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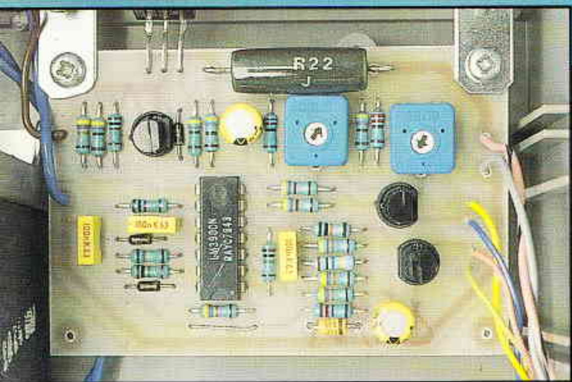

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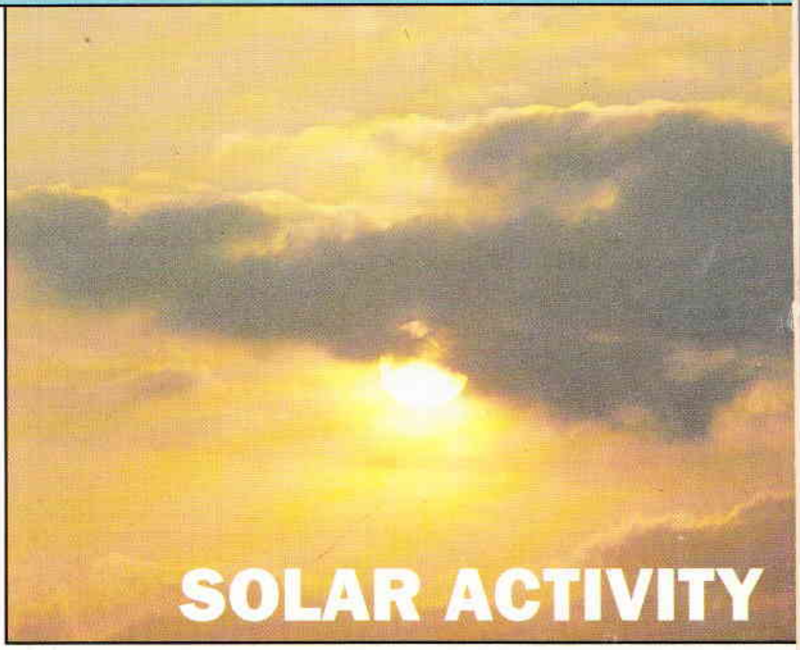
DYNAMIC NOISE LIMITER
DIY AERIAL SYSTEM
DIGITAL TV



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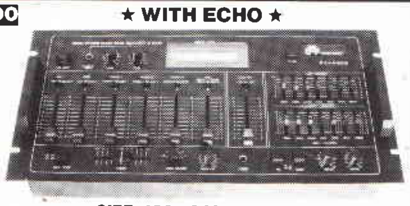

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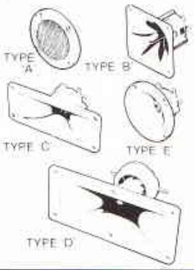

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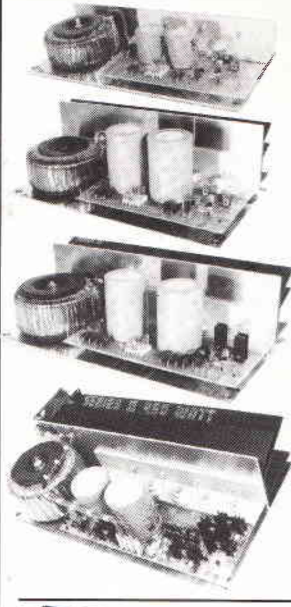

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OMP/MF 450 Mos-Fet Output power 450 watts R.M.S. into 4 ohms, frequency response 1Hz - 100KHz -3dB, Damping Factor >300, Slew Rate 75V/uS, T.H.D. typical 0.001%, Input Sensitivity 500mV, S.N.R. -110 dB. Fan Cooled, D.C. Loudspeaker Protection, 2 Second Anti-Thump Delay. Size 385 x 210 x 105mm.
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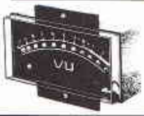

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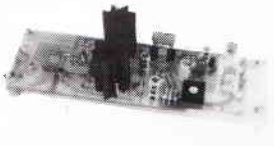

