of the patch may have to be varied from the optimum if a particular bandwidth is required.

There are other much more complex array shapes besides the ones mentioned above, each with its own properties and benefits **(Fig.6)**. For ease of design and construction a rectangular patch is used in this example. The formulae used to calculate the patch dimensions will give a good first order approximation, but if a more exact frequency is required with a particular bandwidth this method may not prove accurate enough. It is possible either to trim the patch aerial with a sharp knife on to the desired frequency, or else to use a set of formulae with built in correction factors to calculate the theoretical size more accurately. ■

Further reading

This article has barely scratched the surface of patch aerial design: it has glossed over many other interesting aspects of these novel aerials, such circular polarization and phased steerable arrays. The following list is just a selection of the literature available:

Microstrip Antennas by I. J. Bahl and P. Bhartia. ISBN:0-89006-098-3

Microstrip Antennas, Design Equations, IEEE AP-S International Symposium Digest June 1979, page 122 to 125.

Theory and applications of broadband microstrip antennas, 6th European Microwave conference, September 1976, pages 275 to 279.

DESIGN EQUATIONS

To design a patch of a given frequency, it is first necessary to choose a suitable dielectric substrate material, bearing in mind that any choice is likely to be a compromise of efficiency, bandwidth or patch size etc. If there is a minimum bandwidth requirement, the thickness of the dielectric substrate can be determined from the following formula:

$$t=\frac{BW}{128\times F_0^2}$$

where F_0 is the centre frequency in GHz, BW is the bandwidth in MHz and t is the minimum substrate thickness in inches.

Choose a dielectric thickness which is suitable and readily available. This usually means alumina ($\varepsilon_r=9.8$), duroid ($\varepsilon_r=2.32$), low-loss foam ($\varepsilon_r\simeq 1.04$), or some sort of glass fibre, whose permittivity could range anywhere from 2 to 5. Low-loss foam can be purchased from wallpaper stores under the brand name "Rosslite". Unfortunately its dimensional stability leaves a little to be desired, but at 5p per square foot it is very cheap! Glass-fibre PCB material can be used at the lower frequencies; it helps considerably if its relative permittivity is accurately known.

Having chosen a suitable dielectric substrate of a certain thickness, we can determine the optimum patch width. By using a patch of a calculated optimum size on a standard substrate, it is possible to make certain assumptions as to the feed impedance at the edge of the patch, and so where the 50Ω feed point will be within the area of the patch. To calculate the optimum patch size, use the following formula:

$$W=\frac{c}{2f_r}\left(\frac{\varepsilon_r+1}{2}\right)^{-1/2}$$

where W is the width of the patch (cm), ε_r is the relative permittivity of the dielectric substrate, c is the velocity of light (3×10^8m/s) and f_r is the desired resonant frequency (GHz).

Length of the patch in centimetres is given by

$$\frac{c}{2f_r\sqrt{\varepsilon_e}}-2\Delta l$$

where ε_e is the effective permittivity (see formula below) and l is the line extension caused by fringing (see formula below).

Effective permittivity of the dielectric ε_e is given by

$$\frac{\varepsilon_r+1}{2}+\frac{\varepsilon_r-1}{2}\left(1+\frac{12h}{W}\right)^{-1/2}$$

where h is the thickness of dielectric, W is the width of the track (patch) and ε_r is the relative permittivity of the dielectric.

Line extension is

$$\frac{\Delta l}{h}=0.412\frac{(\varepsilon_e+0.3)\ (W/h+0.264)}{(\varepsilon_e-0.258)\ (W/h+0.8)}$$

This procedure provides a method of predicting patch resonant frequencies to within 10% or so. If the patch has been designed to the optimum size and based on one of the three common dielectric substrates, then the feed impedance at the edge of the patch will be very close to one of the following figures.

For dielectrics of relative permittivity of 2.5 or less, the feed impedance will be in the region of 240 – 260Ω. For a high relative permittivity material, such as alumina, the feed impedance will be much higher, in the region of 550Ω. Exact values can be calculated from the formulae shown below, but the impedances mentioned above are adequate for experimental purposes.

Self conductance,

$$G_{11}=\sqrt{\frac{\varepsilon}{\mu}}.\frac{1}{\pi}\int_0^\pi \sin\left(\pi.W\frac{\cos\theta}{\lambda}\right)^2 J_0\left(x.2.\pi\frac{\sin\theta}{\lambda}\right).\cos\left(z.2.\pi\frac{\cos\theta}{\lambda}\right)$$

where $\mu=4\pi\times10^{-7}$ and $\varepsilon = 1/36.\pi\times10^{9}$.

$$G_{12}=\sqrt{\frac{\varepsilon}{\mu}}.\frac{1}{\pi}\int_0^\pi \sin\left(\pi.W\frac{\cos\theta}{\lambda}\right)^2\frac{\sin(\theta)^3}{\cos(\theta)}.J_0\left(x.2.\pi\frac{\sin\theta}{\lambda}\right).\cos\left(z.2.\pi\frac{\cos\theta}{\lambda}\right)$$

Then,
$$Z_{in}=\frac{1}{2(G_{11}+G_{12})}$$

Having determined the impedance at the edge of the patch, it is now fairly easy to design an impedance-matching quarter-wave transformer to feed the edge directly, or alternatively to determine the 50Ω feed point within the area of the patch.

To determine the impedance of a quarter-wave matching transformer, use the simple formula:

$$Z_t=\sqrt{Z_{in}\times50}$$

where Z_t is the impedance of the quarter-wave transformer and Z_{in} is the edge input impedance of the patch.

To determine where the 50Ω feed point is within a patch, use

$$l=\frac{\mathrm{acos}\left(\sqrt{\frac{50}{Z_{in}}}\right)}{\beta}$$

where l is the distance in wavelengths from the edge of the patch and β is defined as $2\pi\sqrt{\varepsilon_r}$.

Care must be taken to ensure that the patch is made as close to size as possible, and that the dielectric used is of a known relative permittivity, otherwise inevitably the patch will not perform as expected.

A good method of prototyping these patch aerials is to use sticky-backed copper tape (available from RS Components), which can be trimmed easily on the dielectric subsrate until the desired parameters are achieved.

Waves apart

Andrew Standen discusses Anritsu's advances in RF and microwave spectrum analysis.

Microwave spectrum analysers are indispensable in the research and development, design, manufacturing, installation and maintenance of RF and ultra-high-speed logic systems. But while high-specification, multi-function spectrum analysers have been too expensive, cheaper types have often lacked performance. Obviously, there is a demand for general purpose with a high technical specification and numerous functions at a modest cost.

If you want to provide a measuring instrument which can be operated by a wide range of users, then these criteria can be summarized as follows:

1. Achieving fundamental performance requirements.
2. Ensuring that the instrument is easy to use and understand.
3. Ensuring applicability to automatic measuring systems.
4. Compactness and low cost.

The key to meeting all these requirements is the development of high-grade microwave devices and components, such as filters, attenuators, mixers, oscillators, etc., and of the technology to use these components. However, if you were to develop the highest possible grade components and implement these in our design, then this would conflict with point 4.

In a normal RF spectrum analyser using the fundamental frequency component only of the local oscillator signal, the first IF is generally chosen to be higher than the input tuning range to prevent any direct feed of the input signal through the IF filter without being transformed by the IF stage.

Let us assume we are designing a spectrum analyser with an input tuning range of 0Hz to 2GHz, and an IF at 2.5GHz (good practice, since this is outside the input frequency range). From the tuning equation,

$$F_{IF} = F_{local} \pm F_{input}$$

the local signal can be swept from 0.5GHz to 2.5GHz or 2.5GHz to 4.5GHz to produce an IF signal for any input signal. In the first case we would need to produce an oscillator capable of sweeping over a relative frequency range of five times; whereas in the second, the relative frequency sweep is less than an octave. Obviously, it is much easier to produce an oscillator with a narrower sweep range; so, for our simple example, we would choose the local sweep range of 2.5GHz to 4.5GHz. However, we have one last problem to contend with – that is, to ensure compactness and low cost, analogue hardware should be eliminated from the analyser as far as possible. This has the added advantage that digital techniques lend themselves much more easily to automatic control, through a GPIB interface, for example.

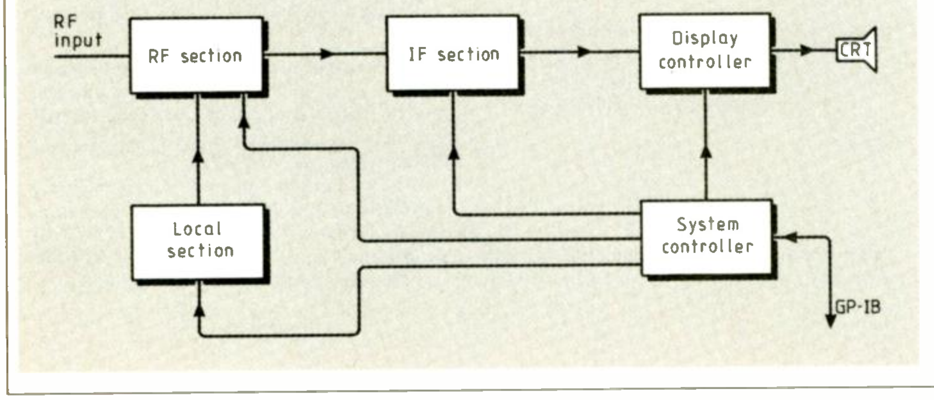


Fig. 1. Block diagram of a spectrum analyser.

Basic principles

A block diagram of a typical modern microwave spectrum analyser, the Anritsu MS710C, is shown in **Fig. 1**.

The RF front-end section is today almost invariably based on the super-heterodyne principle, as used in AM broadcast receivers.

Again there would be the problem of image responses; but because we now effectively have a multiplicity of local signals owing to the harmonics, there would be more of them. There is an additional problem, too, of an input signal mixing with other harmonics to produce more than one response on the CRT (or more than one response at the IF) as the local signal sweeps. These are termed multiple responses. So, one or two signals present at the input could produce a forest of responses on the CRT and make it difficult to distinguish the real signals.

Techniques were developed in the past to deal with this, typically involving adjusting the local oscillator sweep range by twice the intermediate frequency.

Where this happens the real signals remain in the same position on the CRT whereas the image and multiple responses move. Although measurements can be performed like this, the method is quite tedious and time consuming if there are many signals.

To improve the operation, a tracking preselector can be used. The entire family of MS710 spectrum analysers employs this technique, using a YIG

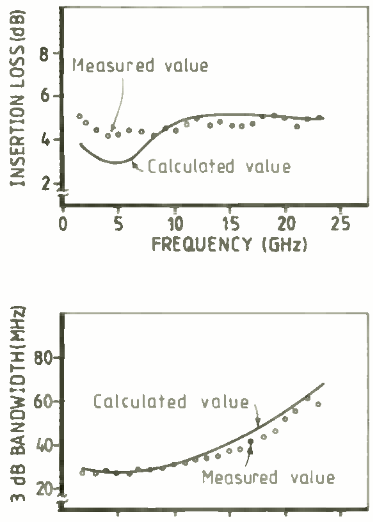


Fig.2. Performance of YIG tuned filter. This variably-tuned bandpass filter provides image rejection.

(yttrium iron garnet) tuned filter. This is a variable-tuned BPF having a bandwidth of tens of MHz (**Fig. 2**). The tuning frequency is controlled so as always to correspond to the input signal frequency. It serves, therefore, as a preselector to eliminate multiple and image signals and obviates the need for signal identification techniques. It also prevents the local signal from bleeding through to the input (the preselector is always off-tuned from the local signal by F_{IF}) and IF feed-through.

Two signals at the input spaced in frequency by $2\times F_{IF}$ will produce a similar response at the IF. Using our example again, an input signal at 1GHz will produce the same IF response as a signal at 6GHz and they would appear superimposed on the CRT. We term these signals image responses. We overcome this problem in RF spectrum analysers by including an image rejection low-pass filter before the RF stage.

However, image rejection is obviously not practical for microwave spectrum analysis where we could typically be measuring signals up to 20GHz and beyond. It is possible to design and implement wide-band swept oscillators from, say, 2 to 20GHz; but this is expensive.

A technique called harmonic mixing is very often used through considerations of manufacturing cost. With this technique, down-conversion of an input signal is performed in the mixer by using harmonics of the local oscillator signal. Thus by selecting various harmonics it is possible to use a signal oscillator to detect high frequency microwave signals. The tuning equation can now be represented thus,

$$F_{input} = nF_{local} \pm F_{IF}$$

where n is the harmonic number.

Another feature of harmonic mixing is that the conversion loss of the mixer is greater for higher harmonics. **Figure 3** shows the basic circuit for the harmonic mixer used in the MS710. Conversion loss is varied by adjusting the bias across the diode; for the first and third harmonics it is minimized by adjusting the bias to zero. The conversion efficiency can be improved slightly by inserting resonance circuits at the input and output of the mixer circuit, tuned to the intermediate frequency.

Ideally, local oscillator control should be performed by a simplified synthesizer technique, to hold down costs while ensuring satisfactory performance. A diagram of the local oscillator control circuit is shown in **Fig. 4**.

The basic difference between this and a full synthesizer local oscillator is that the signal sources f_1 and f_2 (independent quartz crystal oscillators) would be phase-locked to a common reference oscillator and f_3 would be a low frequency synthesizer employing the same reference. However, as it stands, it is possible to achieve a frequency accuracy of better than 5p.p.m.

Of course, it is important to generate accurate sweeps as well and this is performed slightly differently depending on the sweep width. For wide spans, the sweep is generated by adding a d-to-a voltage signal to the PLL compensation voltage while in open-loop configuration. For narrow spans, where accuracy is more significant, this is done in closed-loop mode.

Low frequencies

Often we are not interested just in measurements on the RF carrier but also in the low frequency information or

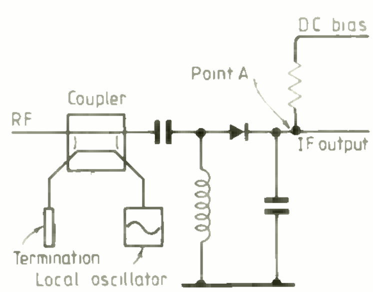


Fig.3. Harmonic mixer: down-conversion of the input signal is achieved by the use of harmonics of the local oscillator signal.

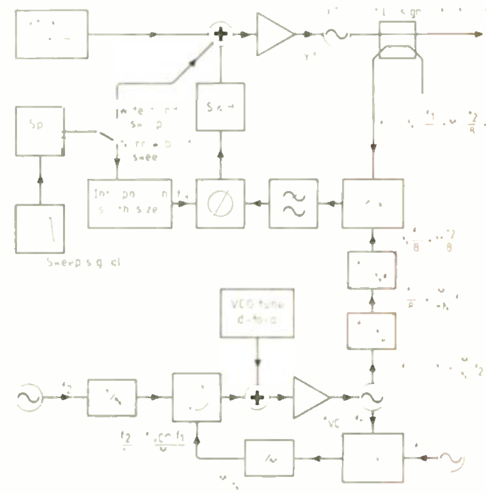


Fig.4. Local oscillator section of Anritsu's MS710C microwave spectrum analyser.

baseband signal. It is cumbersome and costly to use another low frequency analyser for these measurements and some of today's microwave analysers also have the capability of measuring to audio frequencies. This is implemented either by using the same RF stage as for microwave measurements, or by an independent low-frequency superheterodyne stage switched to the common IF section to improve the low-frequency performance.

In military applications such as guided weapons, the signal frequencies used are well in excess of the conventional microwave range (over 100GHz). However, these signals can be measured by using an external mixer and higher harmonics of the local signal to down-convert to the IF. The MS710C is an example of an analyser with these capabilities; it has a frequency range of 10kHz to 140GHz.

Quite a common feature today in *low* frequency analysers is a tracking generator. This is a swept signal source which provides a signal to which the analyser automatically tunes. In this way, the user can perform stimulus/response measurements on devices such as filters, attenuators, etc., components which are used in the baseband and IF stages of microwave sources and transmitters. Not so common is this ability in *microwave* spectrum analysis, where applications often demand characterizations of linear devices used at lower frequency stages. ■

Andrew Standen is an RF and microwave product engineer with Anritsu Europe Ltd.

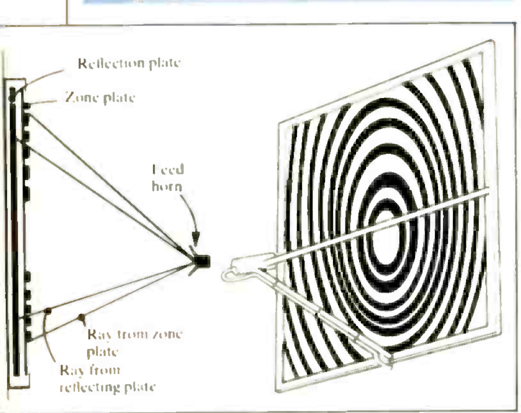
Fresnel antenna

Richard Lambley reports on a novel microwave antenna

More details of Mawzones Ltd's Fresnel antenna for microwave frequencies (May issue, page 469) emerged at a demonstration staged at the DTI's Baldock satellite receiving station, where the company has carried out much of its development work.

According to Mike Wright of Mawzones, flat Fresnel microwave antennas have been re-invented at frequent intervals during this century; but until now they have failed to take account of a large but undisclosed fiddle-factor in Fresnel's calculations, with disappointing results. Getting this right involves some serious maths, as recent papers on the subject testify.

Two versions of the Mawzones antenna were demonstrated: a transmissive type, with its concentric silver rings screen-printed on a flexible plastics sheet; and a reflective type, made of a sheet material similar to the corrugated plastics sandwich used for estate agents' placards, with the ring pattern on one side and a metallized backing on the reverse. The rings are egg-shaped, since in this way the Fresnel antenna can be made to cope with offset feeds or off-axis beams. Both types gave excellent reception of the Astra satellite on 11GHz using a surface area of well under a square metre, and the reflective version (which gave more output) could receive weaker satellites. Mawzones has a computer program which can plot appropriate patterns for satellites at various elevations, for mounting on walls or roofs at a variety of angles. For DBS reception in northern Europe, the alignment tolerance is broad enough for the same pattern to be used at sites anywhere within a 300-mile radius.



Left: signal focusing by reflective zone plate.

For domestic use, transmissive antennas could be made in the form of window blinds, which would be rolled up when not in use. Or they could be integrated with a skylight window, with the microwave feed attached to a strut on the inside. Development work is now concentrated on reducing the effective focal length of the ring pattern, to reduce the length of the strut and so improve its domestic acceptability. Antennas for mounting on exterior walls and roofs can be painted to blend them with their surroundings.

The patterns are screen-printed for Mawzones by Graphic Art, a Cambridge company which specializes in membrane keyboards.

Electrical continuity of the rings is not necessary, so it would be possible to produce very large reflective antennas for the lower microwave bands in a form resembling floor-tiles, which could simply be laid out on the ground. A feed horn would be suspended above them on struts. For the same reason, Fresnel antennas would be highly tolerant of surface damage.

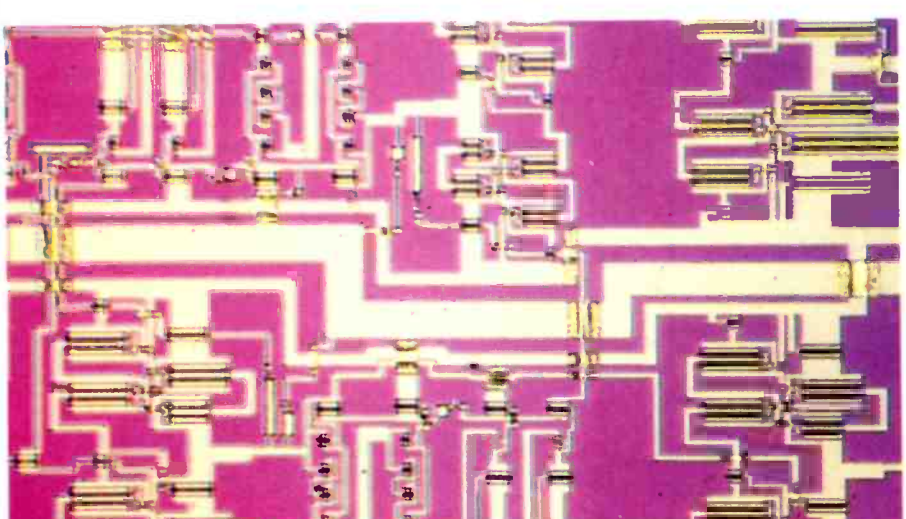
A cheap satellite antenna of this kind could prove an attractive proposition to developing countries, which cannot easily afford the high cost of large, steerable dishes. But the Mawzones antenna could also form the basis of a bargain-price terrestrial communications network. One possibility mentioned by the company is to use a commercially-available PC expansion card (costing about £300) which is designed for backing up hard discs on to cheap video cassettes. This could be used to convert a data-stream into a pseudo-video signal which could be transmitted and received economically using off-the-shelf microwave hardware. Not the PTT-approved way of doing things, perhaps, but it would undoubtedly be cost-effective. ■

Mawzones Ltd, 6 Hodwell, Ashwell, Baldock, Hertfordshire SG7 5QG, tel. 0462-742854.

Information about the Videotrax data formatting card is available from Dark Star Information Systems on 0206-578224 or from Alpha Microsystems on 0753-821922.

Op-amps from GaAs technology

Microwave technology can be applied to make 10GHz op-amps and high-precision switched-capacitor filters capable of operating at 500MHz. Chris Toumazou, David Haigh and Larry Larson report.



Interest in gallium arsenide (GaAs) for high speed analogue applications arises mainly from the higher electron mobility and peak velocity of GaAs compared to silicon[1]. These applications may be divided into higher frequency microwave circuits and lower frequency high-precision sampled data circuits using switched capacitor techniques.

GaAs has not quite lived up to its early expectations because of processing and material problems, but we are now seeing a successful and rapidly growing use of GaAs mesfet technology in the area of sampled data analogue signal processing circuits. One major application has been analogue-to-digital conversion. Despite being one of the most demanding areas of analogue integrated circuit design, 8-bit a-to-d converters[2] with sample rates as high as 1GHz have been successfully implemented in GaAs.

High-frequency switched-capacitor (SC) filters form a related GaAs technology application. Available c-mos technology switches at up to 30MHz[3]. The higher electron peak velocity of GaAs has pushed switching frequencies to 250MHz[4,5] and even 500MHz is now being realized[6]. GaAs SC filters and a-to-d converters will allow higher levels of system integration, optimization of system performance and a greater degree of flexibility.

The problem

Measured characteristic curves for a metal semiconductor fets (mesfets) produced by a typical GaAs process with a gate length of 1µm (**Fig.1**) exhibit typical fet-type characteristics. Such curves indicate a gain figure for a single device of the order of 20, much less than for other technologies such as c-mos and bipolar. The op-amp is a central component in all sampled data analogue circuits. GaAs technology poses a number of difficulties for the amplifier designer: the lack of a p-channel device (due to the low hole mobility of GaAs), a general lack of enhancement mode devices are (depletion mode devices are easier to manufacture) and low device gain. Design techniques must be developed to overcome the drawbacks.

A solution

A typical architecture for a GaAs op-amp, as with c-mos and bipolar, consists of a differential input stage with current source loads. To achieve high a gain, the

Fig.7. At 500MHz, this GaAs switched-capacitor filter is significantly faster than any of its predecessors.

current source must have a high impedance. **Figure 2a** shows a high impedance current source for GaAs mesfet devices which exhibit 'early saturation'. Simply, this means that a device saturates for a value of V_{ds} which is less than that normally required. In Fig. 2a, Tr_5 a is chosen to be much wider than Tr_4. This gives Tr_5 a negative V_{gs}, which is equal to the V_{ds} of Tr_4, and is sufficient to keep Tr_4 saturated assuming that early saturation exists. However, many processes (e.g. that of Fig.1) do not exhibit any or sufficient early saturation behaviour and, even for those that do, the device width required in Fig.2a may be unattractively high.

The alternative double level shifting technique, which solves this problem, is shown in **Fig.2b**. An auxiliary bias chain comprising Tr_7 and Tr_8 has been introduced. Now the level shifting source follower Tr_8, with its negative V_{gs}, assists Tr_5 in keeping Tr_4 saturated. Device Tr_7 similarly helps to maintain Tr_5 in saturation and also helps keep the drain-source voltage of Tr_8 relatively constant in order to maintain adequate source follower performance. The circuit in Fig.2b realizes a high impedance cascode current source without any reliance on early saturation of devices.

Another useful technique for amplifier design is illustrated in **Fig.2c**, which shows a differential input stage with the addition[7] of a voltage follower A. In the absence of the voltage follower, the gain of such a stage using typical GaAs mesfet devices is of the order of only 5. The effect of the voltage follower is to eliminate gain limitation arising from Tr_1 and Tr_2 and hence allow much higher gain. This is at the cost of increased sensitivity of gain versus device gate width.

10GHz op-amp

In the circuit diagram of the amplifier[8] **(Fig.3)**, mesfets Tr_1 and Tr_2 form the differential input pair with a cascoded tail current source comprising $Tr_{7,8}$. A high impedance current source load for the drain of Tr_2 is realized by means of $Tr_{4,6}$. Mesfets Tr_3 and Tr_4 perform two functions simultaneously. Firstly, they form part of the double level-shifting bias circuitry (as in Fig.2b) used to maintain Tr_4 in saturation. Secondly, they implement the voltage follower A in Fig.2c. Mesfets Tr_9 – Tr_{12} and diodes D_1 – D_3 form a level-shifting stage, which feeds an output source follower, Tr_{13}-Tr_{16}.

The amplifier was implemented in a 0.2μm gate length GaAs process. Measured frequency response is shown in **Fig.4a**. The step in the gain characteristic at about 100kHz is due to the frequency-dependent nature of the mesfet's output conductance due to dispersion effects*, but the normal operating region of the amplifier is above this point. The amplifier exhibits a unity-gain bandwidth of 10GHz. **Figure 4b** shows input and output waveforms of the amplifier in an inverting unity gain configuration at a frequency of 1GHz.

* Relatively slowly varying charge is trapped by the surface states between the gate and drain of the mesfet and at the interface between the channel and the substrate.

The demand for future systems with higher operating frequencies creates a need for rapid evolution. Recent development work exploits the high speed capability of GaAs more fully: features

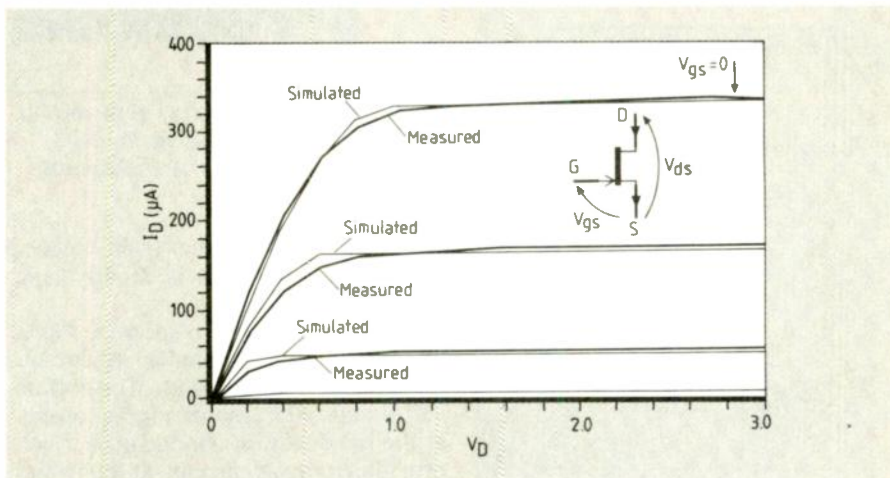


Fig. 1. Curves for a 1μm-gate mesfet exhibit typical fet-type characteristics

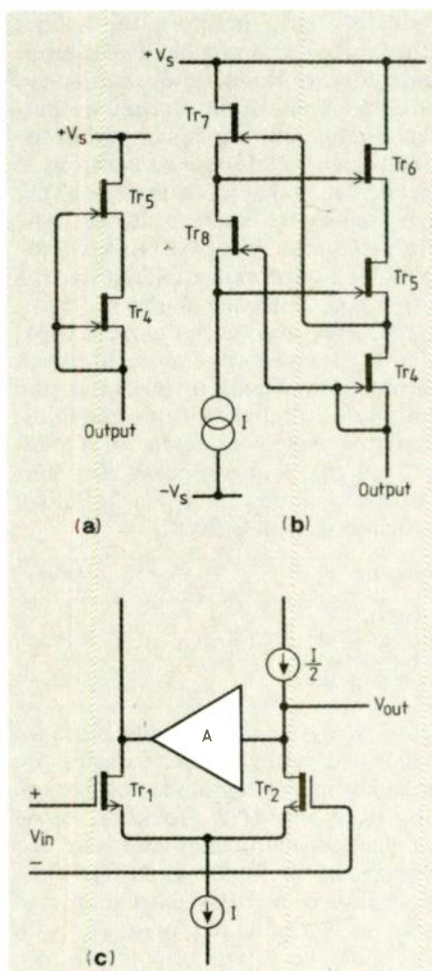


Fig. 2. High-impedance current source suitable for GaAs devices with "early saturation" (a) and a double level-shifting technique for devices without early saturation (b). In (c) is another useful op-amp design technique which solves the problem of low gain in a differential input stage using a voltage follower.

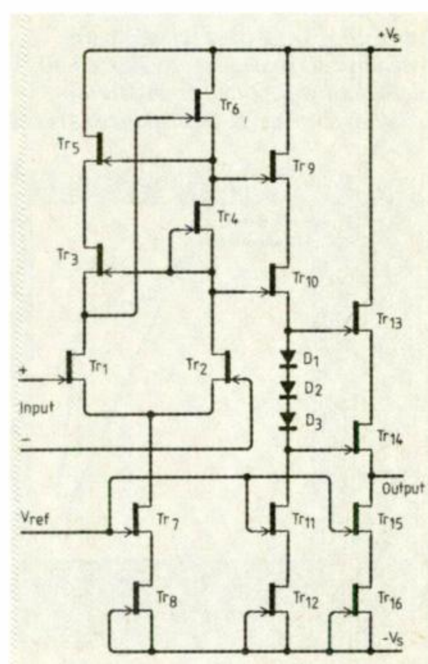
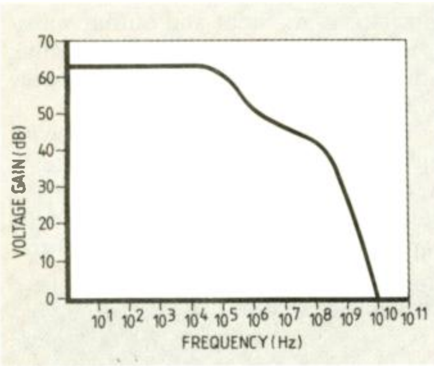


Fig. 3. GaAs fet gate lengths of 0.2μm are used in this 10GHz op-amp.



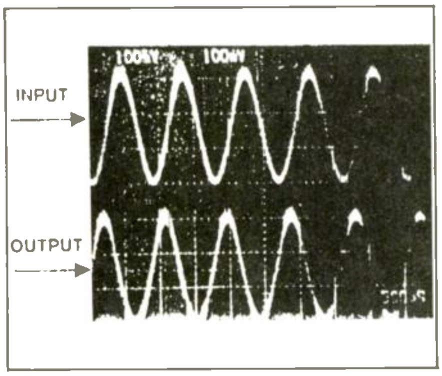


Fig.4. Measured frequency response (a) of the 10GHz op-amp of Fig.3 and its unity-gain input/output waveforms at 1GHz (b).

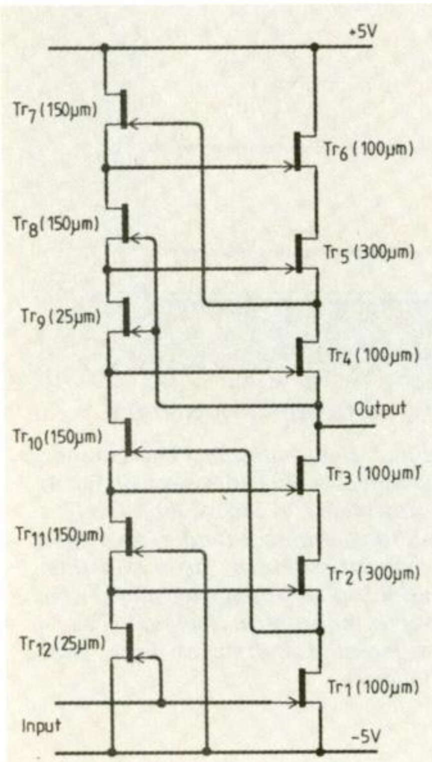


Fig.5. A higher performance op-amp than that of Fig.3 has recently been developed. Its bottom section is a dual transconductance driver and its top section is a very high-impedance current mirror.

obtained include higher gain, higher phase margin (increased stability) and faster settling time.

One such amplifier[9] is shown in **Fig.5**. It consists of two similar subcircuits placed one above the other. The bottom half contains the amplifier input terminal and is a dual transconductance driver providing output currents at the drains of Tr_3 and Tr_{10}. The main driver transistors Tr_1 and Tr_{12} each have *two* cascode transistors and therefore the output impedances are very high. The cascode transistors are biased by an extension of the double level-shifting technique illustrated in Fig. 2b. The upper part of the circuit features different connections to Tr_4 and Tr_6 so that it functions as a very high impedance current mirror. The effect of these blocks is to give the amplifier a push-pull capability at the output node, allowing double the transconductance and output current capability for a given chip area and power consumption. **Figure 6** shows the gain and phase responses of this amplifier, simulated using parameters for a modest 1µm gate width process. Key performance parameters for a 1pF load capacitance are as follows:

Parameter	Value
Gain (dB)	62
GB Product (GHz)	3.5
Phase margin	65°
Min. settling time (ps)	450

This design is a useful building block for high-speed systems. Such systems are currently being integrated and an example of a 500MHz clocking GaAs switched-capacitor filter chip using the amplifier of Fig.5 is shown in **Fig.7**. The application of such design techniques to more advanced GaAs processes with gate widths less than 1µm should produce truly outstanding performance.

GaAs technologies are maturing rapidly: much higher levels of integration will soon be possible. In the development of circuit techniques, emphasis has been placed upon the reduction of chip area and power consumption. From the simulation results for integrated GaAs high-precision filters, we expect circuit switching frequencies as high as 500MHz. Recent amplifier developments have resulted in further reductions in settling time, approaching 200ps for a 1µm GaAs technology[10,11]. This is an encouraging step towards the eventual development of 1GHz clocking switched-capacitor systems. ■

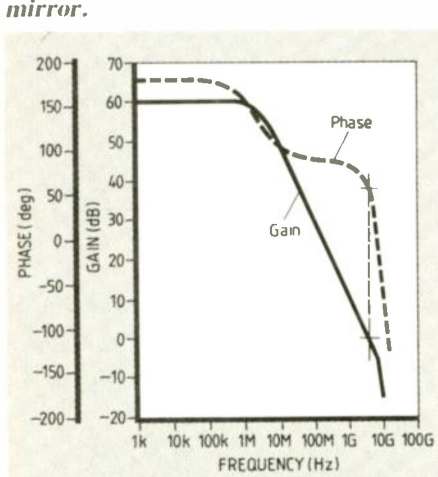


Fig.6. Gain and phase response of the amplifier of Fig.5.

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MMIC-compatible antennas

MMICs have now reached a maturity roughly equivalent to that of silicon in the early 1970s. Alan Sangster surveys progress in the antenna technology associated with them.

Since the first use of gallium arsenide (GaAs) in active microwave devices more than two decades ago, virtually all major microwave companies have been directing effort towards the realization of monolithic microwave integrated circuits (MMIC).

This technology has evolved along both digital and analogue routes. Digital MMICs are fulfilling signal-processing roles such as analogue-to-digital conversion, signal weighting and multiplexing, while analogue MMICs handle receiver and transmitter functions.

MMICs are fabricated by optical photolithography and contact printing techniques on GaAs wafers, which are typically 3 – 4cm in diameter and of the order of 0.1mm thick. One side of the wafer is totally metallized to form a ground plane, while the other side supports microstrip circuit elements such as inductive loops, blocking capacitors, transmission line stubs and passive and active devices (**Fig.1**).

As MMIC technology has advanced, so its potential applications have grown. One of the most significant is in phased-array radars. A phased-array antenna system, using integrated circuit technology to provide the transmit/receive functions at each element of the array, could have a significantly lower cost-per-element. Depending on complexity, phased-array elements can cost between £500 and £5000 each. Since a typical array may contain between 1000 and 100 000 elements, the benefits are potentially very considerable, and could lead to the long-sought goal of economical phased-array radar systems.

Part of the key to realizing an efficient and economical phased array with MMIC-fed antenna elements is to choose a radiator which is microstrip-compatible. Two types of elemental radiator meet this basic requirement: the slot and the patch. Some typical radiator geometries are presented in **Fig.2.** In Fig.2a, for example, a rectangular slot (usually λ/2 long) has been etched into the ground plane below the microstrip feed line from the active circuitry. For the line and slot orienta-

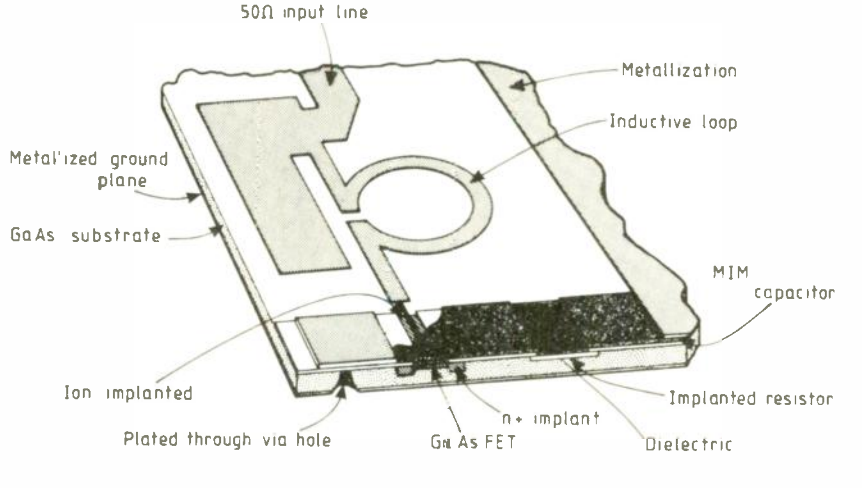


Fig.1. Outline of a monolithic microwave integrated circuit (MMIC).

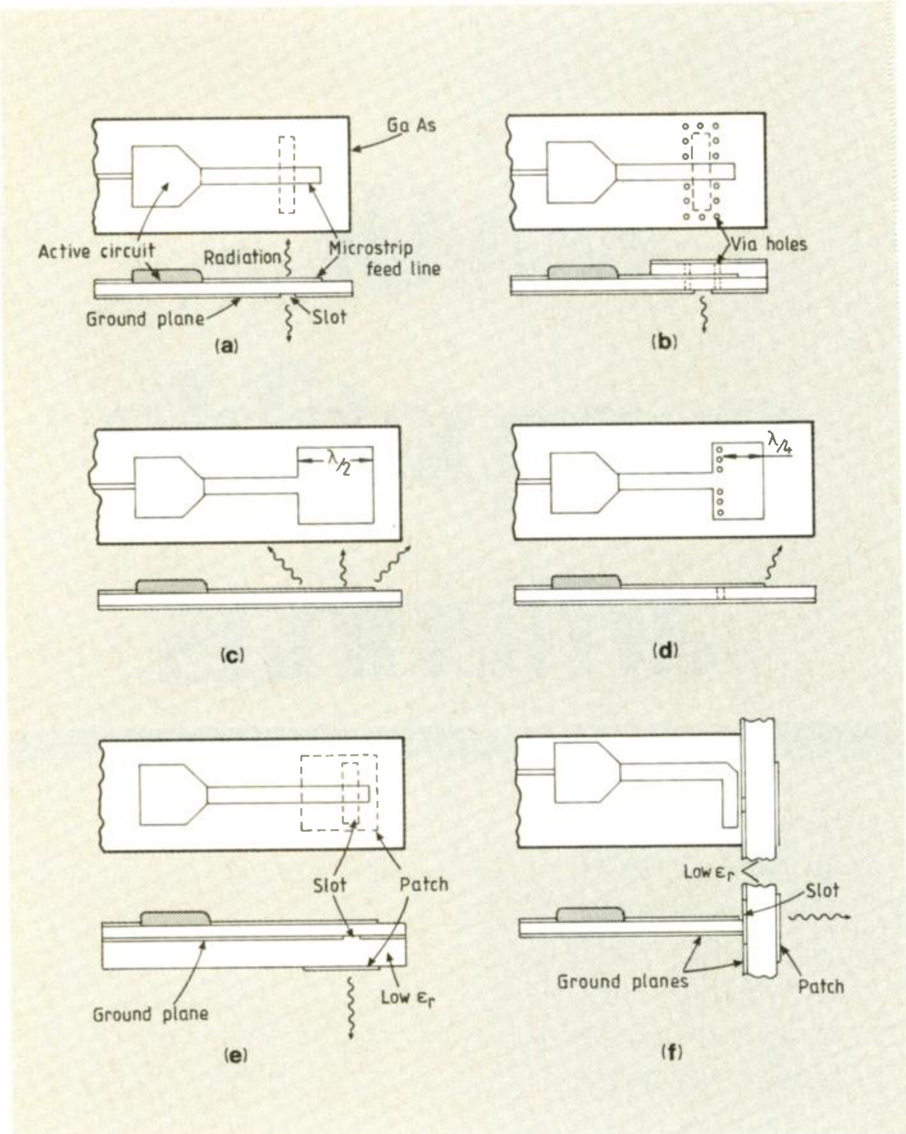


Fig.2. Some MMIC-compatible antennas.
(a) Microstrip-fed slot radiator.
(b) Microstrip-fed cavity-backed slot radiator.
(c) Conventional patch antenna.
(d) Short-circuited patch antenna.
(e) Slot-fed patch antenna.
(f) Edge slot-fed patch antenna.

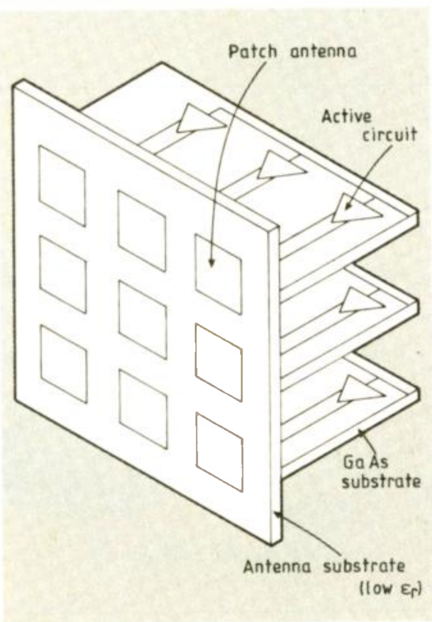


Fig.3. MMIC array antenna.

tions shown, the magnetic field of the microstrip quasi-TEM mode excites the slot and results in radiation above and below the ground plane. This bidirectional radiation feature is, however, not very useful in an antenna; and the arrangement lacks RF isolation between the radiator and the active circuitry. The cavity-backed slot antenna (Fig.2b) can eradicate most of these problems, but is harder to manufacture.

The simple patch antenna arrangement (Fig.2c) is relatively easy to fabricate. If the patch length is of the order of λ/2 in the direction shown, the radiated fields from the transverse edges of the patch add in a direction normal to the surface of the patch. Unfortunately, this simple geometry also suffers from insufficient isolation between the radiator and the active circuitry. The λ/4 patch radiator of Fig.2d, which is shorted using via holes along one transverse edge, provides a useful size reduction in the patch, but isolation is little better.

Two schemes which provide effective isolation yet do not add significantly to the fabrication process are shown in Fig.2e, 2f. A further advantage of these schemes is that the antenna substrate need not be GaAs, and consequently bandwidth improvements can be realized by using thicker substrates with lower relative permittivities. Isolated patch excitation can also be secured by means of probe coupling between the microstrip feed line and the patch. However, the probe coupling mechanism is less compatible with the MMIC geometry and poses constructional problems. It is also pertinent to note here that slots and patches need not be rectangular.

How a phased array could be implemented is shown in **Fig.3**, using the element of Fig.2f. It can be seen that this antenna element can be stacked readily to create a planar array. However, while MMIC technology married to microstrip radiator technology appears to lend itself to active-array development, many practical problems remain to be solved. In particular, these arrays are very sensitive to material and mechanical tolerances. Furthermore, they continue to present design problems due to the fact that the effects of mutual coupling or RF interference between elements, particularly with patch radiators, remain difficult to predict. ■

Dr Alan J. Sangster is Reader in Microwave Electronics in the department of electrical and electronics engineering at Heriot-Watt University, Edinburgh.

Near-millimetre-wave techniques

Access to the near-millimetre-wave region has revolutionized our understanding of the universe by letting us see things never seen before — Stafford Withington.

Fig.1. The James Clerk Maxwell Telescope. This instrument is capable of observing at frequencies ranging from 100GHz to 900GHz. The 15m telescope is housed in a temperature controlled building which is 90ft high and 100ft in diameter.

By receiving and analysing the line radiation associated with molecular and atomic transitions, and the continuum radiation associated with low-energy particle collisions, a wide variety of cold (1K-1000K), and therefore optically dark, astronomical phenomena can be studied. For example, near-millimetre-wave observations have revealed the existence of vast interstellar clouds of complex molecules, the most massive objects known in our galaxy. It has also been discovered that certain young stellar objects radiate huge oppositely directed jets of particulate material; these collimated outflows attest the existence of intricate star-forming structures. The cosmic background radiation, whose form was established when the universe was only minutes old, can be detected and analysed to give information about the creation and nature of galaxies. This cursory list could be extended for pages.

It has only been possible to view the universe at near-millimetre wavelengths (100GHz-1000GHz) by constructing a new generation of radio telescopes. Foremost amongst these is the James Clerk Maxwell Telescope (**Fig.1**), sited on the 14000ft mountain

Mauna Kea in Hawaii. Submillimetre-wave telescopes have to be operated at high altitude because short-wavelength radiation is strongly absorbed by atmospheric water vapour – the atmosphere above Mauna Kea is one of the driest and cleanest in the world. The performance of a radio telescope is, to a large extent, determined by the diameter and the surface accuracy, measured in wavelengths, of the main reflector; a telescope with large smooth surface has high spatial resolution and high sensitivity. The antenna of the James Clerk Maxwell Telescope is 15m in diameter, and it is parabolic to within 0.035mm. This remarkable degree of conformity allows the telescope to operate at frequencies as high as 900GHz with good efficiency. At this frequency, the telescope has an angular resolution of just a few arc seconds.

Conceptually, and to some extent practically, the biggest technical challenges faced by radio astronomers have been, and will continue to be, in the area of very low-noise near-millimetre-wave receiver development. The fundamental problem is that it is extremely difficult to interpolate between those concepts and techniques that are applicable at radio wavelengths and those that are applicable at optical wavelengths. For example, consider power detection. At radio wavelengths, the standard method for detecting power is to apply the incoming signal to a device whose resistance is nonlinear. The oscillatory potential across the devices induces a direct current which is proportional to the amount of power absorbed. Clearly, the phenomenon of power detection is classically understood in terms of the degree of curvature of the stationary amplitude response. It should be noted in passing that according to this description a classical detector can, in principle, be made arbitrarily sensitive by making the detector element increasingly nonlinear.

At optical wavelengths, the devices used as detectors – photoconductors and photodiodes – are all conceptually based on the photoelectric effect, the ejection of electrons from bound states by energetic photons. Thus the language of power detection at optical wavelengths is far removed from that of power detection at radio wavelengths. At sub-millimetre wavelengths, this dichotomy must disappear and the language of power detection must be intermediate between the classical and quantum extremes.

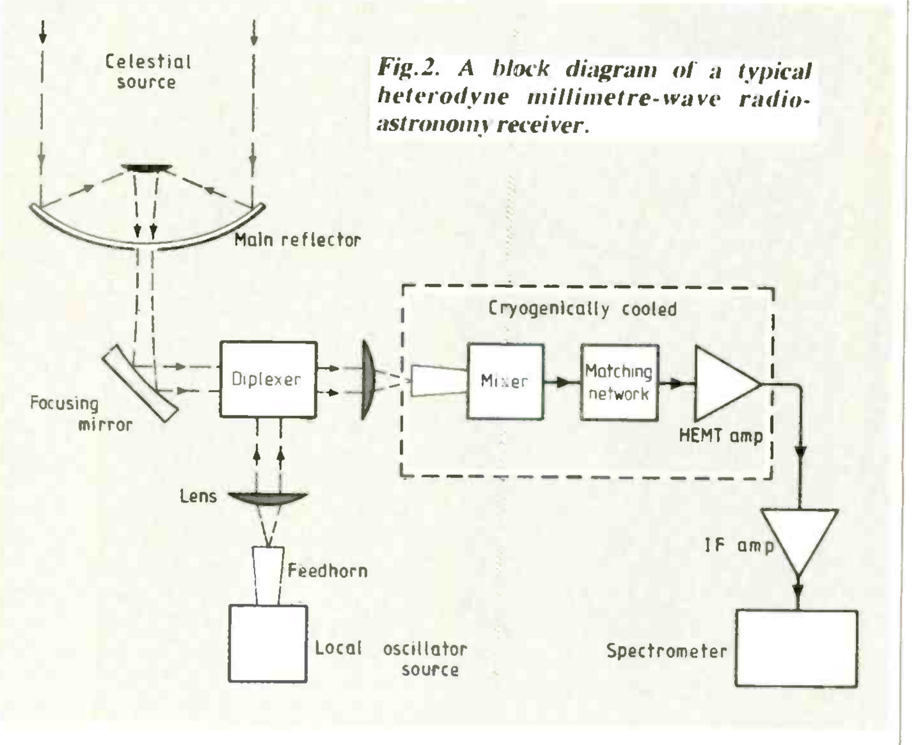


Fig.2. A block diagram of a typical heterodyne millimetre-wave radio-astronomy receiver.

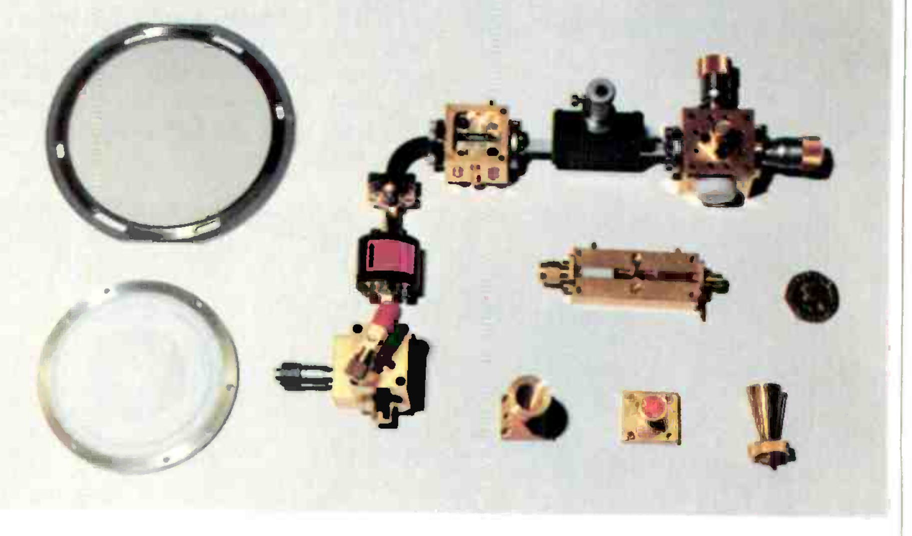


Fig.3. A collection of receiver components. Top left is a wire polarizing grid for 350GHz. Bottom left is a hyperbolic PTFE lens for 350GHz. The assembly in the centre is a 330-360GHz solid-state local oscillator source. This assembly consists of working from left to right: a 90GHz Gunn oscillator, a ferrite isolator, a directional coupler, and attenuator, and a frequency quadrupler. The unit to the left of the 20p coin is a cryogenically-coolable 4GHz HEMT amplifier. Bottom right are three millimetre-wave horns.