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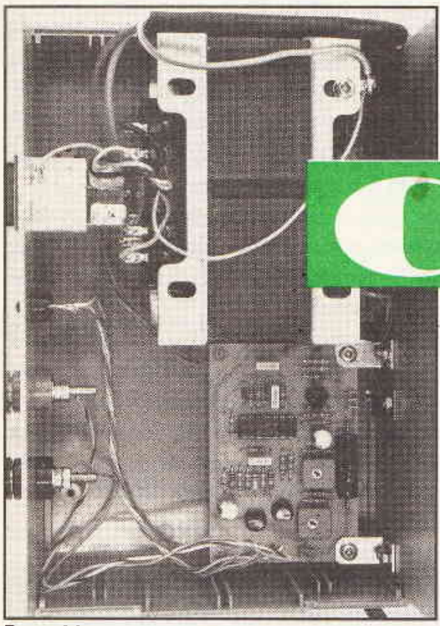
Volume 21 No 8
August 1992

Features & Projects



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Editorial

by Paul Freeman

In our world of digital electronics, the developmental driving force almost seems totally geared to packing in more power to give more facilities at a cheaper price. The last of these I think most of us would encourage, but the drive for miniaturisation and greater complexity has become inextricably linked to consumerism. The consumer is being sold a wonder gadget that not only has the built-in redundancy element but also is so complex that probably most of the functions it offers will never be used. I speak of computers, calculators, videos, synthesisers, and now TV sets. Only the very few will exploit their full potential. It seems one has to buy the complexity to keep the product viable.

Another significant factor not thought about by the manufacturer is associated with the physical size of data storage. Take the CD, an enormous store of information, but good old-fashioned readable sized printing placed on the disc has to tell us something about the contents. Smaller discs, although valuable for what they contain may well be limited as a consumer durable simply because sufficient information cannot be printed directly on the disc.

In the desire to steam speedily ahead with miniaturisation, one major world player in mass data storage had not given this simple visual indication of contents any thought. It could be that only space for a printed code number is available. Then the user would be faced with consulting an index for contents.

OPEN CHANNEL

Modern electronic telecommunications is wonderful, ain't it? Despite many years to get it right, thousands upon thousands of standards, recommendations and regulations, communications between electronics gadgetry such as computers, workstations and PCs is still largely a black art.

Let me explain. Until a year ago, my copy for this column and anything else I wrote for ETI was produced on my computer, at my house. I would print out the copy and send it to Paul, the Editor, by good old-fashioned post. Then, he would mark-up the copy to suit, send it onto ETI's typesetters, where it would be re-typed into high-quality image-set film, ready for printing.

Now, anyone could be forgiven for thinking that ETI as a flagship electronics magazine should have been produced by totally electronic methods. It should have been all written by writers like myself on computer, downloaded by telephone line to ETI editorial offices where the magazine was produced on computer screen. Finally the resultant pages should have been simply downloaded by telephone line, again, to the printer. In an ideal world this should have been so. But magazine publishing has had its financial ups and downs over the last seven or eight years, and ETI was not yet at the electronic stage.

About a year ago, however, ETI took its first tentative steps in this direction. ETI's computer system (ie, Paul's personal computer) could finally accept electronic copy from writers. Great, except that we couldn't get it to work. My computer still couldn't talk to his computer. My computer disks are still different to his. So much for all these standards!

So we hit upon an interim arrangement, which lasts to this day and will probably last a while longer, too. I write my copy and fax it to Paul. This is not a fax in the traditional sense, though. Plugged into the back of my computer I have a special modem which doubles as a direct fax connection. I can send a fax directly from my computer screen. So what I see on the screen I know will be printed out from any fax machine I want to send the fax to. So, I send my computer-generated Open Channel copy to Paul, via ETI's fax machine.

When Paul gets it seconds later, he scans it into his computer's scanner. Once he's done this he uses optical character recognition software to read in the scanned copy. Quality of my faxed material is so good (simply because it's coming directly from my computer - and not being faxed in the traditional sense) Paul tells me he can expect over 98% of words to be recognised by the optical character reader. So a typical length column like Open Channel which has around 1000 words, he has to correct something less than 20 words. From then on in, ETI' s computer takes over.

We know there are easier and better ways to do all this. But when we first tried to do-it-all electronically we wasted so much time and still couldn't get it to work that we simply ran out of time (ETI' s press days can't slip, or the magazine just doesn't get printed) and so gave up. Yes, we know that with a bit of investment, the right software, and a few days more work we could have got it working, but deadlines are deadlines. We now have, instead. a reliable and reasonably accurate method of transferring files, albeit a bit Heath Robinson in appearance and style.

Pheonix Rises From the Ashes

While I've talked about telepoint often enough in past Open Channels, it's time to air it yet again. Telepoint is the name for the one-way portable telephone system, which forms part of the CT2 technology.

At one time there seemed to be hundreds of telepoint services (well. four actually) which was, I am convinced, one of the two main reasons why they all failed dismally to attract users. The other main reason was that they were all incompatible with each other. Users of one system couldn't use base-stations of another.

All that's history, and hopefully won't happen again (UK Government permitting!.). Now there is a new system - based on the old ones but much more promising - currently being set up by Hutchison Personal Communications. Planned to launch in the Granada television region at the end of June and expected to spread to highly populated areas by October, Rabbit is the first system to operate on the common air interface (CAI) standard. This means any handsets using the CAI standard should work on the system.

Hutchison appears to have a realistic pricing policy, too. Its basic handset will retail for less than £200, with network charges of £6 per month and call charges comparable with public telephone boxes.

While I never ceased from complaining about the lack of coordination which previous telepoint network operators put into their systems, I have nothing but praise for Hutchison's Rabbit. Despite a bad start, telepoint finally looks set to pay its way. Unless the Government decides to licence another 50 operators with incompatible systems again, that is.

NICAM Starts to pay

Technology found in all NICAM televisions and videocassette-recorders is patented by BBC. This is the good news, for without the BBC it's doubtful that we would ever have stereo terrestrial television. The BBC was instrumental in developing NICAM, and instrumental in persuading television and videocassette-recorder manufacturers to implement NICAM circuits in their equipment.

Naturally, we should expect the BBC to benefit from this. There were over a million television receivers sold last year in the UK alone with NICAM circuits, there'll be double that this year.

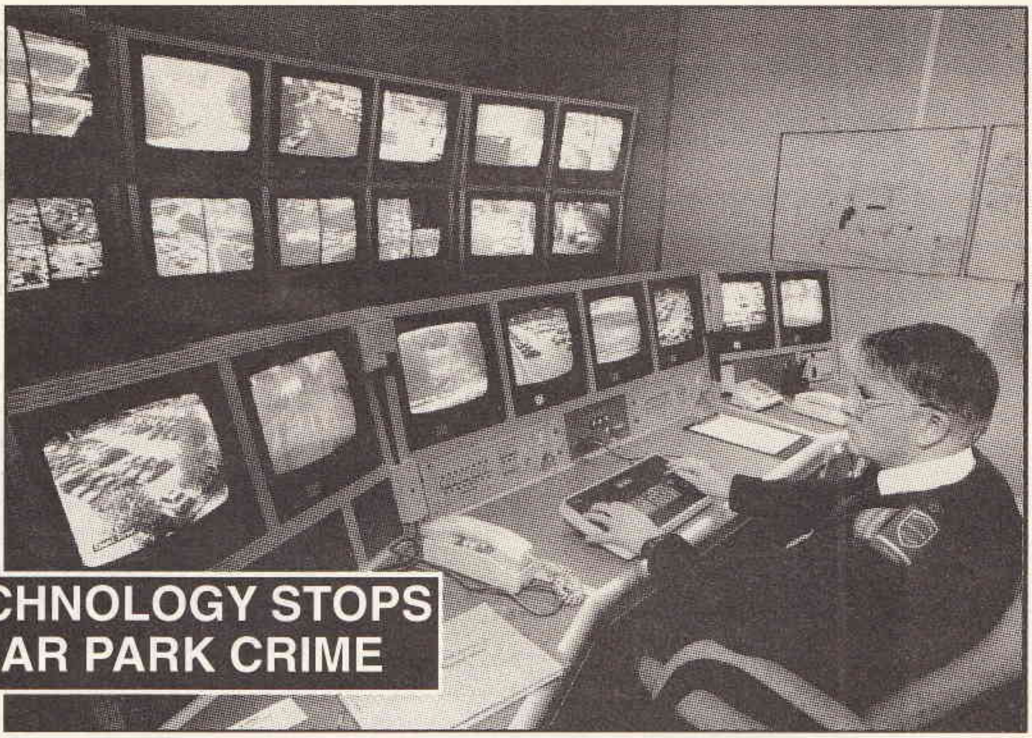
Following discussions with all the related manufacturers, it now looks as though the BBC is set to get royalty payments, which could easily amount to many millions of pounds a year - even if a royalty payment of just £1 per appliance is agreed. The potential is even greater.