Heterodyning

A heterodyne receiver down-converts a band of radio frequencies to a band of more manageable intermediate frequencies. Spectral information and phase are conserved during this process and therefore narrow, high-frequency spectral lines can be analysed through relatively low-frequency signal processing. The configuration of a heterodyne, as distinct from a bolometric, near-millimetre-wave radio-astronomy receiver is shown in Fig.2. The main antenna and subreflector of the radio telescope focus the electromagnetic energy into a roughly collimated beam which is only a few tens of millimetres in diameter. This extremely weak beam is

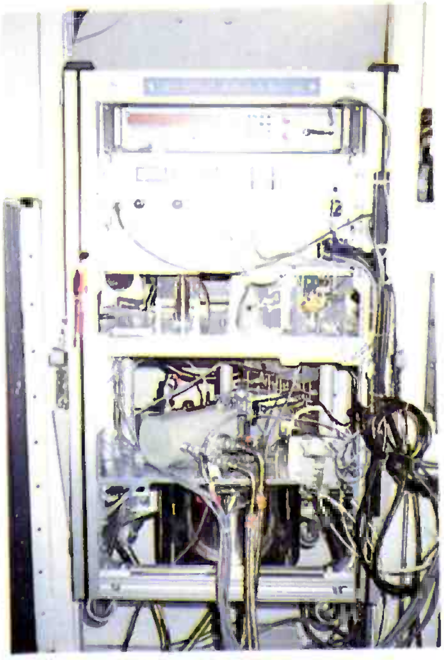


Fig.4. A 230-270GHz dual-polarization cryogenically-cooled Schottky-diode receiver mounted in the receiver cabin of The James Clerk Maxwell Telescope. The signal beam enters the receiver through the aperture above the middle plate.

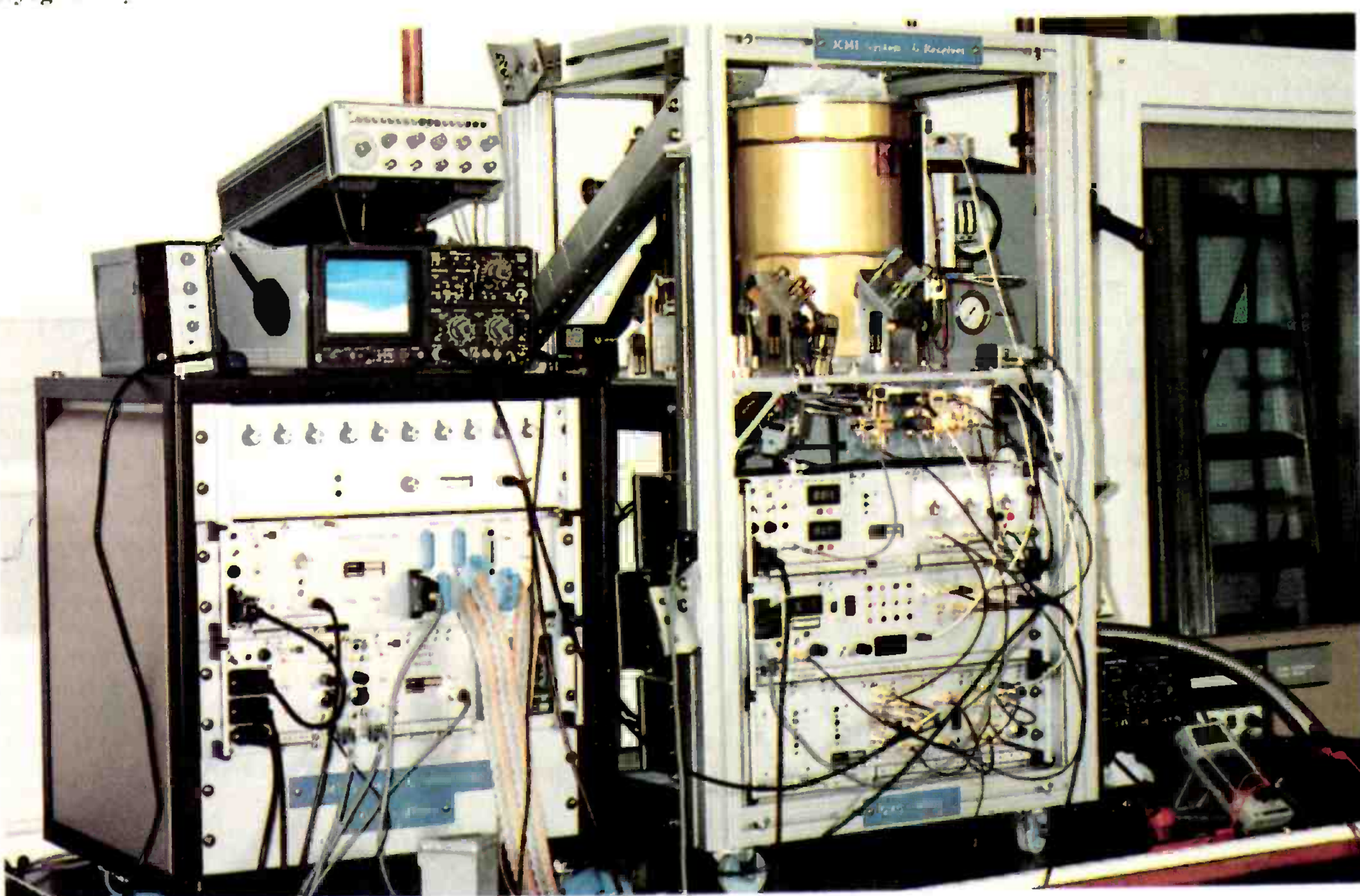
Fig.5. A 460-490GHz dual-polarization cryogenically-cooled InSb receiver.

superposed on a relatively high-level local-oscillator signal in a diplexer, the combined power being then converted from free-space propagation into waveguide propagation by a small, corrugated horn. Once the power is in the waveguide, which typically measures 0.8×0.4mm, it passes through various impedance-matching structures before being applied to a mixer element of only a few femtofarads (10^{-15}F) capacitance. Finally, the microwave signal emerging from the mixer is amplified, in a cryogenic HEMT (high electron-mobility transistor) amplifier similar to the one shown in **Fig.3**, and then processed in an autocorrelation spectrometer. The physical complexity of very low-noise near-millimetre-wave receivers is greatly increased by the need to cryogenically cool the mixer element and the first few stages of IF amplification. Semiconductor mixers only require cooling to a physical temperature of around 20K, whereas superconducting mixers require cooling to 4K; in both cases closed-cycle helium refrigerators are now being used. A complete 230-270GHz Schottky-diode millimetre-wave receiver is shown mounted in the receiver cab in of the James Clerk Maxwell Telescope in **Fig.4.** A 460-490GHz InSb (indium-antimony) receiver is shown in **Fig.5**.

Local oscillators

The construction of powerful, tunable, and reliable near-millimetre-wave local-oscillator sources is a major problem. For many years, reflex klystrons were used at the longer wavelengths, but these devices are expensive and short-lived. High-frequency backward-wave oscillators, also known as carcinotrons, are an alternative, but are mechanically cumbersome, and require high-voltage power supplies. The current solution is to combine an InP (indium phosphide) Gunn oscillator, operating in the frequency range 50-120GHz, with a GaAs varactor-diode frequency multiplier. This type of solid-state source is compact, reliable, and capable of being phase locked to a low-frequency synthesizer. Typically, 1 mW of power can be generated at 350GHz, 100µW at 500GHz, and 30µW at 700GHz. These power levels are sufficient for driving Schottky-diode mixers at frequencies of up to 500GHz, and superconducting mixers at frequencies of up to 1THz (10^{12}Hz). A clumsy but effective solution to driving Schottky-diode mixers at frequencies above 500GHz is to use a molecular laser.

A variety of new oscillators may become available as a consequence of



the rapid advances of III-V semiconductor heterojunction technology. The resonant tunnelling diode is a device which essentially consists of a stack of closely coupled potential wells, the width of each of which is so small that only electrons having particular energies can exist within the wells. As the direct voltage across the stack is varied, the allowed energy levels are aligned and misaligned and when they are aligned, electrons can tunnel through the structure. The resulting negative differential resistance on the DC current-voltage curve can be used to make high-frequency oscillators.

Mixing

A whole range of fundamentally different types of nonlinear device can be used as near-millimetre-wave mixers, two of the most popular being the Schottky-barrier diode and the superconducting quasiparticle tunnel junction.

A Schottky-barrier diode consists of a metallic layer deposited on a lightly doped semiconductor, which is epitaxially grown on a heavily doped conducting substrate. The epitaxial layer is conductive except in the vicinity of the metal, where an insulating depletion zone is formed. Electrons can cross the depletion layer either by thermal excitation or by tunnelling; the former mechanism dominates at room temperatures, and the latter at cryogenic temperatures. At frequencies below 500GHz, it is physically possible to mount a Schottky-barrier diode in a fundamental-mode waveguide structure, where the diode, which is only a few microns in diameter, is contacted by a small wire which bridges the waveguide. A tiny corrugated horn is used to couple the incident electromagnetic energy from the principal Gaussian mode to the waveguide mode; a selection of horns is shown in Fig.3. At frequencies above 500GHz, it becomes exceptionally difficult to machine suitably small waveguide components, and therefore open monopole structures are often used. Cryogenically cooled Schottky mixers achieve noise temperatures of around 100K at 100GHz, 1000K at 500GHz, and a few thousand degrees at 1THz.

The superconducting quasiparticle mixer is an extraordinary device consisting of two superconducting films separated by a dielectric layer only a few nanometres thick. The energy levels of the device are such that when the bias voltage increases beyond some threshold, usually a few millivolts, entities very similar to electrons, called quasiparticles, can tunnel across the structure, causing a discontinuous increase in current; at higher bias voltages the current increases linearly. **Figure 6** shows the DC current-voltage curve of a niobium-aluminium oxide-niobium junction. The characteristic is hysteretic, and has a direct supercurrent due to an additional process which involves the tunnelling of electron pairs; this process can also be used for mixing, but it is noisy. **Figure 7** shows that, when high-frequency radiation is applied, a series of steps is induced. Each of these steps is a photon energy wide, and indeed they occur as a result of photon-assisted tunnelling of quasiparticles across the barrier. The device can be used as an extremely low-noise mixer.

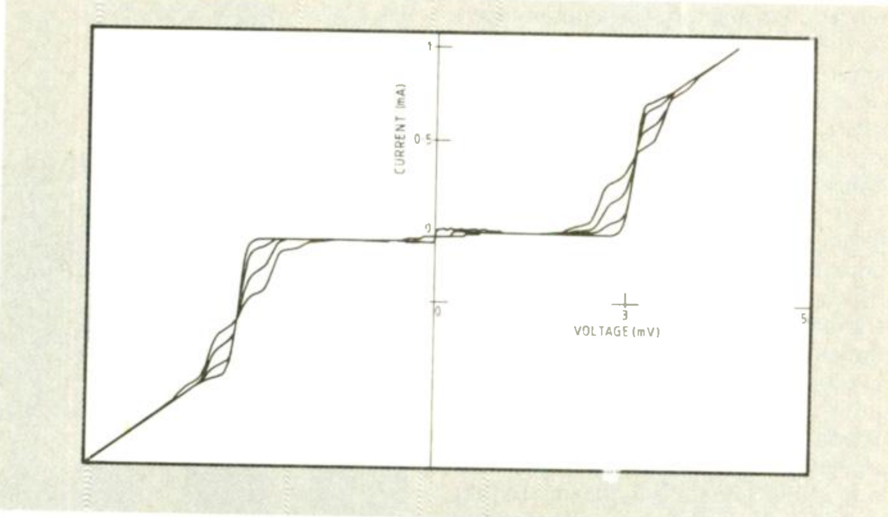


Fig.6. The hysteretic DC current-voltage curve of two superconducting tunnel junctions in series. A DC supercurrent flows at zero bias voltage due to the tunnelling of electron pairs.

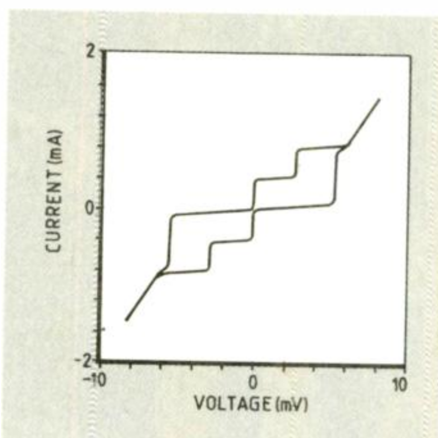


Fig.7. The DC current-voltage curve of a superconducting tunnel junction with various amounts of 100GHz radiation applied. Each step is a photon energy wide.

It turns out that, at low frequencies, where the photon energy is less than the voltage width of the nonlinearity, the mixer behaves classically, whereas at high frequencies, where the photon energy is greater than the voltage width of the nonlinearity, the device behaves quantum mechanically. The smooth transition from classical to quantum behaviour has some remarkable consequences, including the onset of quantum-limited noise temperature (meaning that the sensitivity is limited by the Heisenberg uncertainty principle) and classically forbidden conversion gain. Quantum reactances also appear due to the sloshing backward and forward of quasiparticles between states on opposite sides of the barrier which are not exactly matched in energy. To date, superconducting quasiparticle mixers achieve noise temperatures ranging from 50K at 100GHz to 400K at 400GHz. The upper frequency of operation is likely to pushed into the far infrared within the next few years.

Quasioptics

The loss inherent in fundamental-mode waveguide structures at near-millimetre wavelengths has forced the development of quasioptical components for energy transportation, filtering, beam combining, and switching. Such components are called quasioptical because the input and output beams do not strictly behave according to geometrical optics. Once again, we encounter the situation where the relevant concepts are neither purely radio nor optical, but a curious blend of the two. At optical frequencies, typical beam-forming components are many thousands of wavelengths in diameter, but at frequencies from 100GHz to 1000GHz, typical components are only a few tens of wavelengths in diameter, and as a consequence diffraction becomes important. Fortunately, one does not have to solve complicated diffraction problems every time one wants to design a receiver. Instead, a very elegant scheme called Gaussian optics can be used in a very straightforward manner. The name Gaussian optics comes from the fact that a propagating, expanding, circularly-symmetric beam is considered to be a sum of beams, each one of which has an transverse amplitude distribution whose envelope is a Gaussian function. **Figure 8** shows the quasioptical subassembly of a 330-360GHz, dual-polarization, cryogenically-cooled Schottky-diode receiver. Focusing is achieved with off-axis ellipsoidal mirrors and PTFE lenses, beam splitting and filtering is achieved with very fine wire grids and meshes, and switching is achieved with chopper wheels. The diplexer is a polarizing interferometer. Antisymmetric movement of the two corner cube reflectors changes the differential path length between the split beams, and symmetric movement changes the absolute path length. The former is used to tune the interferometer, and the latter is used to remove the effects of standing waves.

A new discipline

The near-millimetre-wave region of the electromagnetic spectrum contains a wealth of information about the nature of the universe. It is remarkable that by passively observing a distant object it is possible to measure in great detail its physical, kinematic and chemical structure. To gain access to the near-millimetre-wave region of the spectrum it has been necessary for scientists to enhance existing physical concepts and to develop fundamentally new engineering techniques. It is now clear, that this work will form the foundation of a new engineering discipline. For example, a number of university research groups are already beginning to develop near-millimetre-wave integrated circuits in the form of imaging microchips. These and related innovations are going to have a profound impact, not only on radio astronomy, but also on space communications, high-energy plasma physics, atmospheric physics, medical physics, and many other branches of science and engineering.

The James Clerk Maxwell Telescope is principally financed by the United Kingdom Science and Engineering Research Council, and managed by The Royal Observatory, Edinburgh. The Dutch and Canadians also make substantial financial and scientific contributions. The systems shown in this article are the collaborative work of many individuals. The institutions involved include: the Physics Department of Cambridge University, the Metallurgy Department of Cambridge University, the Receiver Group of The Rutherford Appleton Laboratory, the Physics Department of Queen Mary College, London, and Millitech Corporation. ■

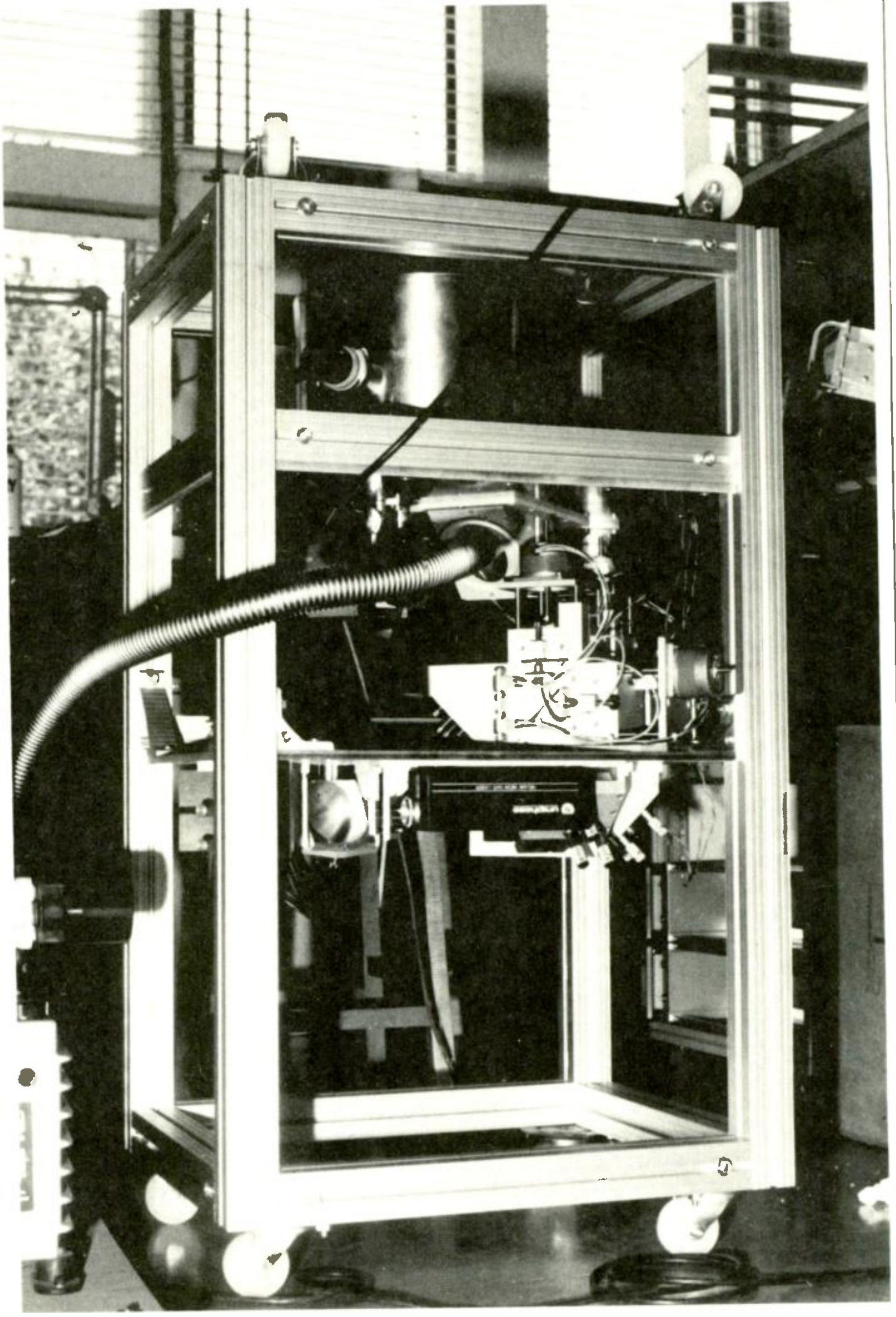


Fig.8. The quasioptical subassembly of a 330-360GHz crygenically-cooled Schottky-diode receiver. This photograph was taken while the cryogenics were being tested. The vacuum vessel at the top of the frame houses the low-temperature mixers and amplifiers. The unit on the middle platform is the polarizing interferometer. A number of focusing mirrors can just be seen beneath the middle platform.

Stafford Withington is with the Cavendish Laboratory, Cambridge

Faster than light?

Pappas's letter (April, 1989), p.414) does nothing to clarify the alleged faster-than-light experiment, apart from a trivial point about the transformer. Instead it just adds more to the fog, while failing to refute our criticisms (Letters, March, 1989).

We could spend years going into more and more arcane explanations and refutations, but this would be pointless (the Catt Anomaly saga shows this). The problems for Pappas and Obolensky is straight-forward: I and several others have given explanations for their results in terms of tried and tested physics, so it is up to them to show that none of these explanations are plausible. Occam's Razor places the burden of proof on them, not us.

As an absolute minimum, Pappas and Obolensky must show that the initiating event (the closing of the relay) occurs at their t = 0 (the point at which the oscilloscope traces first start to change), and not several nanoseconds earlier, as the rest of us contend. As their experiment contains no direct record of the relay closing, I don't see how this can be claimed.

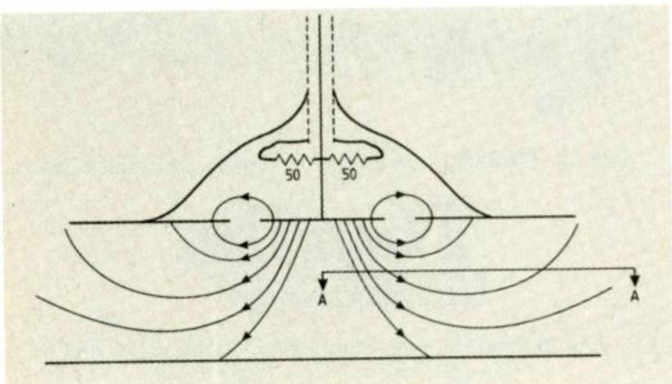
Tim Bierman
Hendon
London NW4

Cross-field antenna

I am afraid that the 'unprecedented concept in antenna theory' is simply a wrong concept. Figure 1 showing H_p and H_D opposed is incorrect. In fact, H_D, will grow linearly out to the edge of the plates and then fall as l/r (neglecting edge effects and assuming the dimensions of the plates are much less than λ). The experiment shonw in Fig.3 tests a different situation. The authors have forgotten the strong field across the gap which results in a field similar to that in the accompanying sketch. Measuring the vertical component of EH along section AA will give a result similar to Fig.5.

Any book on antenna theory which discusses horn antennas will describe the radiation as produced by the distribution of E and H in the aperture. For wire antennas it is useful to use the current element approach. Both approaches are based on Maxwells equations; it is a question of which approach is most convenient for a particular antenna.

Brian Farrelly
Kolstibotn
Landas
Norway



Weinstock

What evidence does Mr Burton (Feedback, June, 1989) have to support his statement that GEC is 'the most profitable engineering group in the U.K.'? In fact, its performance is the worst in its sector. What evidence does he have to support his statement that 'many smaller companies taken into the GEC group would have disappeared but for GEC.'?

What evidence had he to support his statement that 'GEC... has proved to be a good investment for its shareholders.'? I understood that the major shareholders, concerned at the continuing relative decline of GEC, are seeking to oust Weinstock.

In her book The Baroque Arsenal, (Deutsch 1982), Mary Kaldor discusses the effect of dinosaurs like GEC on p.167:

> "They absorb expenditure that might otherwise be spent on investment or consumption in civilian industry. And they absorb skilled people – designers, scientists, engineers – who might otherwise be thinking up the innovations needed to better the future .. warped concepts of technical advance trickle down and distort the application of new technologies, further stunting their development...
>
> Nor is economic decline offset by some definable addition to military power. The growing cost of the modern weapons system is matched by diminishing effectiveness..."

GEC is the ultimate, effete cost-plus dinosaur. I do not know about Siemens, and so I do not know whether the present merger will enable a larger dinosaur to be, as Kaldor puts it, "...carried along in an autistic momentum whose only limit is the size of the (European) defence budget."

I have been writing about my concern over the state of GEC for nearly a decade, to MPs, the Secretary of State for Defence, the MoD, and others. My efforts culminated in a farcical three-hour meeting in Whitehall, where I confronted four of the senior MoD officials who were assiduously pouring billions of taxpayers' money down the drain.

The MoD is now refusing to tell a Commons Select Committee how many of its staff have left the MoD and followed the slush fund into employment in "defence" contractor companies.

Ivor Catt
St Albans

Reference. *Wireless World* November 1980, p.57.

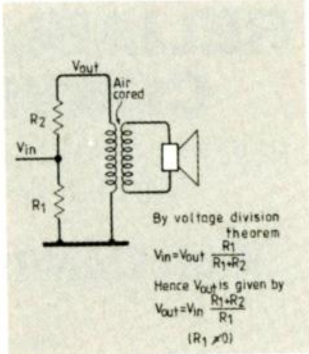
Audio – the last word

I have read with interest the debate over the past months on the objective and subjective assessment of audio amplifier quality.

My own research over the years has revealed a number of startling facts about the nature of the amplifier, verified by rigorous listening tests.

To quantify the results of these tests, I have developed a number of orthogonal parameters which may be used to compare amplifiers precisely, without bias. These parameters are as follows.

Karmic impact, K. This describes the height of spiritual



enlightenment achieved by the listening experience. For a given section of music which, for the purpose of amplifier quantification must be played only on natural acoustic instruments, preferably made of wood, it is dependent only on the amplifier and speaker and associated speaker and mains cables.

Tonal correction, T. This normalizes the listening experience for the presence of tonal components. It is well known in acoustics that the subjective impression of sound is strongly affected by the presence of tonal components.

Visual impression, V. This normalizes the listening experience for the visual environment, i.e. the room size and decor, the amplifier and speaker size, shape and colour. My listening tests were carried out in a room without windows, painted pink with a blue filtered light. Other colours in the optical spectrum interfere with the listeners' objectivity and came make him/her restless and angry.