Once the BBC receives royalties, they'll naturally be doing the decent thing, and reducing all our television licence fees. Won't they?

Keith Brindley

ETI NEWS

TECHNOLOGY STOPS CAR PARK CRIME



BT's fibre optic technology has helped to virtually eliminate car crime from the places where closed circuit TV (CCTV) cameras have been installed in the Norfolk town of King's Lynn. The project has generated great interest, with visits and enquiries being made from all over the UK and even the USA.

Barrie Loftus, Borough Council Project Officer for the system said: "In the first three months of this year there were 164 vehicle thefts in the King's Lynn area, only one of those was from a car park. On average, there used to be two vehicles a week stolen from the car parks, now that kind of theft has been eliminated."

And he added that it is not just car thefts that have disappeared: "There also used to be three or four incidents of damage to cars or thefts from vehicles on the car parks a week. Since February, when the CCTV system went 'live', there hasn't been one incident of this kind reported."

The council points out that it is not just the installation of the cameras that has contributed to their success. Staff put out a press release, held a formal press day and have made numerous appearances on local TV and in the local press.

The local press has also canvassed local people's opinions. "The majority of them were in favour," said Roy Daynes, crime prevention officer in the town, "What we are trying to do is take away the fear of crime. The cameras are there to make people feel safer as well as catch criminals and vandals. King's Lynn is now a safer place to go shopping."

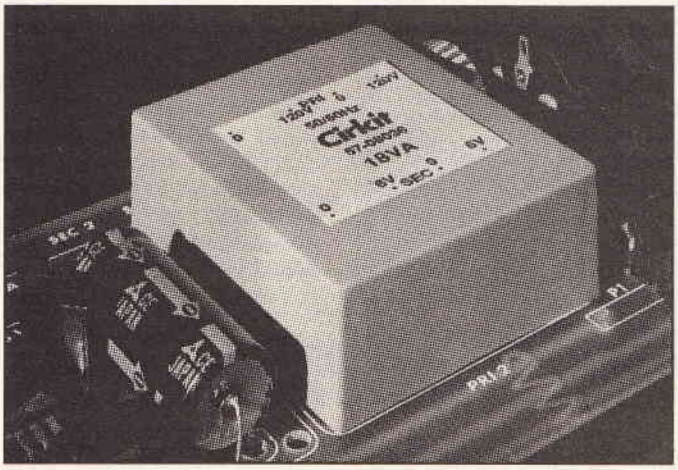
There will soon be a total of 46 cameras in the scheme, all connected to a central control room in the Council's HQ by BT's latest fibre optic technology. In the control room, a single operator, provided by Broadland Guarding Services, monitors movement not just in the town's 17 car parks, but also in a leisure centre, in the area around the council offices and on a major housing estate, Hillington Square.

A manager at the leisure centre remarked: "The level of crime here is far less than in any other centre I've worked in. Most crime is petty vandalism by kids. Now they know the cameras are here and there is hardly any graffiti at all. Moreover, because we can record incidents, we have brought kids into the control room and shown them the screens - word of the scheme spreads more quickly that way.

"More importantly, serious crime in the centre has stopped altogether. At one time we had a whole spate of mountain bike thefts, now I can't remember the last time a bike was stolen. At one stage a new artificial cricket pitch was set alight. There is not much chance of that happening again."

It is sad that yearly running costs are about £180,000 including contract security security staff, BT line rentals and equipment rentals. This is covered by users of the scheme contributing roughly £1,000 for each camera provided, and by a 10p increase on car park tariffs.

LOW PROFILE MAINS TRANSFORMERS



A new range of encapsulated Low Profile Mains Transformers are now available from Cirkit.

Manufactured for Cirkit these high quality, PCB mounting transformers are ideal for applications where space is at a premium, the MA size, for example, is only 44 x 53 x 22mm. Independent primary windings allow 120 or 240V 50/60KHz operation, together with independent secondaries which may be connected for series, parallel or center tap operation, giving a choice of voltage/current combinations. Wound on twin double section bobbins to provide maximum isolation - 4kV between windings and near toroidal characteristics.