These three parameters provide an all-encompassing and scientifically based determination of the performance of any amplifier. An overall figure of merit is FOM=K.V.T. Listening tests carried out on a wide variety of amplifiers revealed the following facts.

- Any performance is strongly dependent on the frequency of the mains. Furthermore, unregulated power supplies give a more natural sound than regulated supplies.
- Capacitors have no effect on the sound. It is well known that these are perfectly linear.
- Wirewound resistors produce a nasal sound, as opposed to carbon resistors which produce a warm fuzzy sound.
- Large amounts of negative feedback produce large amounts of negative Karma.

● Most music contains significant tonal components.
● Many amps are ugly, reducing V. This is mainly due to the size of the cabinet.
● All active components are nonlinear.

These discoveries lead me to search for the ultimate design for an amplification system which would be free from their limitations. We cannot do much about the musical content but, after much research, I developed what I call the Passive Amplifier, which is completely free from the other defects discovered.

The transformer is necessary to reduce the large voltage V_s down to a level suitable for the speaker, and to reduce the impedance of R_1 and R_2 seen by the load.

I hope this may be of interest to your readers.

Mark Poletti
Auckland
New Zealand

Ball-bearing motor

I was reading with some interest the article entitled "The intriguing ball-bearing motor" in the April 1989 issue of *Electronics and Wireless World*. According to the author, the motor violates the principle of the conservation of energy in its operation. He bases his reasoning on the principle of the operation of the motor, and on the calorimetric measurements. However, I detect a flaw in his reasoning in his interpretation of the calorimetric measurements, assuming that they were done with sufficient accuracy.

In the initial stages of the measurements, a stationary motor was placed in the calorimeter, and its "resistance" was measured. Herein is a possible flaw in his reasoning: it is possible that the energy fed into the system could have gone into distorting the outer race, caused by the ball's bulge. If that is the case, then there is no energy violation, because in the second part of the calorimetric measurements, the energy which had gone into distorting the race, was now used to rotate the motor. That is to say, the initial 'resistance' which he measured was not simply the electrical resistance, but instead part of the energy had gone into distorting the outer race.

Kah Yep Zee
Stockton
California
USA

I posted my letter about Marinov's ball-bearing motor (Letter, June, 1989) before I noticed the date on the cover, so I consider myself suitably April-fooled.

The trouble is, the article was no more implausible than most of the other fringe science you publish, though I should have twigged from the box about the episode with the car batteries. Also, I was sure I'd seen something similar on *Tomorrow's World*: maybe they were in on the joke.

That leaves me with only one question: it was a spoof wasn't it?

Tim Bierman
Hendon
London

Having read the article by Dr Marinov, appropriately in the April issue of *EWW*, I thought you may be interested in the results I obtained using parts from a small burned-out induction motor to make a ball-bearing motor.

The ball races used were ¾in diameter and contained six balls. I passed current into the outer race of one bearing along the shaft which I soldered to the inner race to reduce voltage drop and out of the outer race of the other bearing. With 20A flowing, the shaft rotated at 1700RPM with a voltage drop of 2.1V; with 25A, the speed increased to 2000RPM but the voltage drop stayed more or less constant. Spraying the races with oil caused the speed to rise to 2500RPM, again with 2.1V drop.

I then loaded the shaft with a fan which the original motor drove at 3000RPM, taking about 18W. Assuming the original motor to have an efficiency of 70%, the fan would have consumed about 13W. The Marinov motor, with 25A, rotated the fan at 650RPM with a voltage drop of 1.9V. Since the power consumed by the fan blades is proportional to the cube of the RPM, the fan dissipated about 0.13W. The overall electromechanical efficiency of my Marinov motor was thus about 0.3%, which is comparable with a Newcomen steam engine, which enjoyed considerable commercial success. Dr Marinov would, of course, insist that the 0.13W is free power.

As far as the reason for operation is concerned, I think Dr Marinov has put forward a credible theory: at least, more credible than any I can devise using electromagnetic forces and finger-twisting Fleming rules. However, I am at a loss to understand how the expanding balls can do any work if, as he states, the direction of motion and the direction of motion are at right angles. If one postulates a small time delay between the production of heat and the steel expanding, then the expansion of the ball races will take place just behind the point of contact, once motion is established, so that a torque will be produced and work done. However, I have never encountered such a delay and would be surprised if this were the case.

I must disagree with Dr Marinov that the ball-bearing motor has "no back tension, becaus there are no magnets". Just as there is a variation in the contact potential between two dissimilar metals with temperature (used in thermocouples), so there is a potential gradient established in a single piece of metal as the result of a temperature gradient (Thomson effect). As there is a massive temperature gradient within the ball and the race at their point of contact, there will also be considerable potential gradient giving a back EMF analogous to that produced by a conventional electric motor.

I would be unwilling to dismiss as invalid the law of conservation of energy which has served us well in the past when deriving physical formulae, and would therefore suggest that an error in measurement is responsible for the startling results obtained by Dr Marinov's calorimetry. I am aware that doubts have arisen recently in certain particle physics experiments; whether these have been resolved yet I do now know.

May I suggest that someone with access to a ball-bearing making machine might produce bearings in invar or another low-expansion alloy? One could also try a metal exhibiting large magnetostrictive coefficients, for example cobalt, since it is possible that this effect is responsible for changing the shape of the balls in such a way as to maintain motion.

Peter F. Vaughan
Lynton
Devon

Circuit Symbols

Isn't it a little late in the day for Mr McLoughlin (May, 1989) to beat the BSI with the "rectangular resistor"? BS3939 "Guide for graphical symbols..." was first introduced in 1966 and even then was in conformity with many IEC symbols. The BSI may have thrown caution to the wind 23 years ago but it can hardly be accused of having imposed its will on a cowed industry as the circuit diagrams in this journal and many others demonstrate every month.

It was presumably in the interest of a "truly international language" that BSI and IEC symbols were made to conform and I do not believe that the 'wiggly resistor' was in common use in continental Europe prior to 1966.

I do not have any connection with the BSI, but I was under the impression that circuit symbols, and indeed all other standards were the result of the deliberations of committees of experienced people from the industries concerned. We have all heard about camels and committees but why is it that the 'rectangular resistor' raises such ire? Is it so much more difficult to appreciate its ohmic qualities rather than those of a wiggle? Even today the rectangle is easier to draw using the average computer drawing program which was, I understand, one of the original criteria for its choice all those years ago. All the "intelligently chosen" symbols were, after all, the work of the same committee.

The use of different circuit symbols depends to a large extent upon the market place. Firms doing work for UK government agencies and continental Europe may use one set and those doing work for the USA and Japan use another. Being conformist is necessarily a bad thing. I hope that the pupils at Haberdasher Aske's School conform to the basic rules of English grammar and to a different set of rules for French. I hope we all conform to the relevant speed limit when we drive on the public roads. In 1992 we shall no doubt have to conform in many different ways. There must be more important battles to fight than that of the 'wiggly resistor'.

L.P. Best
Fleet
Aldershot
Hampshire

Astra doubles up

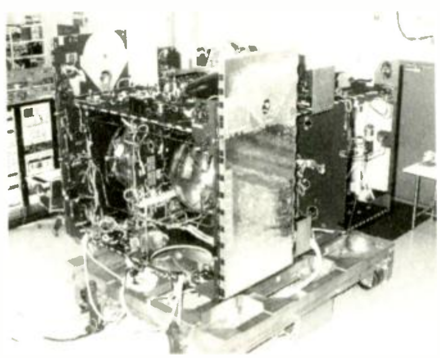
By the early 1990s, Europe will have over 140 satellite TV stations.

Undaunted by a certain coolness in the viewing public's response to satellite television, the Luxembourg company Société Européenne des Satellites is pressing ahead with a second Astra spacecraft. The new spacecraft will be launched as early as next autumn, despite the prospect of a large excess of television channel capacity in Europe's skies.

SES is to take over an existing satellite, Satcom K3, together with its Ariane launch slot in October 1990. Taking the new name of Astra 1B, this will occupy approximately the same orbital station as the existing Astra 1A (an imaginary box 80km wide), bringing viewers the possibility of a further 16 television channels without the need to reposition their dishes or modify their receivers.

Astra 1B is a GE5000 spacecraft, an improved and slightly more powerful version of the GE4000 type chosen for Astra 1A. The craft, which is now at an advanced stage of preparation for launch, needing only to be modified for the Astra channels, was one of several take-over possibilities considered by SES (another was a surplus Canadian Anik C, still unused and looking for a buyer); it meant an earlier launch date than would have been possible with a bespoke satellite, yet had the attraction of full compatibility with Astra's ground-station command system at Betzdorf. With its 60W TWTs (as against 45W for Astra 1A), Astra 1B will give a slightly stronger signal and a larger footprint over Europe, though viewers will be unlikely to notice the difference – they need not even be aware that a second satellite is in operation. It will, however, give SES additional room for manoeuvre in the event of transponder failures.

Who will occupy the extra transponders, and who will be watching the programmes, is rather less clear. By early June, Astra 1A still had four vacant channels, though according to Marcus Bicknell of SES, five German clients were negotiating for them. A further 15 or 16 options had been taken on Astra 1B, among them an Italian soft-porn channel, though not all were likely to mature. However, he said, a glut of transponders would exist in Europe in a couple of years' time, with some 140 channels available: "A lot of blood will be let in the satellite provision industry over the next four years. There will be a lot of hardware up there doing nothing". Among these 140 channels will be those on two recent German craft – the Copernicus satellite successfully launched on 6 June, and the replacement for the first TV-sat, whose solar panels failed to open.



The Astra 1B television satellite, now under construction, will bring an extra 16 channels to dish owners.

According to Astra's marketing folk, the system's big attraction to viewers and advertisers is the "critical mass" of its programme package, made possible by Astra's close channel packing and its placing in the 11GHz fixed satellite service band – provision exists for up to 48 channels, should the as yet undiscussed Astra 1C become a reality. In the 12GHz broadcast band, each country's allocation is limited to just five channels.

Tackling the question of the number of dishes actually in the hands of the public, Bicknell said, "SES refutes entirely that dish sales are disappointing... dish sales have made a very, very healthy start". Quoting a Financial Times survey which revealed that 93 000 had been installed by the end of Astra's first four months, he argued that take-up was twice as fast as that of colour television sets or video recorders in their early days. However, a cynic might observe that colour TVs and VCRs initially cost more than twice what is now being asked for an Astra receiver, and that these had to make their entrance without the benefit of wall-to-wall publicity in News International's stable of mass-circulation newspapers. Despite all this, News International's own Sky Television, whose four English-language channels make it Astra's biggest programme provider, was recently forced to cut its dish sales forecasts for the first year by almost half, from 2M to 1.15M, and privately, seemed to have halved the estimate yet again.

Of the 93 000 dishes, only 60 000 represent special Astra units, the rest having been installed earlier for other satellites. However, since, as Bicknell put it, 75% of consumer electronic goods are sold in the final part of the year, there is still plenty of opportunity for making up lost time.

A further factor in Astra's favour is the postponement of BSB's programme launch until the spring of next year. BSB's problems are more in its technology than in programming or marketing. Among them is the Squarial flat panel antenna, which at the time of going to press had still failed to find a manufacturer and had forced the abandonment of a costly nationwide advertising campaign on the theme "It's smart to be square".

According to David Eglise, the company's technical director, there are actually two Squarials. One is a three-part injection-moulded design, potentially cheap to manufacture but difficult to do so to the fine tolerances necessary at 12GHz. In the other design, which is basically metallic, the problems are transferred from moulding to assembly. BSB was unwilling to give *E&WW* more technical detail until a maker had been signed up.

Despite this hitch, BSB's satellite launch on August 10 goes ahead as planned. However, programmes will begin this year on only one channel, as a showcase for television dealers. The full service, when its starts, will now consist of five channels (the IBA having awarded BSB the remaining two), all transmitted using the D-MAC/packets standard, and with the capability for wide-screen pictures right from the start.

Over at SES, Astra has been fairly free of technical problems – though there was an awkward moment recently when the satellite suffered an Earth-lock fault and flipped its footprint around to point at Greenland for a short time. However, the spacecraft's reserves of hydrazine propellant have been holding up well and its lifetime is now put at as 12.4 years rather than 10. ■

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PIONEERS

Sir Charles Tilston Bright (1832-1888) "The great feat of the century".

W.A. ATHERTON

What the Channel Tunnel is today, in some ways the Atlantic telegraph cable was in the middle of the last century: a visionary engineering enterprise and a link between nations. Samuel Morse called it "the great feat of the century". Once the telegraph traffic started to flow, what some had opposed as undesirable became essential and more Atlantic cables were soon required. At least 17 were laid before the end of the century.

Charles Tilston Bright was Engineer-in-Chief of the Atlantic Telegraph Company, a position he occupied at the age of 24. Imagine a 24-year-old being appointed chief engineer of the Channel Tunnel! Though largely forgotten now he was a famous figure in his day. He was knighted for his achievements at the age of 26, the youngest knight for generations.

Commercial telegraphy began in a small way in 1839 and in the next decade became well established in developed nations. In 1850 England and France were linked by the first submarine telegraph, laid by the brothers Jacob and John Watkins Brett. It soon failed, but on November 13 the next year another cable linked Dover and Calais and saw successful service for 24 years. Demand soon justified other cross-channel links.

After the success of the Channel cables, other short and shallow stretches of sea were crossed, but not always at the first attempt and with the loss of much expensive telegraph cable. London's Thameside spawned a new industry of cable manufacture.

To cross the Atlantic

Meanwhile, a British engineer had held a conversation with an American businessman. The engineer was Frederick Gisborne, who was then constructing a telegraph line from New York to Newfoundland and which was to include a cable across the Gulf of St Lawrence. The businessman was Cyrus W. Field.

Gisborne, who needed fresh financial backing for the Newfoundland project, suggested the idea of continuing the line from Newfoundland across the Atlantic to England: an Atlantic cable. Lieutenant Matthew Maury, a US Navy oceanographer, assured Field that there was a route across the ocean bed where a cable might be laid. With Field's financial and managerial help the New York to Newfoundland telegraph was completed and the cable laid across the Gulf of St Lawrence. For this, Field had to travel to Britain to obtain cable and ships. John Brett, who had laid the Channel cable, became his supporter and adviser.

The idea of an Atlantic cable caught Field's imagination and for the next 12 years he was to be the driving force behind this daring adventure. He enthused leading scientists and engineers (some like Thomson, later Lord Kelvin, taking no pay). A seven-core cable using 20 000 miles of copper wire was constructed, paying out machinery designed, two battleships borrowed, and Charles Bright selected as chief engineer. Within four years of Field's and Gisborne's first conversation, telegraph messages were exchanged between Britain and America. Though success was short-lived because the cable soon failed, they had proved their point. Field had shown the political and commercial potential of the cable and Bright had shown it was technically possible.

The druggist's son

Charles Tilston Bright was the third son of Brailsford and Emma Charlotte Bright. Tilston was his mother's maiden name. His father is described by the Dictionary of National Biography as a "druggist". He was born in Wanstead, London, on June 8, 1832, less than a year after Faraday's discovery of electromagnetic induction and in the same year that Morse had his first thoughts about electric telegraphy.

Though later he was to mix with university professors of world renown there was to be no university education for Bright. At the age of 15 he went

Below: first verse of a poem marking the success of the 1858 cable, from The British Workman (reference 1). That success was, alas, short-lived.

Two mighty lands have shaken hands
Across the deep wide sea;
The world looks forward with new hope
Of better times to be;
For, from our rocky headlands,
Unto the distant West,
Have sped the messages of love
From kind Old England's breast.

Institution of Electrical Engineers

straight from the Merchant Taylors' School into the employment of the Electric Telegraph Company, the company formed to exploit the Cooke and Wheatstone patents. He had joined the oldest telegraph company in existence: it had been formed just two years earlier.

After about five years he moved to the rival Magnetic Telegraph Company. His brother Edward, also an engineer, became manager. In this post, Bright helped to wire up Britain, laying an extensive network of land-based telegraph lines, with thousands of miles of underground wires between major centres including London, Manchester and Liverpool. After becoming Engineer-in-Chief of the company he got his first taste of submarine telegraphy, laying a six-wire cable between Portpatrick in Scotland and Donaghadee in Ireland. Under the Bright brothers, the Magnetic Company prospered.

This cable between Britain and Ireland came after earlier failures. The water was deeper and the currents faster than in previous operations. Bright took charge of the cable-laying machinery. The whole cable was manhandled out of the hold of a steamer, over a pulley, round a drum which measured the speed, and then several times round a brake drum before passing into the sea. It was laid on May 22, 1853, and had a long and successful life. Bright stayed with the Magnetic Company as chief engineer until 1860 and served a further ten years as consultant.

Patents

During his time with the company he received a number of patents concerned with improving telegraph equipment. Two are particularly noteworthy. One awarded in 1855 suggested replacing visual indications at the detector with acoustic ones using two bells, which became known as Bright's Bells. One had a high tone, the other a low one, and they were used on the West Indies network. The other, earlier, patent was awarded jointly to the brothers in 1852 and contains what seems to be the first suggestion of a resistance box for giving a variety of fixed resistance values. This patent covered 24 distinct inventions and marked the arrival of the brothers, and especially Charles, as important figures in telegraph engineering.

A first attempt

With Bright having linked Ireland to Great Britain, and Gisborne and Field having tied Newfoundland to the American mainland, it was natural that all should consider the Atlantic.

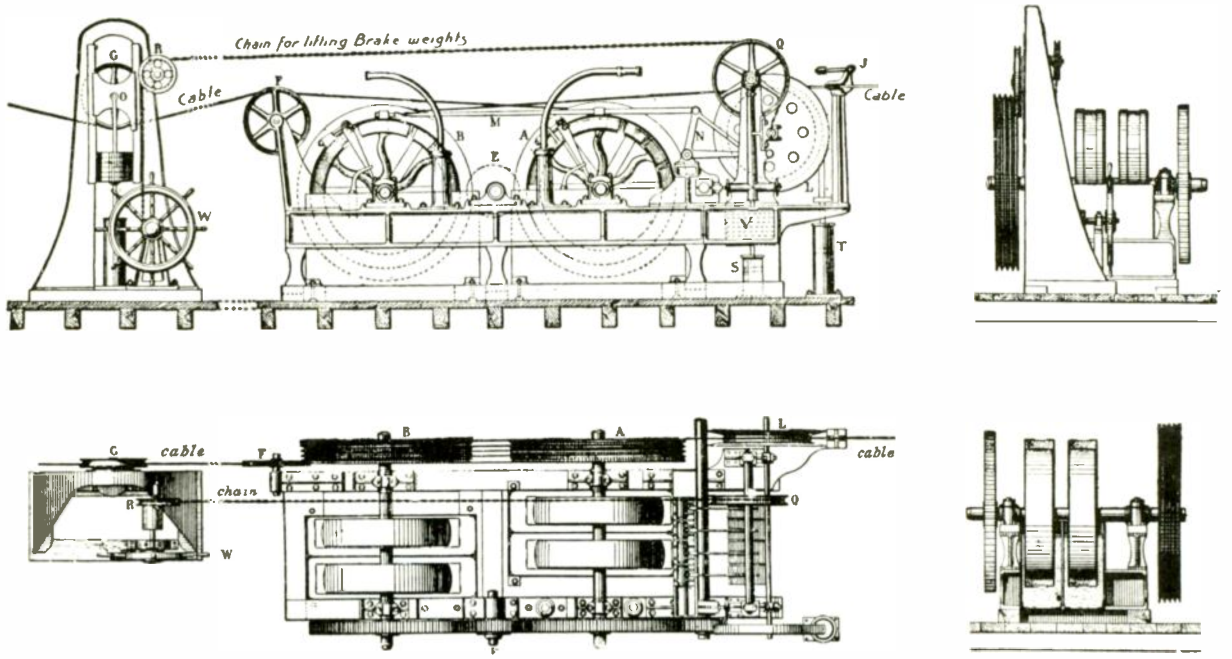
Field was unable to raise the necessary capital in America and so, with Samuel Morse assisting, he sailed for Britain. On September 29, 1856, John Brett, Charles Bright and Cyrus Field pledged themselves to form a telegraph company to operate a telegraph between Ireland and Newfoundland. The company was registered on October 20. Many famous names were associated with it including William Thomson (Kelvin), Isambard Kingdom Brunel and Samuel Morse. The required finance of £350 000 was raised in a fortnight. Bright became chief engineer, and his colleague from earlier ventures, E.O. Whitehouse, was appointed the company's "electrician".

Field's "dynamic energy" pushed the project ahead at breakneck speed. Bright objected to the size of the single conductor and wanted it increased, but was overruled. He also wanted the two ships involved to start laying by splicing the cable in mid-ocean and then sail for opposite shores, but was again overruled. Kelvin wanted to wait for the completion of the *Great Eastern*, which was to be the biggest ship in the world, and he found variations in the quality of the copper core. Field again ignored the warnings: he had ships on loan from the respective navies and wanted to lay the cable in mid-summer when the weather could be expected to behave.

The cable was loaded into the *USS Niagara* and *HMS Agamemnon* and the shore end landed at Valentia in Ireland on August 5, 1857. When laying began the cable broke after only five miles. When it broke again after 380 miles the end was lost and the attempt abandoned. The cable was stored for the winter while additional lengths were manufactured ready for a fresh attempt.

As Bright put it, "It has been proved beyond a doubt that no obstacle exists to prevent our ultimate success; and I see clearly how every difficulty which has presented itself in this voyage can be effectually dealt with in the next". He was a little optimistic.

The original backers put up more



Paying-out machinery for Bright's transatlantic cable of 1858. After only a month, the cable failed.

cash, the two navies agreed further support of ships and men, and the paying out machinery was redesigned. The following summer the fleet reassembled and sailed for mid-Atlantic, Bright's plan now being adopted. The two lengths were spliced and cable laying began again; but again the cable broke after 160 miles and the ships returned to base independently.

Another try

Disappointment brought disagreement. The board chairman recommended abandonment of the project and some members agreed with him. Some resigned. But Field pushed on and won the day. The fleet put to sea again on July 17. The cable was spliced in mid-ocean on July 29 and without further ado it was laid successfully, both ships reaching shore on August 5. During the subsequent celebrations the roof of the New York City Hall was set on fire by fireworks.

Signalling through the cable was never easy and after only one month and 732 messages the insulation failed. Probably several factors were involved, including Whitehouse's use of too high a voltage. But to have achieved any success at all on such a venture when practice was so primitive was remarkable in itself. Though the success was short lived, Bright as chief engineer and Field as chief proponent had proved that telegraphic communication could be achieved across the Atlantic.

Financial backing for a third attempt proved much harder to find. After other expensive submarine cable also failed a British Board of Trade inquiry was set up to look into the technology and methods used. Bright was amongst those consulted. It reported its findings in 1861 and many improvements resulted. Inevitably a new attempt was made, with Bright as consultant to the project. In 1865 the *Great Eastern* laid cable all the way from Ireland to within 600 miles of Newfoundland before it broke. The next year complete success was achieved with yet another new cable and the 1865 cable was grappled, spliced and also completed.

Five years before the 1866 success, Bright had resigned from the Magnetic Telegraph Company to go into business as an independent consultant in partnership with Latimer Clark, another famous engineer of the time. They experimented with the insulation of wires and are remembered for "Bright and Clark's compound", a bituminous sealant used with later cables.

Bright was consultant to many telegraph companies needing major submarine cables including the Anglo-Indian, the Anglo-Mediterranean, the British-Indian Extension, and the China Submarine Telegraph Company. It was he who broke the jinx of failure in the deep waters of the Mediterranean. Also he was instrumental in setting up the British Association committee on electrical standards, on which he served with other distinguished scientists and engineers such as Maxwell and Wheatstone. It was this committee which established electrical units such as the ohm and the farad.

Into Parliament

For three years from 1865 Bright pursued a completely different career, as the Liberal Member of Parliament for Greenwich. Late in his life he was involved in an unsuccessful mining venture in Serbia.

He received honours, naturally. As well as the knighthood from the British government, the French granted Bright membership of the *Légion d'Honneur*. He was a member of the Society of Telegraph Engineers (later the Institution of Electrical Engineers) from its inception and was its president for the year 1886/87.

It was soon after his presidency that Bright died, suddenly, on May 3, 1888 of heart disease at his brother's home in Kent. He was buried in Chiswick churchyard. Though a marble bust was made of him, his lasting memorial is the fact that he was engineer-in-chief of the first transatlantic telegraph cable. He linked the New World with the Old. ■

Further reading

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2. Dictionary of National Biography.
3. V.T. Coates and B. Finn, "A Retrospective Technology Assessment: Submarine Telegraphy, The Transatlantic Cable of 1866," San Francisco Press, 1979.

Next in this series of pioneers of electrical communication: Joseph Henry.

Tony Atherton is author of "From Compass to Computer: A History of Electrical and Electronic Engineering", Macmillan Press, 1984.

Single-bit, oversampling A-to-D converter

Erik Margan developed this simple converter for a battery-powered telemetry transmitter, which needed a signal-to-noise ratio of more than 70dB over a 5kHz bandwidth

Delta A-to-D conversion technique is probably familiar to readers of this journal; most recently, it was described in a Circuit Idea by D.J. Greaves[1]. This type of conversion is attractively simple, needing only a comparator, a D-type flip-flop, a sampling clock and a low-pass filter in the feedback loop (see **Fig. 1(a)**). In contrast to the more commonly used converters which attempt to sample the input signal as accurately as possible at each sample, delta converters make use of the oversampling technique, allowing the error to be arbitrarily large at each sample and reducing the error by averaging the samples at the output. The bit stream which is produced by clocking the flip-flop is integrated by the low-pass filter and fed back as error signal for the comparator.

This configuration has a few drawbacks, the most serious being the degraded signal-to-noise ratio with frequency, which follows the inverse characteristic of the feedback filter. The cause of this degradation lies in the progressively smaller error correction in the feedback loop which, in turn, causes slew-rate limiting and thus smaller undistorted output amplitudes at higher frequencies (**Fig. 1(b)**).