These low profile transformers have industry standard pinouts, are encapsulated with UL94V-O self-extinquishing resin and in a rugged moulded case to provide excellent environmental protection and designed to BS 415, VDE 0550-0551, IEC 65 and BS 3535, Class 2 specifications.

The stock range covers 4VA to 30VA in the most commonly used voltages with custom wound versions available to special order.

Further information contact Cirkit Distribution Ltd, Tel:(0992) 441306

CD COMPATIBLE HEADPHONES

Maplin Electronics has introduced a pair of high quality headphones intended for listening to compact disc and digital music material. The drivers feature high energy magnets and diaphragms designed to bring out the best in the dynamic range and superior signal to noise ratio of compact disc.

The unit maintains sensitivity and stability at low volume levels. It has an adjustable padded steel headband and leather earpads, and a 2.7m long extra flexible lead terminated in a 3.5mm stereo jack plug. A 1/4in. jack adaptor is included.

The phones have a frequency response of 16Hz to 22kHz and cost £14.95.

$2MILLION GRANT TO DEVELOP FRACTAL TRANSFORM VIDEO CHIP FOR HDTV

The US government has awarded Iterated Systems a $2 million grant to help develop a special Fractal Transform video decompression chip for use in high definition (HDTV), videophones and multimedia computer systems.

The fractal chip project, funded by the National Institute of Standards and Technology (NIST) under the Advanced Technology Program, is called High Fidelity Digital Image Compression. The chip will decompress digital images fast enough to keep up with the frame rate of television.

Alan McKeon, Managing Director of Iterated Systems Ltd, the UK subsidiary commented "The next step is developing a Fractal Transform chip to compress video images for HDTV in real time, to save the production industry time and money."

Referring to the strategic importance of projects funded under the program, George Bush, President of the USA, said "Today our government must help carry that research forward and contribute to the development of generic technologies that build on basic discoveries." He continued "In this way, we can help leverage the R&D of the private sector, helping whole industries advance in an increasingly competitive global market."

INTERACTIVE TRUCK DRIVELINE CONTROL

Lucas Automotive Ltd., UK, is introducing Electronic Control Units for the control of functions such as fuelling, transmission and braking. As a result considerable improvements in vehicle performance for both passenger cars and trucks. However, further improvements in driveline performance are possible if greater interaction occurs between the sub-system controllers than is currently the case.

This opportunity has been grasped by Lucas who in 1989 embarked on a collaborative research project with two other international vehicle component suppliers, to explore the issues surrounding truck driveline performance by use of high speed data bus communications between driveline control systems.

In the first phase of work a mid-range diesel engined truck has been equipped with a Lucas electronic diesel engine management system, an electronic automated mechanical transmission and electronic ABS/ASR control. The Electronic Control Units for these three sub systems have been interconnected via a high speed 'Controller Area Network' (CAN) databus in order to allow rapid controller interaction. This data bus effectively replaces the electromechanical linkages which would otherwise be necessary for the transmission and braking controllers to change the engine torque during gear shifting and ASR operation; in addition it allows much greater exchange of information between the control units. This significantly improves the driveline performance.

At the outset of this work, no agreed standards existed for message structures for the transmission of data in interactive systems of this type so the project team developed their own message format.

The vehicle fitted with the interactive system has demonstrated significant improvements in driveline dynamic performance, particularly during ABS/ASR operation, owing to the improved communication between the engine, transmission and braking controllers, compared with that obtained with sub-systems connected by conventional electromechanical linkages. The mechanical and wiring complexity is also reduced.

Further developments to the interactive control strategies are also being made to further improve the dynamic driveline performance and improve fuel economy.

For further information please contact:-
Don Hiett at Lucas Automotive.
Tel:021-627 3939