Improving s:n ratio

A rearrangement of the feedback configuration leads to another type of the delta converter, known in theory as the sigma-delta converter, but rarely used (**Fig. 2(a)**). Such a circuit has been recently described in an excellent article by R.W. Adams[2]. The difference is that the input signal is first summed with the output of the flip-flop and then the sum is filtered and compared to a DC reference. The consequence of this change is that the overload level is now flat with frequency (in the band of interest), while the noise spectrum rises with frequency, producing about the same s:n ratio (**Fig. 2(b)**).

This does not seem to be an improvement, but putting the filter after the summing stage gives us another degree

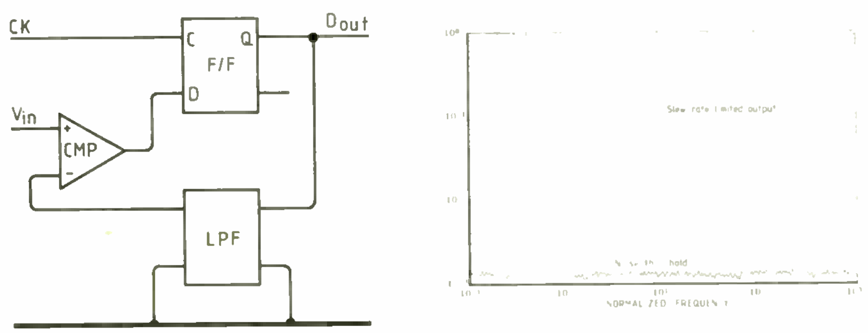


Fig. 1 The Delta-type converter circuit at (a) compares the input signal with the filtered output big stream. When following high-amplitude signals of frequency greater than the filter cut-off, the converter runs into slew-rate limiting, degrading the S/N ratio, as seen at (b).

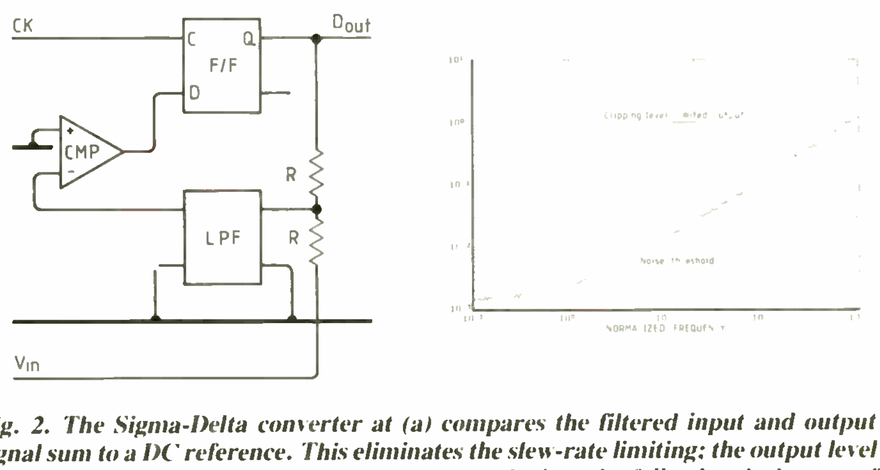
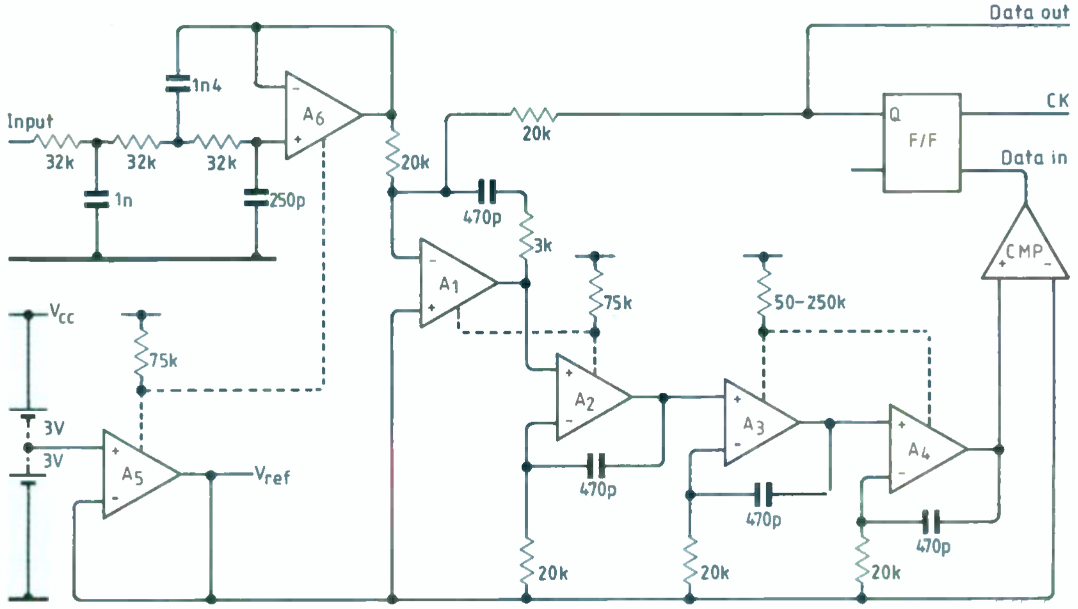


Fig. 2. The Sigma-Delta converter at (a) compares the filtered input and output signal sum to a DC reference. This eliminates the slew-rate limiting; the output level is not affected by the filter frequency response, only the noise following the inverse of the filter response shape, as in (b).



of freedom. We note that the frequency response is now flat, regardless of the filter response shape, while only the noise spectrum follows the inverse of the filter shape. From this we conclude that by making a suitable choice of filter, we can concentrate most of the noise outside the band of interest and filter it out after the conversion has been performed.

In this way, the s:n ratio can be greatly improved, but this is not an easy task. There are several points to take into account simultaneously:

- selecting optimum filter configuration to achieve required noise shaping;
- reducing the influence of the choice of sampling clock frequency on the noise spectrum;
- maintaining the stability of the closed feedback loop in the broad range of sampling clock frequencies; and
- optimising the idling pattern of the converter for low distortion (the quantisation process should introduce noise only, independently of the signal level or slope).

Each of these requirements introduces some restriction in other areas, so the system must be treated as a whole. The first two points restrict the choice of filter configuration to those with zeros and poles, implying a pronounced attenuation well above the band of interest. Circuit complexity imposes another restriction and the last two requirements are best served if we implement some kind of slew-rate control.

Fig. 3. The proposed circuit diagram of the Sigma-Delta type. Using the programmable c-mos op-amps and comparator (MC 14573 and Mc14575) allows us to control the stability of the feedback loop through controlling the bias and the slew-rate (a non-linear mechanism!) of the last two filter sections.

Adams[2] states in his article that single-bit converters are difficult to stabilize, and for that and other reasons he opted for a four-bit configuration. Using a similar filter configuration, but with four zeros and four poles, I came to the conclusion that the required feedback stability can be achieved through slew-rate control of filter amplifiers.

Design

The whole circuit is shown in **Fig. 3.** The programmable operational amplifiers A1, A2, A5 and A6 are contained in an MC14573, while A3, A4 and the comparator (CMP) are contained in an MC14575. The D-type flip-flop (F/F) is one half of a 74HC74. The signal first passes through the input anti-aliasing filter (A6 – a third-order Bessel type was found to be enough) and proceeds to the inverting integrator A1 where the input and output are summed at the virtual-earth point. This eliminates a separate summing stage. The resistor in series with the integrating capacitor sets the transfer-function zero location, and is selected so as not to saturate the comparator's input even under the lowest sampling frequency and largest signal level. The other three zeros are made equal to the first through the use of non-inverting integrators (A2, A3, A4) for the remaining filter sections, with equal RC constants defining the pole position. A further control of the filter is available through the bias control of the programmable operational amplifiers that also influences the slew rate of the signal at the comparator input. Slew-rate limiting is implemented only in the last two filter stages and is adjusted for low distortion and a stable feedback loop at the lowest sampling frequency.

As stated before, the filter is the key to the circuit. We have already stated the main requirements, but now is the time to justify the filter choice given in Fig. 3. First, as the filter is used inside the feedback loop, at least one of its stages should be configured as an integrator to ensure the stability of the loop DC level and to make it equal to the reference. This, in turn, ensures equal positive and negative clipping levels and maximises the signal dynamic range. Next, the loop must have a flat frequency response to well above the desired bandwidth to make the noise spectrum flat inside this band, thus achieving constant s:n ratio. To concentrate the noise in the upper frequency band, the loop gain must be intentionally lowered in this region. This attenuation reduces the loop error correction and, being randomly distributed, the resulting error appears as high-frequency noise.

One can hardly resist a temptation to make this attenuation as deep and as narrow as possible and this, as correctly assumed, improves the in-band noise reduction. The price to pay, however, is that the loop self-oscillating frequency becomes better defined, making it harder for the loop to rebalance the changes introduced by the input signal. The consequence is that some kind of "bit hysteresis" results, producing a very distorted signal. In fact, looking more closely at how a "bit" is defined, we can see that the loop rebalances to the new signal level by changing the phase (and frequency) of the self-oscillating waveform; thus, for a smooth bit-to-bit transition, a low-Q notch is necessary.

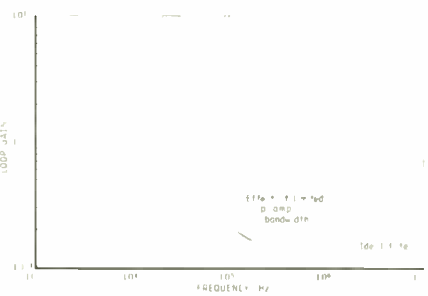


Fig. 4. The analysis of the filter continuous-time loop frequency response. Eliminating the comparator and the flip-flop and closing the filter feedback loop results in these response curves. Finite open-loop bandwidth of the filter op-amps modifies the response around 1MHz. When discrete time steps are introduced in the loop (by putting back the comparator and the flip-flop) most of the error will be produced in the response trough (100-500kHz).

As can be seen from **Fig. 4**, the transition from a flat response to attenuation is not smooth. The pronounced resonance peak results from the four poles of the filter being present inside the feedback loop. This peak gives a very high loop gain and reduces the noise in this band to a very low level, but this is not its main function: it is needed mainly for the high phase jump produced at the resonance, which provides an effective boundary that prevents the loop self-oscillations from breaking into the desired signal-frequency band.

Finally, the filter zeros determine the frequency response at the highest frequencies, but it is altered somewhat by the limited op-amp bandwidth (the small second peak).

Now, this continuous-time loop-analysis result is not greatly affected by the introduction of discrete time steps (the sampling clock frequency) if the steps are small enough (less than 2μs). As the feedback factor is lowest in the response trough (100-500 kHz) most of the noise will be concentrated in that band. The self-oscillating frequency is not fixed, but varies from about 120 to 470 kHz and the period is incremented or decremented in discrete time steps ("phase noise"), the width of the step being determined by the sampling clock frequency (**Fig. 5**).

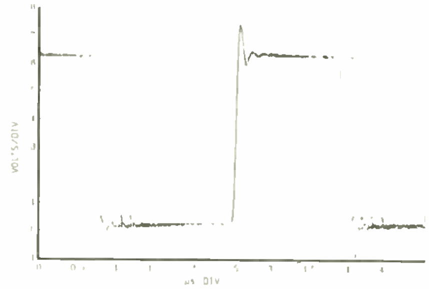


Fig. 5. The flip-flop output with no input signal. Sampling frequency 5MHz. The output changes state randomly between discrete widths, the steps being determined by the sampling frequency. Recorded with 50% pretrigger.

The effect of slew-rate limiting is not made obvious from Fig. 4. To understand it, we must visualise the time pattern of the signal at the comparator's input. The switching of the flip-flop output is transferred through the filter by the transfer-function zeros, appropriately attenuated. This is the fastest part of the waveform and slew-rate limiting will modify its slope. This slower slope introduces a delay proportional to the error produced at the current sample, the slower part of the waveform, which is due to error integration, being passed unaffected. Bear in mind that the output bit stream is undergoing constant integration, so with no input signal the average width of the high logic state equals the average width of the low logic state. If the error at the current sample is large, it will be compensated to a large extent immediately with the next opposite state width, to which the delay introduced by the slew rate limit will contribute. In this way, we have gained a fast bit-to-bit error correction, transferring the noise spectrum to higher frequencies. Fast error correction contributes also the loop stability at higher frequencies.

To evaluate the s:n performance and the noise spectrum shaping effect, the output bit stream was sent to two filters: a single pole 1kHz filter, used to evaluate the noise shaping and a 7-pole, 5kHz Bessel filter, used to evaluate the s:n ratio and signal distortion in the 5kHz bandwidth. The Bessel-type filter was used to preserve the shape of the input impulse waveform that my application required, but for other purposes a 5th-order Butterworth or Chebychev filter type would give much lower noise in the 5-15kHz band.

The outputs from the filters were recorded and transferred to an IBM PC AT machine for further analysis. The 1kHz filter output is shown in **Fig. 6** representing the first 500 samples of a 4096-byte sequence. Such a long record was needed to enhance the resolution of the spectrum at lower frequencies (as the spectrum low-frequency limit is defined by the length of the time window). The sequence was broken into 512-byte packets, Hanning-windowed, FFTed, and put together again to produce the 256-point intensity spectrum displayed in **Fig. 7**. The smooth line above the spectrum shows the maximum noise level estimation, corrected by the inverse of the 1kHz filter response.

The s:n ratio achieved inside the required 5kHz bandwidth can be judged from **Fig. 8**, where a 20mV p-p sine wave was presented at the converter input and the output signal from the 5kHz 7-pole filter was recorded. Sampling frequency was 5MHz. Here we can see that the noise peak-to-peak level is about 1mV. If we consider the fact that the system supply voltage is 6V

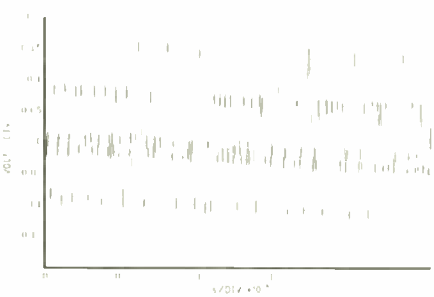


Fig. 6. Passing the output bit stream through the 1kHz single-pole filter to evaluate noise performance. The randomness of the "bit" pattern is clearly displayed. Worst case condition – 500kHz sampling frequency.

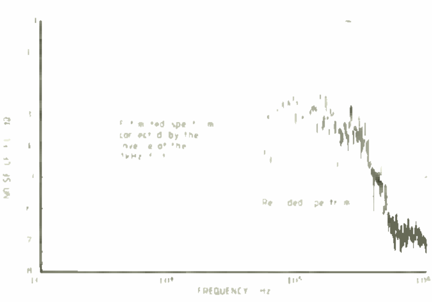


Fig. 7. Noise spectrum estimation and the resulting spectrum of the signal from Fig. 6. The effect of concentrating the noise outside the required signal frequency band (5kHz) is evident.

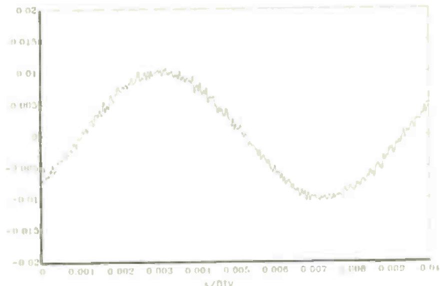


Fig. 8. A 20mV$_{p\text{-}p}$, 120Hz signal was presented to the converter and the output bit stream fed to the 7-pole, 5kHz Bessel filter produced this figure. Sampling frequency was 5 MHz. Noise level is about 1mV$_{p\text{-}p}$. Compared to the maximum undistorted signal level of 5V$_{p\text{-}p}$ gives 74 dB s:n ratio.

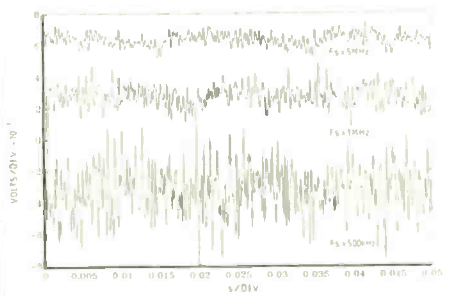


Fig. 9. Output bit stream filtered by the 7-pole, 5kHz Bessel filter, with converter input tied to the DC voltage reference. The influence of the sampling frequency to output noise is shown.

(two 3V lithium cells) and if we assume maximum undistorted output of 5V p-p, we arrive at the peak-to-peak voltage ratio of 5000:1 or 74 dB. If we take into account that the noise peak-to-average ratio is much greater than the sine-wave peak-to-average ratio, we can add some 4-6 dB, arriving at nearly 80 dB.

Further improvement in s:n performance can be achieved either by raising the supply voltage, and thus maximum input and output signal levels, or by increasing the sampling clock frequency. Unfortunately, if you insist on battery supply, this results in either prohibitive power drain, or space and weight. However, even if you renounce the battery supply you are faced with the 18V supply limit of the 4013 c-mos flip-flop used in place of the 74HC74. A CA3080 transconductance amplifier can be used instead of the feedback summing resistor, as well as in the place of MC14573, and a PMI's CMP-01 in the place of the MC14575 comparator stage, thus arriving at about 25V p-p signal swing and resulting in more than 86 dB dynamic range. On the other hand, there is little benefit if the sampling frequency is raised beyond 5MHz, since the noise at the input of the comparator and the filter-stage amplifiers determinate the final result. A better approach is to improve resolution by using more comparators (a flash converter) and digital filtering as in Adams' article[2].

Finally, in **Fig. 9**, we can compare the effect of changing the sampling frequency on the noise. A 3:1 improvement is achieved by raising the sampling frequency from 500kHz to 1MHz and a further 2:1 improvement results from increasing it to 5MHz.

A single-bit converter with flat overload level and nearly flat noise spectrum more than 70 dB below overload in the 5kHz bandwidth is, as has been shown, realisable and stable and the possibility exists of varying the sampling clock frequency in one decade-wide range. With a 1MHz sampling frequency, the s:n ratio is roughly the same as in the standard 12-bit, successive-approximation A-to-D converter system with 15kHz sampling frequency. If there is a need to convert the received data back to analogue form, a simple analogue filter will do the job. If a digital format is needed, the received bit stream can easily digitally be filtered and resampled with the required resolution at a lower rate, single-bit filtering being much simpler than 12-bit, and will easily run in real time in most cases.

The author feels greatly indebted to the excellent article by R.W. Adams, as it triggered my curiosity and offered a good guideline to this successful design.

References

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2. Design and implementation of an audio 18-bit analog-to-digital converter using oversampling techniques, R.W. Adams *Journal of the Audio Engineering Society*, vol. 34, no. 3, March 1986, p. 153-166 ■

Erik Margan works on nuclear magnetic resonance, liquid crystals and electro-optics at the Jozef Stefan Institute in Ljubljana, Yugoslavia

Useful Network Theorems, fifth edition, by Dr Harry E. Stockman. Comprehensive collection of theorems, with historical notes on their origin, applications and a set of worked problems. The theorems, some of which are contributed by the author, include those dealing with the periodic steady state and those for transient analysis. Appendices provide guidance on methods of calculation, tabular information, definitions and short biographical notes on people who have made notable contributions to network analysis. The book is in typescript form with diagrams drawn and annotated by hand; on occasion, this makes it difficult to read, since the printing quality is also rather poor. Sercolab, Box 767, E. Dennis MA. 02641, USA. 160 pages, soft covers, $11.50.

Advanced BASIC scientific subroutines, by B.V. Cordingley and D.J. Chamund. Extensive collections of mathematical and statistical subroutines, written in a structured form of Basic which should run without significant modification on IBM PCs and various other computers. Subject groupings include the generation of random numbers; probability, density and distribution functions; analysis of variance; matrix operations; interpolation; numerical analysis (Chebyshev polynomials and Fourier series); calculus; solution of equations, and complex numbers. The routines are liberally commented. Macmillan Education, soft covers, 178 pages, £9.50. The software is also available from the publishers on a disk suitable for the IBM PC.

Taming Technology: how to manage a development project by Geoff Vincent (senior consultant with PA Technology). Concise, readable, practical guide to improving one's competitive position, for both engineers and managers. Sections cover the development cycle, from concept to product launch; projects and the project leader; the need to plan, and the improved planning methods made possible by the microcomputer; creating an effective organization; and estimating cost and time. The author states persuasively the case for a scientific approach to project management, as distinct from simply playing it by ear. Kogan Page, in association with the British Institute of Management, 173 pages, hard covers, £14.95.

Electronics Sourcebook for technicians and engineers, edited by Milton Kaufman and Arthur H. Seidman. Generously-filled crib (it's nearly 45mm thick) for technicians and students, describing the properties and uses of electronic components, circuit elements and instruments of all kinds. Entries include a little theory, a lot of useful advice and, where appropriate, a scattering of worked examples. This edition is a condensed, soft-cover version of the same publisher's Handbook of Electronics Engineering Technicians. McGraw-Hill, £19.95.

Analogue/digital divider

Circuits for digitally-programmable gain control are quite common now but this one is different in that it accepts serial digital input.

Serial-input d-to-a converters like the PMI DAC-8143 shown here are particularly useful in multi-converter applications where access speed is not of primary importance. They have cascading outputs and so one address-decoder line can control any number of converters. Another advantage of serial control is that many converters can be fed by only three lines, namely for the serial data, clock and load signals.

As with any such serial system though, there are two disadvantages that affect access speed. Firstly, serial data takes longer to transmit than parallel data and secondly, to access any device in a cascaded serial chain you need to re-send data to all the devices; to update one eight-bit converter in a four-converter chain, for example, you have to send out a 32bit serial word.

The 8143 is a 12bit c-mos d-to-a converter designed specifically for serial-data multiple-converter applications. Its data sheet shows how devices are strung together, how they are interfaced to microprocessors, and how the analogue side operates.

In multiply mode, the 8143's transfer function is

$$V_o = -V_{in}\left(\frac{A_1}{2^1}+\frac{A_2}{2^2}+\frac{A_3}{2^3}+\dots\frac{A_{12}}{2^{12}}\right)$$

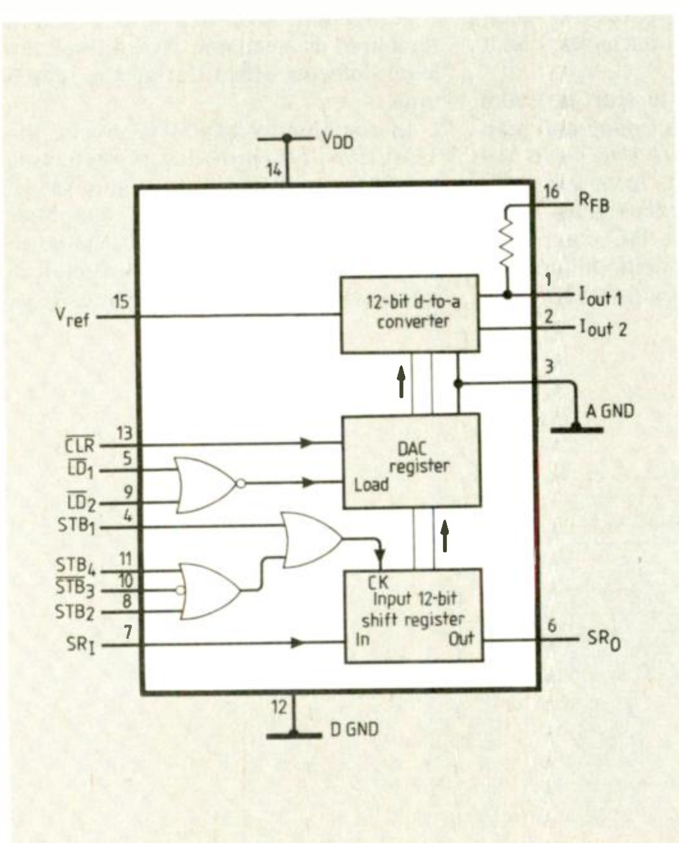
where A_x assumes a value of one for an on bit and zero for an off bit.

When the d-to-a converter is placed in the feedback loop of an op-amp, its transfer function is modified to become

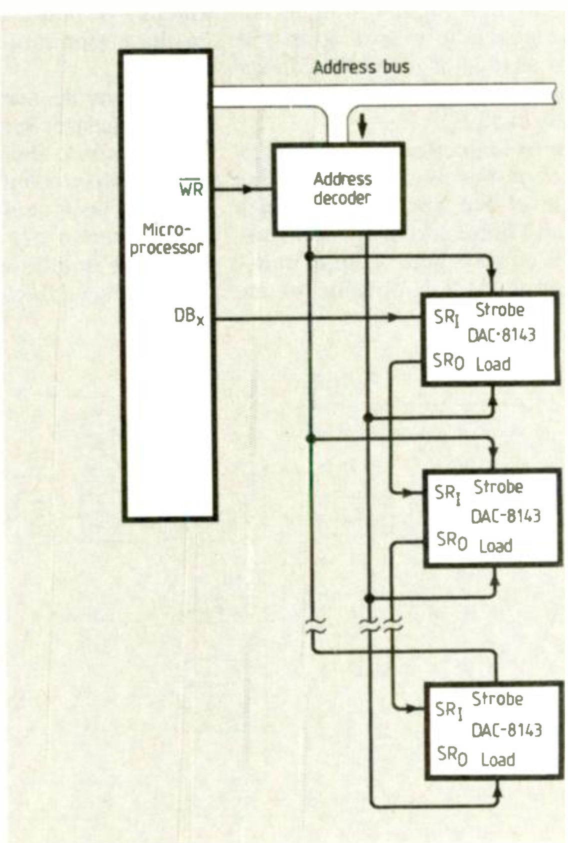
$$V_o = \left(\frac{-V_{in}}{\frac{A_1}{2^1}+\frac{A_2}{2^2}+\frac{A_3}{2^3}+\dots\frac{A_{12}}{2^{12}}}\right).$$

This transfer function is the division of an analogue word at the reference input by a digital word. When all bits are off, the amplifier output goes to the rails since gain, at divide-by-zero, is infinity. When the least-significant bit is one, gain is 4096 and when all bits are one, gain is one.

PMI, 90 Park Street, Camberley, Surrey GU15 3NY; 0276 692392.



Block diagram: DAC-8143



DAC-8143 micro interface

AM/FM radio chip-set with high selectivity and sensitivity

A new chip-set from Philips is designed for high-fidelity AM/FM tuners and radios.

Keyed AGC for the RF preamplifier, determined by narrowband and wideband AGC components, is provided by TDA1574T integrated FM tuner. The tuner also contains a linear IF buffer amplifier to compensate for insertion loss of the IF filters, improving the signal-to-noise ratio.

At the heart of the chip-set is the TDA1596T IF amplifier/demodulator, which not only provides IF limiting and demodulation but also static and dynamic mute control of the audio signal levels. Further, this mute control provides a buffered control voltage for the stereo blend control in the TDA1598T stereo decoder and narrowband AGC in the TDA1574T FM front-end.

If the RF input signal is distorted by external interference, fading or multipath signal reflections, the dynamic mute system reacts rapidly to maintain an undistorted audio output from the chip-set for as long as possible. Effects of external interference are, therefore, substantially damped.

Mute control circuitry sees the audio signal level not only as a function of the IF signal level but also of the signal de-tuning and noise and distortion content of the audio signal. These parameters control the tap position of an electronic potentiometer in the mute attenuator via a diode network, so that the FM chip-set reacts quickly to changes in the IF signal or interference levels.

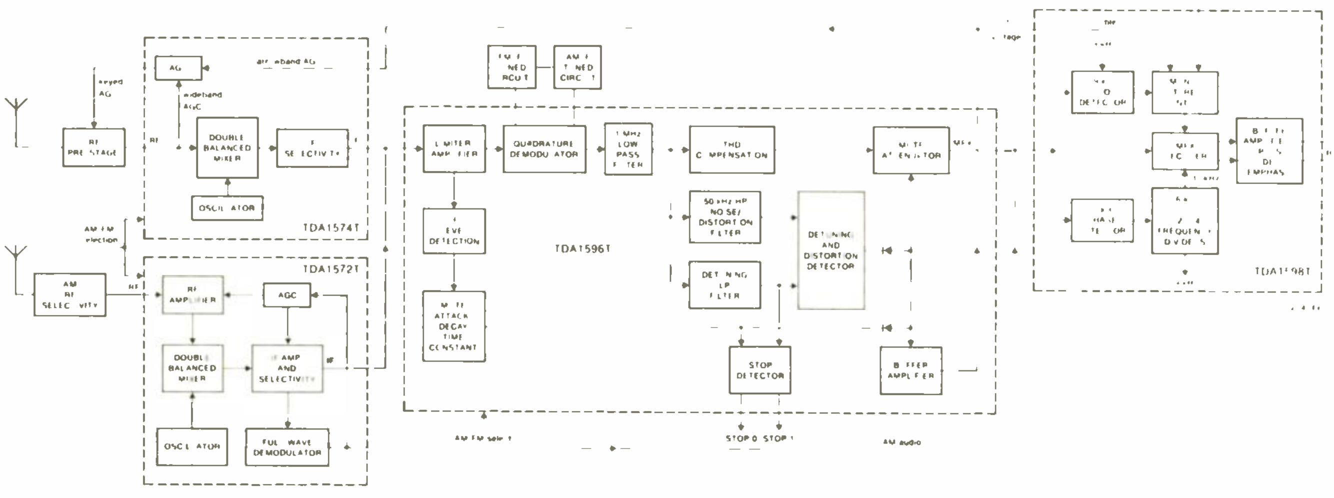
The time constant with which the mute control responds to amplitude variations in the IF signal level is determined by an external capacitor and an externally-applied voltage to provide either equal attack and decay times or fast attack but slow decay time.

Demodulation of the FM IF signal using a quadrature demodulator normally results in a total harmonic distortion of 0.6% (a single tuned circuit with a Q-factor of 20). The total harmonic distortion is reduced still further by a built-in THD compensation circuit to less than 0.3%. The THD compensation circuit has a frequency response characteristic which is equal and opposite to that of the quadrature demodulator tuned circuit. Addition of both frequency response characteristics results in the minimum total harmonic distortion.

To provide accurate stop detection for synthesizer search tuning and scanning control, the TDA1596T has two open-collector outputs from a stop detector. The two reference signals for the stop detector are the DC component from the quadrature demodulator and the IF level detector output. These two signals are fed to comparators and logic gates to provide a basic search threshold of ±18kHz.

A second level-dependent search threshold is available by shifting the level detector offset during the search time.

In combined AM/FM receivers, the TDA1596T also provides stop detection for AM signals. In this case, the AM IF signal is fed to the limiter-amplifier input, a second tuned circuit (tuned to the AM IF frequency) is connected in series with FM IF tuned circuit of the



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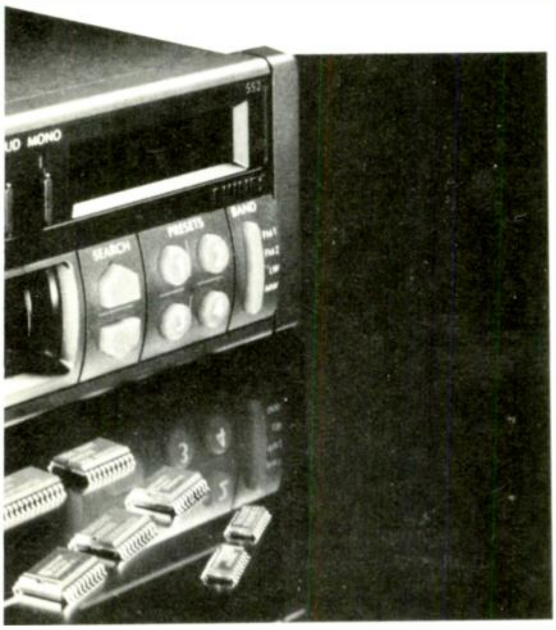
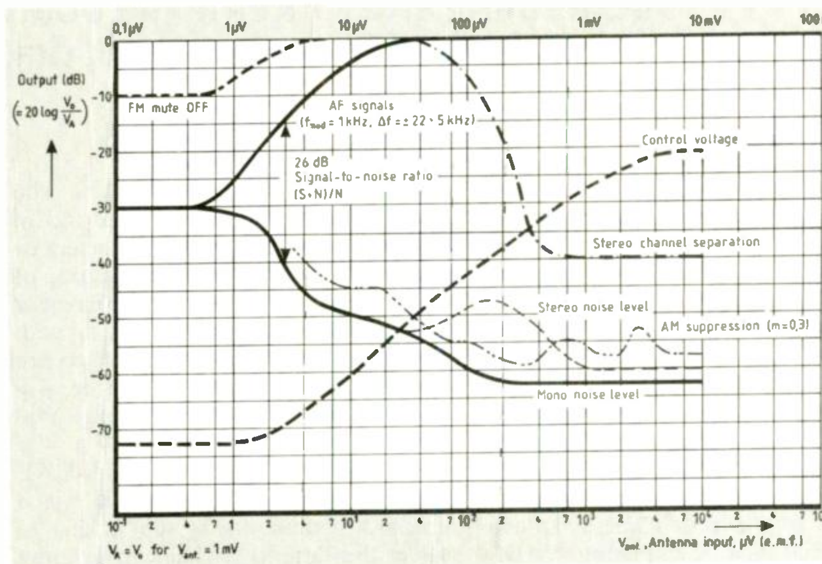




FM	
S-to-n ratio	26dB @ 2.6µV
	60dB @ 1mV
AM suppression	55dB
THD	0.3%
Stereo separation	40dB @ ≥1mV
AM	
S-to-n ratio	26dB @ 15µV
THD	~.5%

quadrature demodulator and the mode switch is set to the FM-off position. This blocks the FM signal path and the distortion detector is disabled, but all the remaining functions of the TDA1596T are still operational to generate stop detection outputs for AM signals.

The TDA1598T time-multiplex stereo decoder is matched to its signal source (the TDA1596T's multiplex output) by selection of an external resistor. Audio gain is determined by the feedback resistors for the output amplifiers. The TDA1598T left and right channel output signals are typically 0.75V RMS.

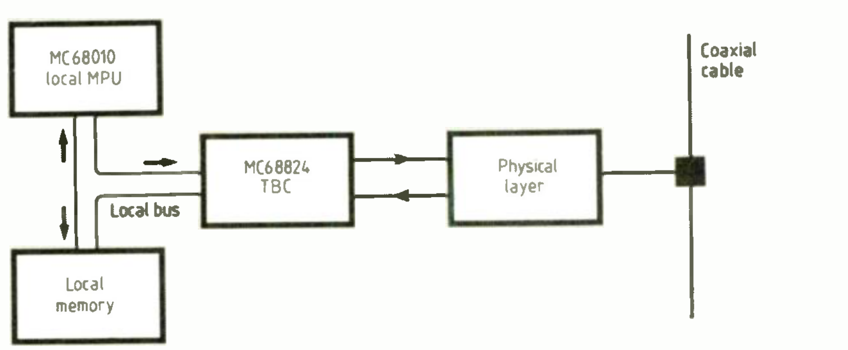
Stereo blend, using the control voltage from the TDA1596T, depends on the IF signal level as well as the interference and noise content of the multiplex signal. The starting level and the slope of this automatic changeover from stereo to mono depend on external resistors.

The TDA1072AT AM receiver circuit includes all the functions necessary for AM reception. The oscillator, especially designed for diode-tuning, contains internal amplitude control to provide a constant oscillator output voltage of 130mV over the frequency range from 0.6 to 40MHz.

In addition to the functions provided by the TDA1072AT, the TDA1572T AM receiver provides an amplitude-controlled IF output for AM stereo operation. The low oscillator phase noise of both the TDA1072AT and TDA1572T ensures that they are well-suited for AM stereo reception. External components needed for the complete chip-set are reduced to those necessary for frequency selection and time constant definition.

Token-bus controller interface

There is an application note written specifically for designers of 68000-based systems interested in networking to token bus. Produced by Motorola, the note details hardware requirements and performance considerations for interfacing the MC68824 token-bus controller to the 68010. *Motorola, Macro Marketing, Burnham Lane, Slough.*



Pads Superstation

An automated PCB design package, generating and routeing the board from data derived from the circuit diagram

With this package, the level of sophistication increases and approaches probably about the most complex a PC would comfortably handle. Automation is the name of the game here, with the ability to generate and route the PCB automatically from information derived from the schematic drawing. This is known as 'logic capture', generally more at home on mainframe and fast workstations than the PC.

PADS Superstation came as an evaluation package, comprising PADS-CAE (for schematic drawing/logic capture) and PADS/PCB (for PCB generation). A further package, PADS Superrouter is available but was not supplied – this adds the ability to guarantee 100% routeing on a board, although this will tie up the PC for, typically, 18-20 hours.

Installation of both programs was via a custom-install program, with the software supplied on 1.2Mbyte AT-style discs. A comprehensive overview manual guides you through the procedure. Requirements are any of the PC XT/AT/386 range with 7 Mbyte of free hard-disc space. An EGA/VGA colour-graphics card is essential. Due to the complexity of the package, you really need a 386 or very fast 286 PC for best results.

The evaluation version of PADS/CAE is apparently derived from the normal software but with reduced facilities – for instance, you cannot save newly created parts or print any output. Nor is it possible to transfer the information generated from the schematic to the PADS-PCB software. However, a number of evaluation files are provided and used in conjunction with the tutorial/guide enable you to get a good idea of the facilities available.

The tutorials for both PADS-CAE and PADS-PCB are comprehensive but not easy to comprehend – they are written for a USA audience and could do with changes to suit the UK. The style is sometimes patronising and in places the instructions are incorrect or confusing, with a lack of explanation of terms or reasons for taking a particular action. This caused some problems with the review, and for anyone who has not used logic-capture software before, will distinctly detract from the package evaluation.

PADS-CAE. Taking PADS-CAE first, entry to the program is via a graphical menu, using the mouse to select the desired package from those installed. A batch file then calls the relevant software.

The maximum sheet size for a PADS-CAE schematic is 60 × 60in – you can use any size below this and have as many sheets as are necessary to hold the complete circuit. Placing parts is relatively simple, using menus located to the left of the entry screen. Parts are taken from the library by typing their name, after which is the identifying description it will be known by throughout the rest of the process – an example might be 'R 1/4W 100k 5%'.

The software tries to automate as much of the process of design as possible. One major example is the assignment or entry or Reference Designations used by the database for keeping track of parts, for example, R1/C1/L1 etc. On most logic-capture systems, this is a manual process, with the operator needing to enter these at the keyboard, and keep track of them across the various sheets of the drawing. PADS-CAE does this automatically at the time you place the part on the drawing – the first resistor you enter becomes R1, or in the example above, 'R1 1/4W 100k 5%'. Similarly, gate and pin assignments are totally automatic. Parts can be moved, deleted, rotated, duplicated and have their identities changed (R1 to R20, for instance) at will. There are also facilities for creating multiple parts such as decoupling capacitors without the need for entering each one individually.

Once all parts are on the schematic, you add the conductors using similar editing facilities to EASY-PC. While you do this, each conductor is assigned a signal name in the database using the format $$nn – you can view this at any time using the F8 Identify key. If preferred, you can change the automatic assignment to one of your own choice. If you inadvertently identify two with the same name, both conductors appear in red on the display, indicating an error condition, otherwise conductors appear in white. There is also a facility to start and stop conductors on other conductors using Tie Dots, and to carry conductors across several sheets of the schematic. At all times, error checking shows whether you have left conductors unconnected.

Bus lines for power, ground and other major connections are drawn separately, being assigned part names in the database such as VCC or GND. If you need to, blocks of the schematic can be moved or copied around the drawing or to other sheets.

Since PADS-CAE holds all the information on the sheets in memory, you can obtain instant reports at any time on the state of your drawing. The reports facility, directed to any matrix printer, include Signals, Buses, Parts, Decals and even the Library. Although not available in the evaluation, the full version allows you to plot the schematic to suitable plotters.

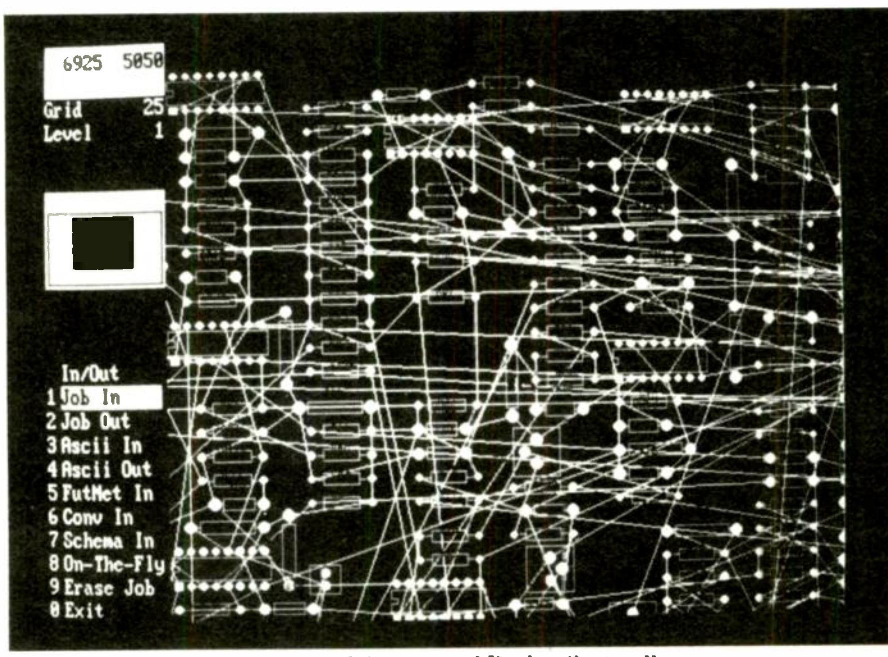
Once the schematic is generated and saved, the next stage prepares the data for forwarding to the PADS-PCB package. The first stage is creation of a database for the PCB software using a process called Initial Forward Annotation. This stage is usually fairly complex in most CAE systems, involving batching all sheets of the design together, then checking across the sheets for errors and their subsequent correction. PADS-CAE generates the Forward Annotation file immediately, since the error checking has already been carried out and the remainder of the information is in memory, even across multiple sheets. The output format can be in either PADS-PCB format, or FUTURE NET for other board-design systems using this format.

The required netlist is generated from an archive file of the drawing. After doing this, PADS-CAE's unique ability comes into play, allowing the user to assign track widths to conductors for use by the PCB package. The list of connections is presented on the screen, from where you can select specific track widths for conductors. You can default all widths (say to 12 mils), then alter any other you wish, such as making V_{cc} 50 mils. Once this stage is finished, your schematic generation is complete and you are ready to move the information over to the PCB generating package.

To be of any use, a logic-capture program must be able to take care of the inevitable changes to schematics, known as Engineering Change Orders. PADS-CAE has full ECO features using a batch file process with little input needed from the operator It can also perform Backward Annotation, whereby changes made at the PCB design stage, for instance pin swapping, can be fed back to the CAE program to correct the schematic.

The Parts Library supplied with PADS-CAE contains around 3000 components. New parts are quite easy to create, although you cannot save them in the evaluation. A 'part' consists of a Decal (the graphical display ou see on the screen) and the Part Type Description, used to combine the decal and electrical information into a single entity.

This part of the package seems to work very well. I found the mouse interface much easier to use than that of EASY-PC, with equally fast screen updates and a logical structure to operations. It would have helped considerably if some form of output facility for the drawings were available; perhaps a limited number of plots rather than none at all. Also, I would like to have seen support offered for draft matrix or laser printer output of drawings.



Working with PADS ("Personal Automated Design System").

PADS-PCB. Moving on to the PADS-PCB package, one of the first things that intrigued me was whether it could generate a PCB in the same manner as EASY-PC does, that is by hand drawing. The answer is yes, and PADS-PCB describes this as 'designing-on-the-fly'. The process is more complicated than simple hand drawing, since you have to tell the software about each part you wish to place on the board before moving on to the layout stage. However, it is really designed as an automated package and should be treated as such.

The working screen is consistent with that of the schematic entry and takes little getting used to. The largest board size is 32.7 × 32.7in or anything smaller, with up to 30 layers. The latter are represented on screen in different colours, user-definable from the standard 16 EGA colours via a set-up menu together with other parameters.

Since you cannot import any files directly from the CAE package in the evaluation, a number of example files are provided to allow use. This could be regarded as a negative point – it does not allow you to try out the package on boards of your own design in typical situations and see how it copes with problems that may be specific to your products. Also, I suppose one could argue that the supplied files are designed to show the package in a favourable light.

The first step in designing a PCB from the database generated by the CAE package is placement of the components. This is slightly different from the way you design a manual board, which is usually done in sequences of placing some components, routeing them, placing more and so on. That way, you continually see how much space is available for tracks.

With PADS/PCB, bringing the CAE generated file into memory, via the Job In command, displays the board outline (already defined in the sample file) plus all the components grouped together around the periphery of the board. Many of these lie on top of each other, but all have every one of their logical connections shown as grey lines. At this stage, all the components except the ICs are 'glued' to the board and will have to be unglued later before placing.

There are two ways of placing components – interactive and auto. With interactive placing, you tell the system to place the ICs on a predefined matrix, then rotate and move them with the mouse in conjunction with the Net Length feature. The latter tells you how long the total connection lengths (or Nets) are, and with the Length Minimisation option allows the program to reduce net lengths automatically by reordering the nodes in each net. With this done, you can move ICs and other components around with the mouse. As you do so, the logical connections move with it, letting you get some idea of the best placement for individual parts so that, for instance, ICs with many connections lie next to each other. If necessary, say with SMD designs, you can move components to the other side of the board.

The process continues with minimising V_{cc} and GND connection lengths, and placing the discrete components such as decoupling capacitors next to ICs. If youi need to, components such as resistors can be stood on end and/or rotated at any time.

With automatic placement, you take a slightly different approach. Since this process attempts to place every component on the board in a suitable place on the matrix with as short connecting lengths as possible, you first have to glue down those components with fixed positions (such as connectors, switches, etc.) in the right place. Having done this and made sure everything else is unglued, selecting Autoplace starts the automatic placement routine, initially for the ICs. The sample file is supposed to be placed within a minute, according to the guide, but actually took about 20 seconds, so the timings are probably based on something like an 8MHz AT rather than a 386.

This first run will not necessarily be perfect and other options help you improve on this. They mainly concern SWAP operations, such as interchanging IC pairs, gates and pins until no further improvement in connection length results. Any changes are saved and can be used by the CAE program to update the original schematic.

With the ICs placed, the placing of the decoupling capacitors follows, then the discrete components.

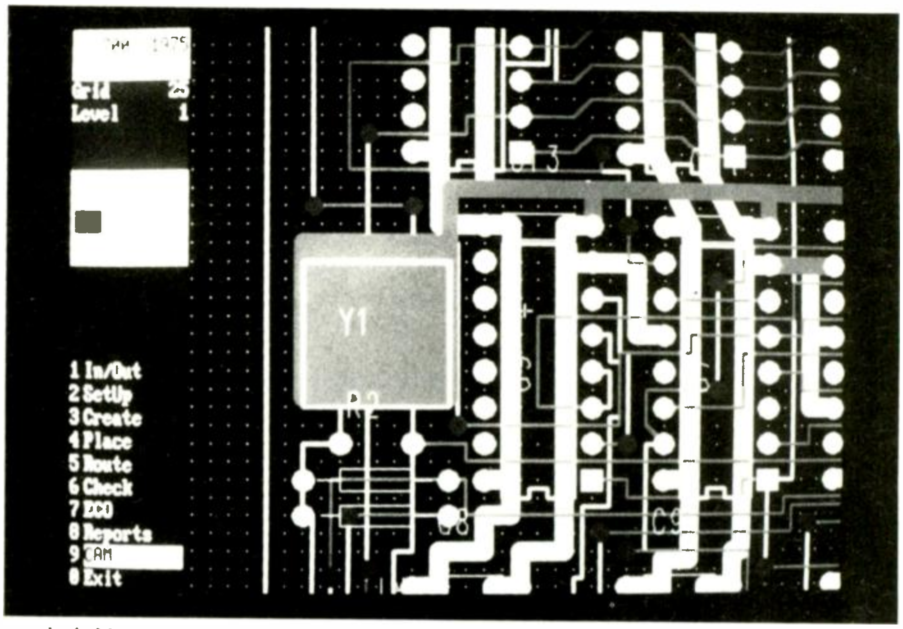
The final stage, and the most complex for the program (and the user) is the physical routeing of the tracks on the PCB. Again, there are two ways to accomplish this – interactive and automatic. It is unlikely that any cad program will ever succeed in automatically routeing 100% of connections on every board it encounters, so the interactive routeing option is always required.

In manual routeing, the display shows the PCB with its placed components and their logical connections. To assist identification, the component outlines are yellow, pads are green and the logical connections are purple. Once you select a logical connection with the mouse, it changes to red for a short distance at the starting pin, then to white until it reaches the destination pin. At all times, a yellow information display area shows data about the connection. For instance,

DATA6
U3.5
U7.9
Width 25

tells you that the signal path currently selected is DATA6, connecting pin 5 of IC_3 to pin 9 of IC_7 with a width of 0.025in.

Moving the mouse in combination with various function keys then lets you route the physical connection, changing layers with automatic via insertion if



needed. No matter where you are in the route, the uncompleted part is always shown in white, and you cannot complete the route to the wrong pin. As you might expect, you can unroute connections if needed.

With autorouteing, PADS-PCB does most of the routeing work for you. The evaluation package contains PADS-ROUTE, which use three separate routers. Two are specialised, for memory and power bus connections, while the third is a ten-pass general-purpose router for all other connections.

The power and ground bus router is heuristic, that is it works to a predetermined pattern, in much the same way that a human tries to lay out a similar bus manually. The memory router (for memory chips) works in the same way. The Maze router (general-purpose) handles the remaining connections on up to two layers at a time, needing multiple passes if there are more layers.

When you select Autoroute, there is a large number of options in a menu list that require setting before you go ahead. The manual is quite explicit as to those to select, but very vague as to why you select certain of them. Since the options are not always self-explanatory, I was left wondering at times what the program was actually doing.

Using all three of the routers on one of the supplied example files, the software routed 264 connections out of a total of 271, or around 97%, in about four minutes. The routeing is shown on screen as it takes place, a fascinating process, with the logical connections being replaced by routes as each is made. The program also gives a running data display on the state of the passes and success/failure at all times. You can also choose to route specific nets (connections with the same signal name), or specific connections if required. Any connections not made when the automatic routeing has finished need to be made manually.

PADS-PCB includes a CHECK option, used to make sure your design does not violate spacing rules (defined in the SETUP menu), and is accurate to 0.0005in. This includes tracks colliding with text as well as direct short circuits.

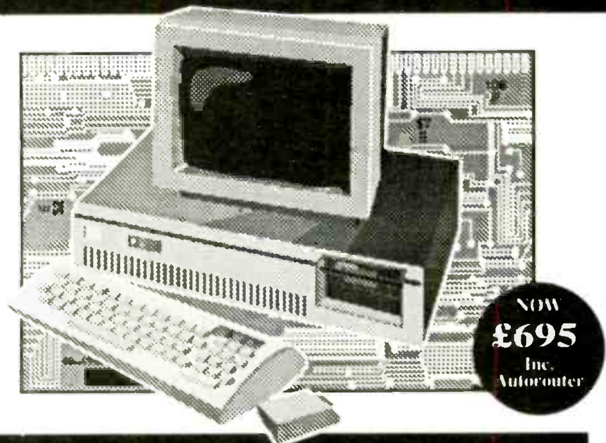
Having used the PADS-CAE/PCB programs on and off for several weeks, I still have mixed feelings about them. They appear to do an excellent job on the supplied files, and all indications are from the specifications that the package does the work it was designed for. But I would still have liked to have run a known circuit schematic through the whole process and seen what came out the other end. It would also have helped if more explanation of terms and actions had been given in the manual at certain points.

This sort of highly automated cad program certainly has a future on the PC now that a high degree of processing power is available for a reasonable price. No longer will a small company need to spend many tens of thousands of pounds for a dedicated workstation when designing PCBs – software like this and the not-so-humble PC will see to that.

Tony Bailey

PADS Superstation
Microtel Systems Ltd,
Glanty House,
The Glanty,
Egham,
Surrey TW20 9AH
0784 39881/35364

PCB TURBO V2



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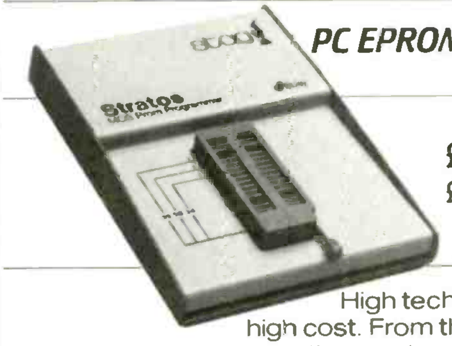

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NEW ARCHITECTURE OP-AMPS

High speed current-feedback op-amps were first described in the research journals about five years ago. Over the past year or so, a new style op-amp has emerged – with several analogue semiconductor manufacturers now producing standard parts. Though each has its own approach, the basic architectures are remarkably similar. Elantek was first with the EL2020 and EL2022, followed by Comlinear Corporation's CLC400 and CLC401. More recently the competition for specific sectors of the high speed op-amp market has been hotting up with devices such as the dual op-amp OP-260 from Precision Monolithic Inc. and the AD844 from Analog Devices, both released earlier this year.

Why now?

Way back in the early days of production op-amps such as Fairchild's 709, the structure was based on a standard BJT architecture with a long-tail pair (LTP) input stage driving resistive collector loads, feeding perhaps a second LTP for additional open-loop gain, followed by a DC level-shifting stage and output buffer/power amplifier. About the time of these first monolithic op-amps the analogue circuit designer's cry was first heard, *"My oscillator won't but my amplifier will!"*

Then came the breakthrough of the internally-compensated op-amp, such as the 741, with active loads replacing the resistive collector loads of earlier designs.

And then we all became familiar with the two fundamental limitations of this op-amp architecture. The first is that the gain-bandwidth product is constant, resulting in a direct trade-off between closed-loop gain and closed-loop bandwidth. Secondly, the slew-rate of these op-amps is not particularly high, often being limited by the same internal compensation capacitor that makes the op-amp such a stable and well behaved device. Slew-rates are typically 0.5 to 10 V/μs. We are so used to these limitations that we are almost inclined to believe that they are fundamental laws of physics and not a result of a particular op-amp architecture.

Incremental changes have been made in op-amps designs over the years to improve particular parameters, such as input-referred offset voltage, power supply rejection ratio and so forth, but these changes have been made without altering the basic architecture.

Introduction of fets within op-amps has resulted in lower power consumption, with reduction of input bias currents by up to three orders of magnitude in some cases, and a consequential increase of input impedance by the same order. However, fets brought no major changes of architecture.

In the latter half of the 1970s, about the only brave new architecture op-amp was National Semiconductor's LM3900 Norton op-amp. However, despite a radical design concept and its promise of versatility, it did not change the face of analogue design and it has not displaced the conventional architecture op-amp at all.

So why after such a long time are we seeing major changes in the basic architecture of the op-amp? Silicon n-p-n BJTs are faster by virtue of their higher electron mobility and so a high-performance BJT monolithic process has always been predominantly n-p-n with p-n-p realized as lateral rather than vertical devices. These p-n-p devices have a very poor β and high-frequency performance. This has cramped analogue designers' style into non-symmetrical circuits with the signal path handled mainly by n-p-n transistors. But recent technological advances enable high quality vertical p-n-ps to be fabricated and so the designer is free to use almost as many p-n-ps as n-p-ns in the signal path without degrading performance. With this new-found freedom, circuits with a radically new topology can be fabricated. The classical op-amp architecture can be dropped and new approaches taken up.

Current-feedback op-amps

Figure 1 shows the basic structure of the current-feedback of the op-amp, as presented by Derek Bowers of PMI at the IEEE's Bipolar Circuits and Technology Meeting held in Minneapolis last October. The now common current-mirror symbol has been used to simplify the diagram, the driven side of the mirror indicated by the arrow. Though each semiconductor manufacturer has individual refinements, all have adopted a very similar architecture.

Resistor R_2, connecting the inverting input to output, closes the feedback loop. The input voltage buffer Tr_1 to Tr_4 forces the inverting input node voltage to equal the non-inverting potential. Imbalance in the collector currents of Tr_3 and Tr_4 is summed at the voltage gain node z. Transistors Tr_5 to Tr_8 form an output voltage buffer to produce a low impedance at the output node. With an additional resistor R_1 between the inverting terminal and earth, the overall closed-loop voltage gain of the amplifier is $(1+R_2R_1)$, which is exactly

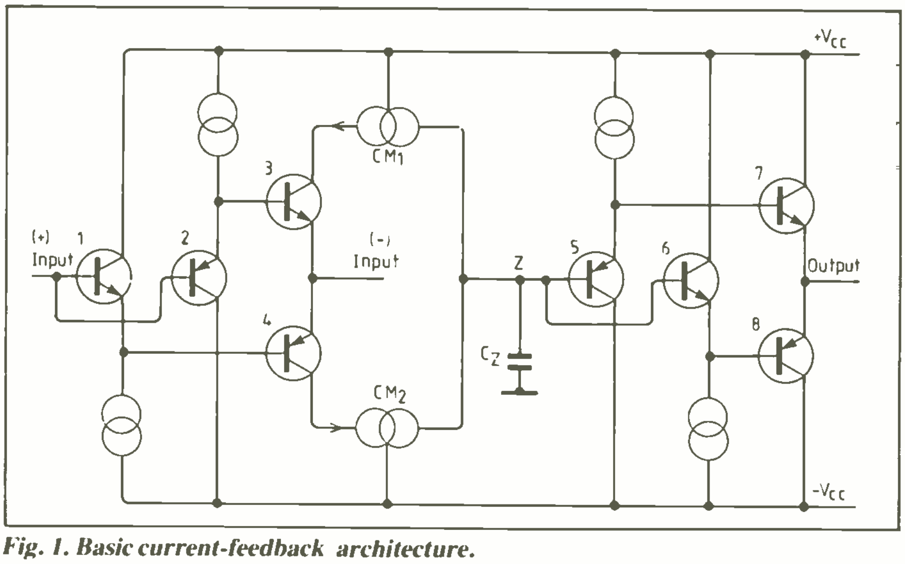


Fig. 1. Basic current-feedback architecture.

what would be obtained for a conventional op-amp.

The main difference in the new architecture is that the time-constant of the system is predominantly set by the product of resistor R_2 and the z node capacitance of C_z. The z-node is the only high impedance internal node in the amplifier. With this structure the gain-bandwidth trade-off of the conventional op-amp does not occur: instead, the bandwidth remains virtually constant.

Also the slew-rate of the amplifier is very high. In conventional op-amps the slew-rate is generally determined by the ratio of long-tail current to compensation capacitor, which is difficult to increase above a few tens of volts per microsecond. But the new architecture does not suffer from this limitation. Current available to charge the capacitor C_z is always proportional to the input error current, the slew-rate limitations associated with conventional op-amps just do not occur. In practice, rise and fall times are almost independent of signal level, and the input current will eventually cause the mirrors to saturate. The effective $(dV_{out}/dt)_{max}$ is as high as 2000V/ms in the case of the AD844, some 500-1000V/µs in PMI's dual op-amp OP-260, 700V/µs in Comlinear's CLC400 series and around 500V/µs in the Elantek EL2020 and 2022 devices.

Applications

Clearly, wide bandwidth and high slew-rate are achievable at low cost with these new op-amps, though the applications engineer is advised to read the manufacturers' notes very carefully. There are subtle points to consider when using these devices, such as gain-peaking effects in non-inverting applications due to stray input capacitance at the inverting input node.

A particularly elegant use of the current-feedback op-amp is in automatic gain control (AGC) applications. One of the shortcomings of voltage-mode op-amps in AGC amplifiers is that the bandwidth varies with closed-loop gain. The constant bandwidth feature of the current-feedback architecture overcomes this problem completely. **Figure 2** shows the circuit of an AGC amplifier using two halves of the dual OP-260 from PMI. Amplifier A_{1a} is used as the gain stage with A_{1b} set up as a positive precision rectifier. Op-amp A_2 is an integrator; this feeds an n-channel j-fet which is used as a voltage-controlled resistor to set the gain of A_{1a}. The 604kΩ resistor pulls a constant current from the inverting input of A_2. If the average amplified signal output falls below 15/604k = 25mA, then the gate drive to the j-fet will reduce R_{ds} of the j-fet, with the result that the gain setting of A_{1a} rises to compensate, hence the AGC action. **Figure 3** shows the pulse response of the amplifier for inputs of 100mV (at a) and 1V (at b, upper trace). The output responses are the lower traces, which show virtually identical rise and fall characteristics. In fact, the loop maintains a constant peak output amplitude for a square wave input signal range of ±20mVpk-pk to ±6.0Vpk-pk.

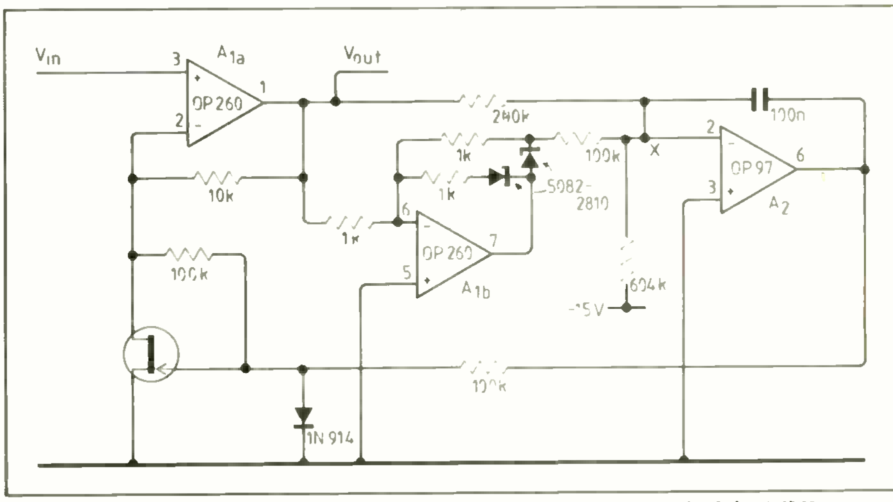


Fig. 2. PMI's OP-260 eliminates the problem of variable bandwidth in AGC amplifiers using voltage-feedback op-amps (Precision Monolithic Inc.).

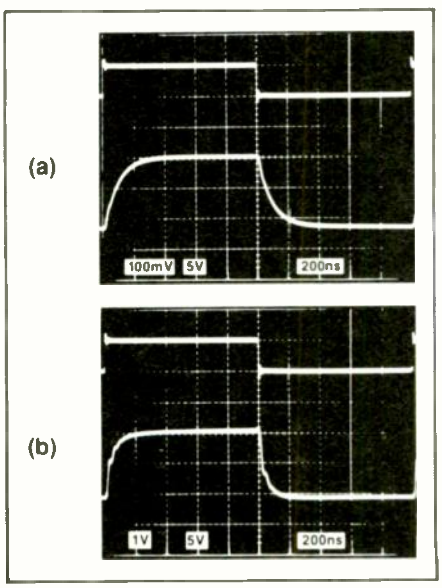


Fig. 3. Pulse response of the AGC amplifier of Fig. 2: (a) low-level input signal (b) large signal. (Precision Monolithic Inc.)

As yet the DC performance of the new architecture op-amp is not as good as the conventional op-amp. Figures for CMRR, PSRR are as yet well down. But, if it is wide bandwidth together with very good large-signal fast response you need, at low cost, then take a look at this new stable.

P-n-p pair with 3% matching

Drawing on the technology that has enabled devices like the current feedback op-amps to be developed, PMI is offering the MAT-03, a dual p-n-p matched pair with excellent parametric matching and high frequency performance.

Its low noise characteristics (1nV/√Hz maximum at 1kHz), broad bandwidth (190MHz typical at 10mA collector current), and low offset voltage (±100mV) make the MAT-03 an excellent choice for demanding low-noise and high-speed preamplifier applications. Tight current gain matching (3% maximum mismatch) and high current gain (100 minimum) over a wide range of collector current make the device ideal for current-mirror source applications. With a low value of bulk resistance (typically 0.3Ω), it is well suited to applications requiring very accurate logarithmic conformance.

Analogue Action is compiled by Dr John Lidgey of Oxford Polytechnic

PHASOR TRANSFORMS

Joules Watt explains how eliminating the time element in sinusoidal circuit analysis by means of phasor transforms yields an enormous simplification.

Yet again, the early development of what many of us take to be a relatively modern technique reminds us that there is nothing much new under the sun.

On this occasion, I am talking about the use of *phasor diagrams* in electronic circuit theory. The story began with the efforts made by Heaviside, Steinmetz and Kennelly to understand and characterize reactance and impedance[1,2,3]. In his 1887 paper, Heaviside talks about "admittance" and appropriately uses his operator "p" (for d/dt), interpreted for sinusoidal driving signals. Steinmetz and Kennelly offered phasor diagrams in their 1893/94 papers, although they called them vectors. For our present purposes, my earlier discussion about complex quantities sets the scene[4], but you could take a further look at the theory of them if you feel a bit rusty[5].

Of course, phasors appear on the page as *directed line segments*, similar to vectors but they represent an amplitude and phase plot of a time-varying sinusoidal quantity; see **Fig. 1**. Vectors, on the other hand, represent directed space-varying quantities. In some discussions you might come across *phasor-vectors*, where spatially distributed vector fields also oscillate in time with amplitude and phase variations.

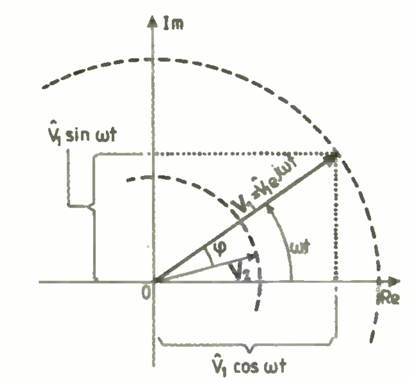


Fig. 1. A directed line, drawn from the origin on the complex plane, represents a phasor. This geometrical picture shows immediately the meaning of the angle and magnitude. The real and imaginary components also appear naturally as projected line segments along the axes. The smaller phasor V_2, is said to lag the larger, V_1, by ϕ radians.

Since phasors describe sinusoidally varying quantities, their relevance to radio-electronic engineering theory is obvious. Of all the "signals" we use, the sine wave wins hands down. Such signals form simple steady-state models of circuit action, including the easy introduction of something known as the impedance concept. At first sight you might consider sine waves rather boring from an information point of view and therefore not very relevant to most problems. In practice, they are relevant because, as we have seen, by Fourier analysis[6] all periodic waveforms contain a fundamental sinusoidal component, together with all the appropriate sinusoidal harmonics.

Transforms

But hereby hangs a tale. You can look upon phasors as a particularly simple example of a whole range of mathematical techniques based on *integral transforms*. The most obvious one that comes to mind involves Fourier again; Fourier transforms correlate non-periodic functions in, say, the time domain to a corresponding distribution in the frequency domain. You will find other Fourier pairs that associate: for example, space distributions transform to angular ones. Other transforms include Laplace, Hankel (*EWW*, May 1989, p.458), Mellin and even more exotic types.

Often, you will come across phasors discussed directly without reference to transforms. Nevertheless, they possess similar properties in that time functions, in the form of linear differential equations with constant coefficients, transform to algebraic equations in the frequency domain. The useful results arise because by employing a simple-harmonic forcing function you obtain a straightforward particular integral of the differential equation. In solving such differential equations, the natural response turns up, of course; this is the complementary function solution of the homogenous equation (that is, the differential equation equated to zero instead of the forcing function). You will find that the natural responses in all networks containing resistance die out as time passes. Responses in high-Q circuits might take a considerable time to die, but eventually even in these only the steady-state response remains.

Writing down the equation

An example always helps to clarify the situation. Consider the circuit in **Fig. 2.** I have chosen a two-loop LCR network, excited by a sinusoidal generator, v_s. One or two questions arise with this kind of circuit: for instance, you might want to know what value v_o or i_l takes. A standard procedure involves using Kirchhoff's voltage law round each

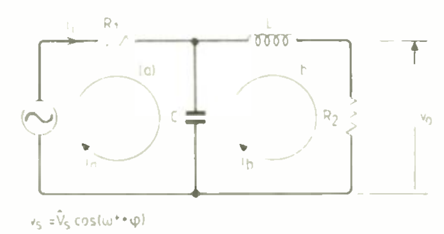


Fig. 2. The typical LCR network shown, has an input v_s. I have chosen two possible 'outputs', i_l and v_o. Setting up the loop currents i_a and i_b makes the circuit suitable for analysis by Kirchhoff's voltage law. You can see that $i_l = i_a$.

loop, or mesh, after choosing relevant loop currents. I have labelled these currents i_a and i_b in Fig. 2.

This mesh analysis yields simultaneous equations that you have to solve. Only two equations arise in this simple problem, but that is bad enough when you consider they will be differential equations. First, go round loop (a),

$$i_aR_1 + \frac{1}{C}\int i_a dt - \frac{1}{C}\int i_b dt = v_s.$$

Now loop (b),

$$-\frac{1}{C}\int i_a dt + R_2 i_b + L\frac{di}{dt} + \frac{1}{C}\int i_b dt = 0.$$

These are integro-differential equations. Your next move is to differentiate right through with respect to the time in each case, which eliminates the integrals,

$$\therefore \quad R_1\frac{di_a}{dt} + \frac{i_a}{C} + \frac{i_b}{C} + \frac{dv_s}{dt}$$

and

$$-\frac{i_a}{C} + R_2\frac{di_b}{dt} + L\frac{d^2 i_b}{dt^2} + \frac{i_b}{C} = 0.$$

Now you have to solve these simultaneously with the known forcing function v_s in place. A direct approach to this means struggling with di_a/dt and interpreting the integrating process.

At this point, some people try an "operator" method, (by writing d/dt = D and $\int ...dt$ = 1/D) and you will find some simplification accrues by using this traditional method[7].

Enter the transform

All LCR combinations form linear circuit networks whose natural responses to a transient input at some instant turn out to be exponential functions in every case. What you find looks like Ae^{st}, with A representing some kind of signal amplitude and s taking on various forms; real, imaginary or complex, according to the circuit. The interesting thing about exponential functions arises from the fact that they reproduce themselves when differentiated or integrated.

In the sinusoidal steady-state, or phasor approach, we choose an imaginary power in the exponential so that the driving function becomes the real part of $Ae^{j\omega t}$. In other words, we choose s = $j\omega$, which we assume to have begun indefinitely in the past and to go on indefinitely into the future. You might think this a bit artificial, but the implication is "you are in the middle of a long period – a sufficiently long time to disregard any transients". This means that no awkward initial or boundary conditions remain to bother us and explains the other title, "steady-state analysis", we give to this approach.

Complex exponential functions possess another interesting property which I have often used in my discussions. This arises from the mention of Euler's identity,

$$e^{j\omega t} = \cos\omega t + j\sin\omega t.$$

By taking the real part of this you see immediately that

$$\mathrm{Re}\,(e^{j\omega t}) = \cos\omega t.$$

***Fig.3.** In (a), the 'unit phasor' has a length equal to 1 and rotates anticlockwise at an angular rate ω. The amplitude of the phasor in (b) corresponds to a voltage with peak value V_s. The presence of a phase angle φ advances the phasor relative to direction ωt. By removing the dependence on ωt, in other words, by taking the reference direction along the real axis, a fixed phasor V_s represents the amplitude and phase angle, as shown in (c).*

Write the complex conjugate of $e^{j\omega t}$ and its Euler expansion,

$$e^{-j\omega t} = \cos\omega t - j\sin\omega t.$$

Thus, by adding the two, you get,

$$\cos\omega t = \frac{e^{j\omega t} + e^{-j\omega t}}{2}.$$

Subtracting them would give you the imaginary part, namely $j\sin\omega t$. This shows that you can write sinusoidal forcing functions as complex exponentials. After working the problem with these exponentials, you then take the real, or "cos" bit, or the imaginary, or "sin" part as the solution.

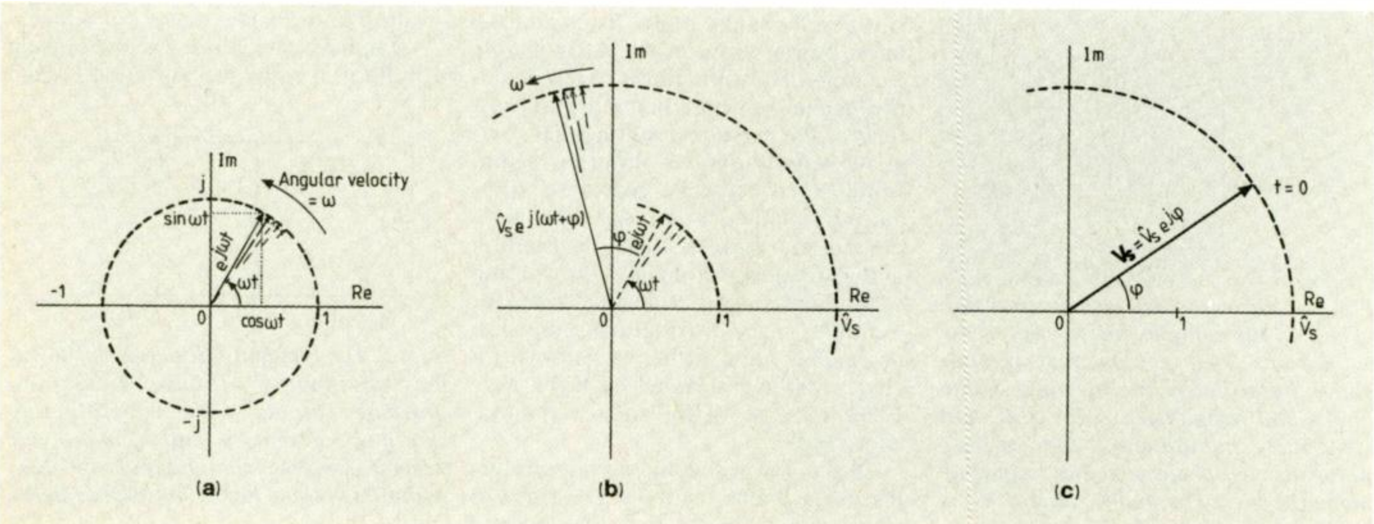
The expression $e^{j\omega t}$ has a magnitude equal to unity all the time. You can see this by taking the magnitude of its Euler expansion,

$$|e^{j\omega t}| = |\cos\omega t + j\sin\omega t|$$
$$= \sqrt{\sin^2\omega t + \cos^2\omega t} = \sqrt{1} = 1$$

As the time ticks away in the exponent, this unit length, looked on as a phasor, spins anticlockwise with angular velocity ω; **Figure 3 (a)** illustrates this. If there is an amplitude $\hat{V}s$ and a phase angle φ in the forcing function, that is, $v_s = \hat{V}_s\cos(\omega t + \phi)$, then this causes no trouble as Euler's identity now appears as

$$\hat{V}_s e^{j(\omega t+\phi)} = V_s e^{j\phi} e^{j\omega t}$$
$$= \hat{V}_s(\cos(\omega t + \phi) + j\sin(\omega t + \phi))$$

This shows that the anticlockwise rotating unit phasor is multiplied in length by $\hat{V}_s$ and has φ radians added to its angle, as in **Fig. 3(b)**. Finally, by convention, we stop the rotation of the phasor by removing the factor $e^{j\omega t}$ from further consideration until the end of the problem. We do this by taking a "snapshot" at t = 0 so that $e^{j\omega t}$ = 1. You can mop up the $e^{j\phi}$ into the amplitude $\hat{V}_s$ to form a



complex amplitude $\mathbf{V}_s$. This we usually call the *phasor;* in other words, we apply the term after removing the time rotation part, as in **Fig. 3 (c).**

$$\therefore \qquad \mathbf{V}_s = \hat{V}_s e^{j\phi}$$

Such functions of the independent variable at the input mean that the dependent variables, or outputs, follow the same form. Therefore, if you take v_o as the output, then you can write $v_o = \mathrm{Re}(\mathbf{V}_o e^{j\omega t})$. Also all the other values of voltage and current will have a similar form, so that for example i_l will equal $\mathrm{Re}\,(\mathbf{I}_l e^{j\omega t})$. These are the phasor transforms. Notice we have made the complex amplitudes $\mathbf{V}_o$, $\mathbf{I}_l$ independent of time, but they are functions of frequency.

If you return to the differential equation I used in the example earlier, you will see how the phasor transform yields a huge simplification. Before proceeding, remember that only the real part of the phasor expression has physical relevance. But if, further, you remember the properties of the complex conjugate, then obtaining the real part turns out to be simple. Any complex number z has a real part given by

$$\mathrm{Re}(z) = \frac{z + z^*}{2}$$

where the asterisk denotes the conjugate; see **Fig. 4.**

$$\therefore \quad \mathrm{Re}(\mathbf{V}_o e^{j\omega t}) = \frac{\mathbf{V}_o e^{j\omega t} + \mathbf{V}_o{}^* e^{-j\omega t}}{2}$$

a simpler version of which we have already used to obtain the cosine.

Now insert the phasor expressions into the first differential equation,

$$R_1 \frac{d}{dt}\left(\frac{\mathbf{I}_a e^{j\omega t} + \mathbf{I}_a{}^* e^{-j\omega t}}{2}\right) + \frac{1}{C}\left(\frac{\mathbf{I}_a e^{j\omega t} + \mathbf{I}_a{}^* e^{-j\omega t}}{2}\right) - \frac{1}{C}\left(\frac{\mathbf{I}_b e^{j\omega t} + \mathbf{I}_b{}^* e^{-j\omega t}}{2}\right) = \frac{d}{dt}\left(\frac{\hat{V}_s}{2}\left[e^{j\omega t + \phi} + e^{-j\omega t + \phi}\right]\right)$$

Then carry out the differentiations because, as mentioned, the $\mathbf{I}(\omega)$, $\mathbf{V}(\omega)$ have no dependence on the time, so

$$\frac{d}{dt}(\mathbf{I}_a e^{j\omega t}) = j\omega \mathbf{I}_a e^{j\omega t},$$

$$\text{in other words, } \frac{d}{dt} = j\omega.$$

After collecting terms and cancelling the 2, you should end up with

$$\left(j\omega R_1 \mathbf{I}_a + \frac{\mathbf{I}_a}{C} - \frac{\mathbf{I}_b}{C}\right) e^{j\omega t} + \left(-j\omega R_1 \mathbf{I}_a{}^* + \frac{\mathbf{I}_a{}^*}{C} - \frac{\mathbf{I}_b}{C}\right) e^{-j\omega t} = j\omega \hat{V}_s e^{j\phi} e^{j\omega t} - j\omega \hat{V}_s e^{-j\phi} e^{-j\omega t}.$$

And similarly with the second equation,

$$\left(-\frac{\mathbf{I}_a}{C} + j\omega R_2 \mathbf{I}_b - \omega^2 L \mathbf{I}_b + \frac{\mathbf{I}_b}{C}\right) e^{j\omega t} + \left(-\frac{\mathbf{I}_a{}^*}{C} - j\omega R_2 \mathbf{I}_b{}^* + \omega^2 L \mathbf{I}_b{}^* + \mathbf{I}_b{}^* + \frac{\mathbf{I}_b}{C}\right) e^{-j\omega t} = 0.$$

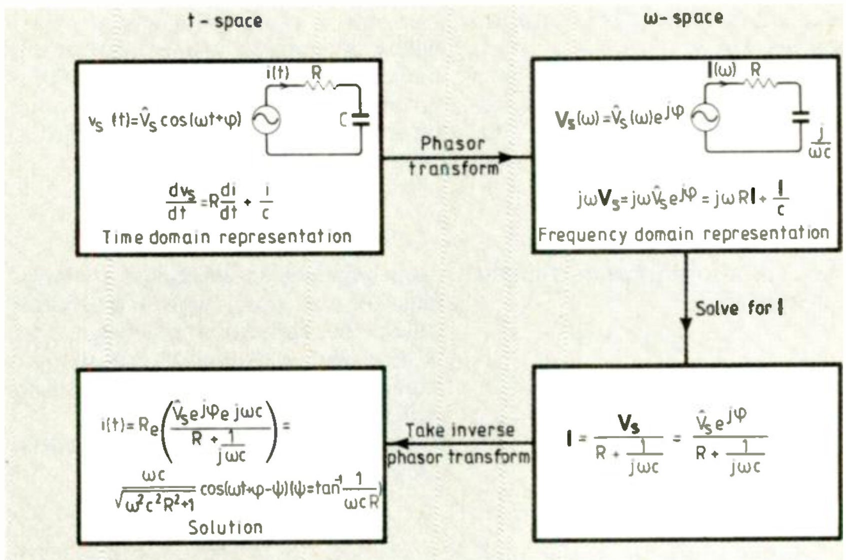


Fig.5. Using a transform method follows a procedure. The direction goes from the time domain into the frequency domain. Solving the algebraic equations there is relatively simple. Finally, an inverse transform yields the time solutions.

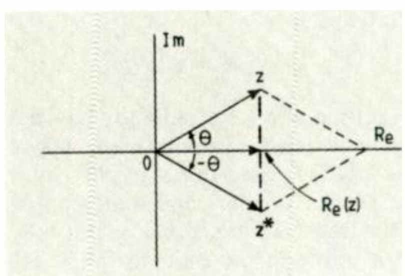


Fig.4. The complex number z has a conjugate z* represented geometrically as a reflection in the real axis. If you take the sum of z and z* you will always get a real number result.

We have to struggle on a little further, but with the pleasure of knowing that in all future problems, you will be able to write down the final phasor equation straightaway. You only have to go through the above tortuous route once to see the method.

Notice that the only way the equations can equal zero for all times requires each coefficient of the exponential factors separately to equal zero. The other way you can state this is to say that $e^{j\omega t}$ and $e^{-j\omega t}$ are *linearly independent,* and again because of this the coefficients must be zero. Mathematically, the complex conjugate terms contain all the information which the other terms contain, so we only need one set of coefficients. This means that, after dividing through by $j\omega$,

$$R_1 + \frac{1}{j\omega C}\mathbf{I}_a - \frac{1}{j\omega C}\mathbf{I}_b = \hat{V}_s e^{j\phi} = \mathbf{V}_s$$

and,

$$-\frac{1}{j\omega C}\mathbf{I}_a + R_2 + j\omega L + \frac{1}{j\omega C}\mathbf{I}_b = 0$$

...which you recognise as the final result.

Note the $i_a(t)$, $i_b(t)$ transform into $\mathbf{I}_a(\omega)$, $\mathbf{I}_b(\omega)$ and so on. Also $d/dt = j\omega$. This means that, by looking at the equations, you could have written down the results immediately and thus obtained the algebraic phasor equations from the differential time equations in one step.

You can now draw up a short table of phasor tranforms.

$$\varphi[v] = \mathbf{V}$$

$$\varphi[av] = a\mathbf{V}$$

$$\varphi\left[\frac{d^n v}{dt^n}\right] = (j\omega)^n \mathbf{V}$$

$$\varphi[\cos(\omega t + \phi] = e^{j\phi}$$

$$\varphi[v_1 + v_2] = \varphi[v_1] + \varphi[v_2] = \mathbf{V}_1\ \mathbf{V}_2$$

This last relation works because of the linearity of the equations. The whole procedure cycles in a loop, which you will find characteristic of transform methods. See **Fig. 5.**

The solution always appears as a sinusoid, because we applied a sinusoidal excitation, or forcing function. This always means you can write solutions in the form,

$$v_o = \hat{V}_o\cos(\omega t + \phi)$$

The phasor transform appears from the earlier definition as,

$$v_o = \mathrm{Re}(\mathbf{V}_o e^{j\omega t})$$

$$\therefore \hat{V}_o\cos(\omega t + \phi) = \mathrm{Re}(\mathbf{V}_o e^{j\omega t})$$

so that by using Euler's identity and the conjugate,

$$\frac{\hat{V}_o e^{j\phi}e^{j\omega t}}{2} + \frac{\hat{V}_o e^{-j\phi}e^{-j\omega t}}{2} = \frac{\mathbf{V}_o e^{j\omega t}}{2} + \frac{\mathbf{V}_o^* e^{-j\omega t}}{2}$$

which, together with the linear independence of $e^{j\omega t}$ and $e^{-j\omega t}$, means we end up with

$$\hat{V}_o e^{j\phi} = \mathbf{V}_o$$

as I have already pointed out.

This yields the *transform pair*

$$v_o = \mathrm{Re}(\mathbf{V}_o e^{j\omega t})$$
$$\mathbf{V}_o = \hat{V}_o e^{j\phi}$$

These tell you how to get v_o from a known $\mathbf{V}_o$, and how to get $\mathbf{V}_o$ from a knowledge of v_o. Finally, you can consider $\mathbf{V}_o = \varphi[v_o]$ as the phasor transform of v_o and $v_o = \varphi^{-1}[\mathbf{V}_o]$ as the inverse phasor transform of $\mathbf{V}_o$.

Impedance

The great value of this transform method lies in the ability to bypass the time equations completely. The linearity of the situation means you can write down mesh and nodal equations directly in the phasor frequency domain form.

Because $v_l = L\frac{di_L}{dt}$, $i_c = C\frac{dv_c}{dt}$ and $v_r = Ri_r$ transform directly into,

$$\mathbf{V}_L = j\omega L\mathbf{I}_L,\ \mathbf{I}_c = j\omega C\mathbf{V}_c \text{ and } \mathbf{V}_r = R\mathbf{I}_r$$

you can take the quotient, $\mathbf{V}$ divided by $\mathbf{I}$ and thereby obtain the familiar impedance relationship $\mathbf{Z} = \mathbf{V}/\mathbf{I}$. Notice that the impedance, as the quotient of two complex numbers, itself consists of a complex number. The quotient of a voltage with a current has the dimension of ohms, of course.

Impedance applies while in the frequency domain. If you try to talk about it in the time domain, you will be discussing something that has no meaning. Straightaway, you see that the familiar impedance relationship for an inductor,

$$\mathbf{Z}_L = \frac{\mathbf{V}_L}{\mathbf{I}_L} = j\omega L,$$

has only a positive imaginary part – whose magnitude usually carries the name inductive reactance, X_L. For a given inductor, the reactance varies directly with the frequency. Similarly, a capacitor has an impedance

$$\mathbf{Z}_C = \frac{\mathbf{V}_C}{\mathbf{I}_C} = \frac{1}{j\omega C} = -\frac{j}{\omega C}$$

again yielding an imaginary quantity, negative this time, with a magnitude called the capacitive reactance, X_C. Notice the well known result that capacitive reactance varies inversely with the frequency.

The impedance of a resistor equals its ordinary resistance

$$\mathbf{Z}_R = \frac{\mathbf{V}_R}{\mathbf{I}_R} = R$$

and, being a real quantity, it shows a zero phase angle must exist between phasors $\mathbf{V}_R$ and $\mathbf{I}_R$, together with no variation in R with frequency.

In a general way, the impedance and phasors enable you to write down the phase angles in the inductive and capacitive cases.

$$\therefore \quad \frac{\mathbf{V}_L}{\mathbf{I}_L} = j\omega L = \omega L\angle 90°$$

This tells us that an inductor has the positive reactance I mentioned, whose ohms lead resistive ohms by 90°, as it were. It also tells us – and some people get confused at this point – that the current through the inductor must *lag* the voltage across it by 90°. You can see this by noting that $\mathbf{I}_L$ in the denominator of $\mathbf{Z}_L$ requires −90° relationship to $\mathbf{V}_L$ in the numerator, to yield a resultant +90° for the quotient. In other words,

$$\frac{\mathbf{V}_L}{\mathbf{I}_L} = \frac{\hat{V}_l\angle 0°}{\hat{I}_l\angle -90°} = \frac{\hat{V}_l\angle 0° + 90°}{\hat{I}_l} = X_L\angle +90°$$

In a similar way, but with everything reversed (the "dual"), you can determine the phasor results for capacitors.

If you take the reciprocals of all the impedance relationships, you end up with admittances. You have probably come across the terms conductance and susceptance for the real and imaginary parts of these.

Network immittances and functions

People talk about impedances and admittances in the same breath. The term *immittance* soon became established in engineering circles to mean the quotient of phasors $\mathbf{V}$ and $\mathbf{I}$ taken either way up. (Although W. H. Chen[8] offers the alternative *adpedance* in place of immittance.)

If you consider one of the quotients $\mathbf{V}_i$ with $\mathbf{I}_i$ at the input terminals of a network or system, then you obtain the driving point immittance. If taken at the output, then you end up with the output impedance (or admittance). A touch more subtle perhaps: if you form these quotients by taking a phasor voltage or current or voltage from another part, then you get a transfer immittance.

In this latter case, you might have noticed the possibility of dividing one phasor voltage by another, (or a current by a current). You obtain a ratio now, which cannot be an immittance of course, but such ratios exist and possess important meanings. The voltage ratio gives the voltage gain (or loss) between the two sections. The ratio of currents yields the current gain (or loss). Engineers call these ratios the transfer functions of the system. Clearly, since immittances must be functions of ω (or jω), so must transfer functions.

These concepts nicely overlap with my earlier discussion on transfer resistances and so on, as applied to amplifiers[9]. Generalising to include impedances would broaden that discussion to include complex numbers, with their phase and amplitude responses as function of frequency. You would find the present material relevant.

This brings me back to our unsolved pair of equations, which still need interpreting in the light of what I have just said. The phasor transform greatly simplified the approach, but what do we do with it next? First, solve the equations for either $\mathbf{I}_a$ or $\mathbf{I}_b$, employing any ordinary method. I will use Cramer's rule with determinants[10]. For $\mathbf{I}_a$,

$$\mathbf{I}_a = \frac{\begin{vmatrix} \mathbf{V}_s & -\dfrac{1}{j\omega C} \\ 0 & R_2 + j\omega L + \dfrac{1}{j\omega C} \end{vmatrix}}{\begin{vmatrix} R_1 + \dfrac{1}{j\omega C} & -\dfrac{1}{j\omega C} \\ -\dfrac{1}{j\omega C} & R_2 + j\omega L + \dfrac{1}{j\omega C} \end{vmatrix}}$$

$$\therefore \mathbf{I}_a = \frac{V_s(R_2 + j\omega L + \frac{1}{j\omega C})}{R_1(R_2 + j\omega L + \frac{1}{j\omega C}) + \frac{1}{j\omega C}(R_2 + j\omega L)}$$

Immediately, you get the driving point admittance $\mathbf{I}_a/\mathbf{V}_s$ for our example network.

On the other hand, by finding $\mathbf{I}_b$ in a similar way and by noting that the product $\mathbf{I}_b R_2$ gives the output voltage $\mathbf{V}_o$, you obtain the voltage transfer function.

$$H(j\omega) = \frac{\mathbf{V}_o}{\mathbf{V}_s} = \frac{R_2}{j\omega CR_1(R_2 + j\omega L + \frac{1}{j\omega C}) + (R_2 + j\omega L)}$$

You may enjoy obtaining this result as an exercise.

H(jω) gives you voltage "gain" information as you process the sinusoidal input signal through your network. Plotting |H(jω)| as a function of frequency yields the amplitude response. Similarly plotting H(jω) gives the phase response of the network.

I hope you see how economical in time and effort the use of phasor transforms have made calculations of impedances and responses of LCR linear networks, when they are driven by sinusoidal test signals. But, you may ask, what about the transient response we have neglected? Also, does the use of linear amplifiers – in other words, controlled sources – in the network modify the results? They do and, as for transient response, the departure from simple sinusoidal driving signals to abrupt switching types certainly complicates the issue. But similar transform methods enable these to handled algebraically too. I add the comment that we might find time for a discussion of these and other points on another occasion. ■

References

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UHF direct frequency synthesizer

RF Commentary (*E&WW* April 1989, p.423) outlined recent work by the American company Digital RF Solutions on direct digital synthesis (DDS) for use in HF/VHF equipment. Work on DDS in the UK by Dr Peter Saul and others at Plessey Research, Caswell, has included the development of the SP2001 DDS 40-pin IC package first announced in 1987. This directly generates the code required for an output sinewave anywhere within the range 54kHz to 100MHz with a clock frequency to over 350MHz.

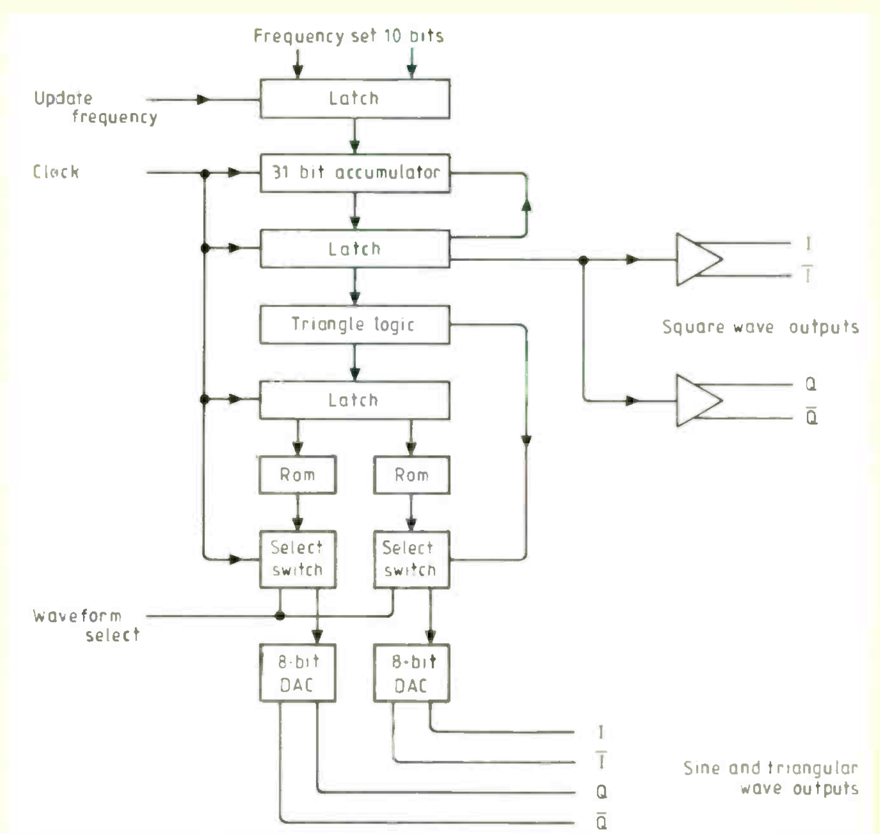
More recently, Peter Saul and David Taylor have described a new 2GHz-clock direct frequency synthesizer in which the two integral d-to-a converters each have a faster operating specification than any currently available, with devices tested to above 2.5GHz clock rate. The device, intended primarily for radar and electronic warfare applications, generates square, triangle and sinewave outputs over the range 1Hz to 500MHz with an accumulator length of 31 bits and a nominal clock frequency of 2147 483 648Hz with 1Hz frequency increments. In a paper presented at the recent international conference on frequency control and synthesis, the authors claim this as a new class of UHF frequency synthesizer of unprecedented performance:

> "The simplicity of operation, coupled with the complete absence of any setting-up or alignment procedures, makes this class of device very attractive. In some applications, the extreme speed of operation can make possible products which cannot be realised in any other way. Particular examples of this are in fast-hopping signal generation for EW and radar. Also, the device can provide almost all of the functions for a multimode signal generator operating over the 1Hz to 500MHz range."

Close-to-carrier noise is predicted as –135dBc/√Hz at 10kHz, with maximum spur levels of –65dBc at 30MHz, –58dBc at 50MHz and –30dBc at the 500MHz output.

First samples of the DFS have recently been evaluated with a 1.5GHz clock and are fully functional with the exception of the sinewave outputs. The project has been partially funded under the Alvey programme.

Block diagram of the new Plessey Research 0-500MHz direct digital frequency generator with clock rate up to 2.5GHz.



Wavell wins through (just)

For years, most progress in telecommunications has been hailed as stemming from the marriage of communications and computer technology. But marriages are not always successful or smooth-running and this seems to have been the experience of the British Army. It has previously been reported ("Ptarmigan at last", *E&WW* July 1985, page 75 and "Packets on the move" July 1987, page 755) that the development and introduction of the British Army's first battlefield C^3I (Command, Control, Communications and Information) system has taken well over 20 years. Even now the complete system – Ptarmigan distributed battlefield trunk network; Treffid, the radio-relay sector; and the Wavell automatic data processing system – is recognized, although now at last in service, as requiring further improvement in hardware, software and managerial structures.

Wavell was initially deployed in 1985 but it was not until the 1988 exercise *Summer Sales* that the Army began to have any real confidence in the system. As Major C.L.G. Wright (*J. Royal Signals Institution*, Winter 1988) has admitted: "Unfortunately Wavell has had a chequered history ranging from unreliable hardware to untrustworthy software. These problems were severe enough to erode staff confidence in it to the extent that the whole project was in danger of being cancelled.

During 1985-86, the system continued to be plagued with problems: lack of spares to repair the data processors (Plessey MDP1000); software "bugs" resulting in totalling mechanisms corrupting the database; and unreliable interprocessor commmunications, floppy disc drives and bubble memory units. System availability during exercises often proved to be less than 50%. The considerable problems of managing the system with Army rather than civilian specialists are now seen to have been underestimated.

Extensive modifications and improvements have been undertaken by Plessey, the prime contractor, mostly during 1987, including "massive surgery" to the hardware.

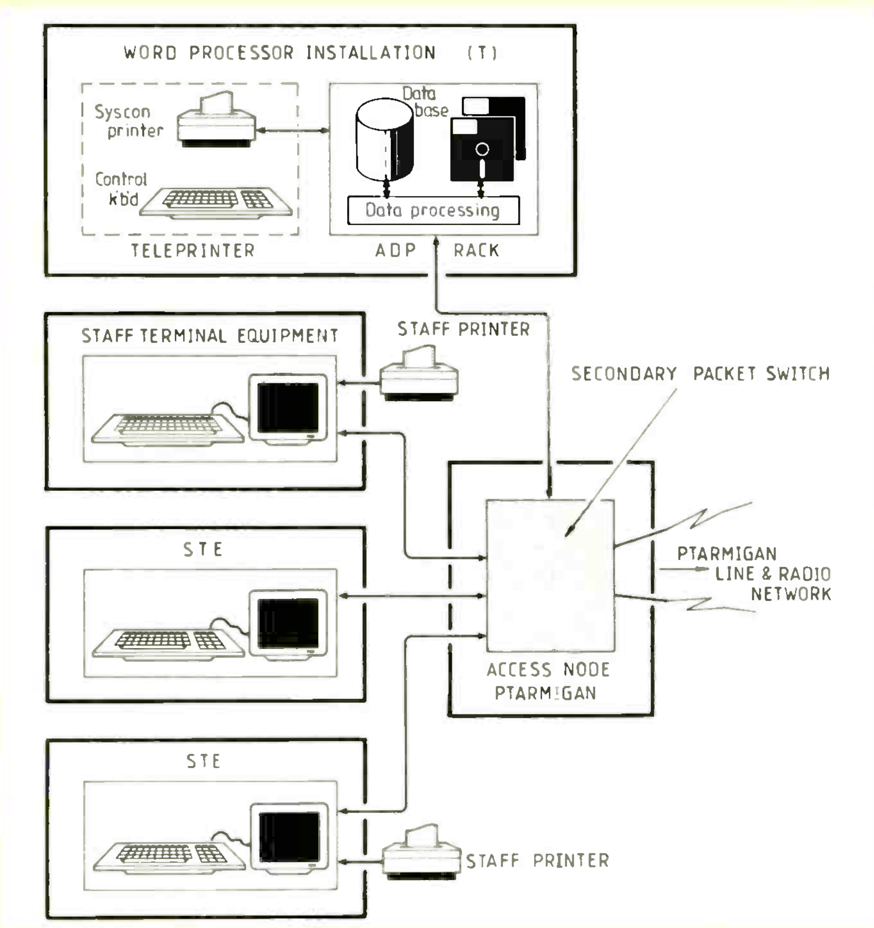
Wavell is intended to provide commanders and staff officers with a battlefield computer system to automate state boards, collate and process information and intelligence, with the ability to pass messages and orders as data over the packet-switched network (PSN) of the Ptarmigan system.

Major Wright stresses that: "The fundamental lesson that was learned with Wavell and indeed with Ptarmigan is that complex, dynamic ADP systems are bound to fail (not necessarily completely) on first fielding. The very nature of these systems would indicate that they are unlikely to work first time and this failure does not just occur initially, but every time the system is increased in size. These size-related system faults are unlikely to be found in the factory testbed no matter how rigorous the quality-assurance procedures... A sensible precaution should be to expect failure and have an agency available which is capable of identifying exactly what went wrong and then be capable of providing a palliative measure to circumvent the error. This will require a system support organization which will have as an integral part of its framework, hardware and software specialists... The Ptarmigan-Wavell System Support Team now provides this essential support for Wavell, but initially it also required further contractor support when Wavell was deployed on exercises... The final lesson is that the C^3I system will not necessarily save manpower... it needs an operational manager responsible for the currency of the critical information, a system manager responsible for the availability of the network, and probably a communications manager responsible for the interface with the various bearer systems."

In July 1985, I reported the belief of some senior officers of a developing "crisis in software" in the British Army. This foresaw the possibility of entire area networks such as Ptarmigan/Wavell failing at critical moments due to lack of skilled software support.

The chequered history of Wavell tends to support this view while at the same time highlighting also the hardware problems. Could it be that, 45 years on, the lessons of the disasters of Arnhem ("*Market Garden*") have been forgotten? A simple, limited capacity link that works is better than a sophisticated automatic data processing system that fails at a critical time. And what are the prospects for such complex systems as the American SDI ("Star Wars") for which no realistic full-scale trials will ever be possible?

Wavell automatic data processing system forming part of the British Army's C^3I battlefield system.



EMC directive "deeply disappointing"

Although the DTI continued to express reservations and, together with Ireland, abstained from voting when on May 3, the EMC directive of the European Community was finally submitted to the Internal Market Council (IMC), the text was agreed unanimously with no contrary votes. The final version of the text of this important directive has been modified in various, mostly minor, ways since the original proposal of November 11, 1987. By the time these notes appear, it may have been published in the official journal of the European Communities, and will take effect from January 1, 1992 (or January 1, 1993 in areas where there may still be no appropriate national standards).

Some of the recent modifications are regarded as "helpful" by the DTI but it still considers that: "In view of the extensive and, at times, promising lobbying campaign mounted by the UK over the last six months in particular, the eventual outcome of the Directive is deeply disappointing".

One of the main worries of the DTI appears to be in respect of the possible impact of this all-embracing EMC directive on telecommunications terminals, including subscribers' apparatus. The DTI believes there is no validity for treating most telecommunications equipment with a more onerous specification than information technology apparatus. This may perhaps be why British Telecom reacted strongly when last year I revealed that while considerable improvements had been effected over the past three or four years, many of the recent "electronic" telephones and associated units continued to suffer from relatively poor immunity to local RF fields, and BT endeavoured to conceal such information from would-be users.

A very important aspect of the directive about which there has been some confusion is that it has now been confirmed that the Directive *does* apply to specific items of an existing product design which are sold after January 1, 1992 (or 1993 in the absence of national standards), as well as to any brand-new apparatus marketed for the first time after the directive comes into force.

This summer, the DTI will issue a consultative document outlining the ways in which the EMC directive will be implemented in the UK and DTI's thinking on many of the important issues that have been identified. Views and comments will be welcome, although it is now too late to change the text of the directive. In addition, a booklet in the DTI's "Standards and the Single Market" series is being prepared. Other publications in this series include "Standard: Action Plan for Business" and "The New Approach to Technical Harmonisation and Standards" providing introductions to this area of EC policy. ■

RF Connections is compiled by Pat Hawker.

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