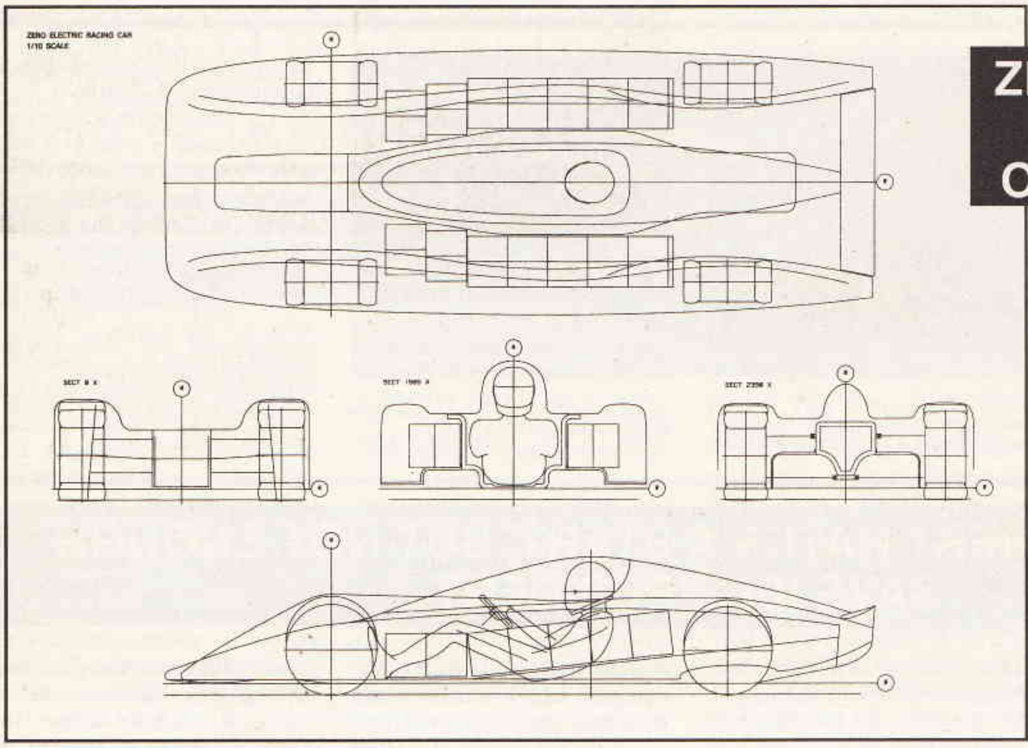
LIVE VIDEO LINK IN UK CIVIL COURT

Evidence will be given by live video link for the first time in a civil case in the UK on Wednesday 13th May.

The County Court is to sit on that day in the Bar Council's videoconferencing studio at 3 King's Bench Walk in the Temple. The court will hear evidence in the case of Henderson v SBS Realisations Ltd. live from a studio in the Massachusetts Institute of Technology in the United States. The case concerns a claim for damages arising from the disappearance of an antique clock from a family home in England.

The County Court was allowed to sit in the videoconferencing studio following a decision in the Court of Appeal.

Bernard Weatherill, Chairman of the Bar Services Committee, which has pioneered the use of live video links in the courts in the UK, said, "This is an important first. The use of videoconferencing can make justice much more affordable, not only in cases where witnesses are abroad, but also where they are in different parts of the UK. I hope that the use of this technique will spread as a natural development of the legal process."



ZERO EMISSION RACING ORGANISATION

A new company has been formed called ZERO Ltd. ZERO stands for Zero Emissions Racing Organisation. It is the aim of ZERO to design, build and develop an electric racing car. The desire to build such a vehicle, has come from the clear objective for the companies involved to remain firmly at the fore-front of technology.

Those involved in this new company are all well known in the Motor Industry. Day to day running of ZERO will be in the hands of Managing Director, Gerard Sauer. He will also be responsible for the gathering of research-data and information. Building of the chassis and its auxillary components will be undertaken by Lola Cars. All electronics, electric motors, and batteries will be developed by Zytek Systems.

The company should have its first prototype car ready by the beginning of September 1993, in readiness for the 1994 competition season.

Commented a spokesman for Zytek Systems: "A major part of Zytek Systems' business is in the development and supply of electronic equipment for use on competition vehicles worldwide. The electric vehicle is viewed by many as the future of road transport and it is therefore logical that Zytek should combine these technologies in the development of the ZERO electric racing car". A spokesman for Lola Cars commented: "We see this as part of our policy of always looking forward - environmental issues are not going to go away, so the sooner we seriously investigate the technology of zero-emission racing cars, the stronger our position will be when 'green' pressures really start to bite. There are fascinating challenges involved in this project, some of which will undoubtedly lead to technology 'spin-offs' for our more conventional racing programmes".

Arthur Large will be responsible for the funding of the company, its City contacts, and also acts as company secretary. Commented Automotive Consultants founder, Gerard Sauer: "Over the last 6 years, we have acted as marriagebrokers between many 'leading-edge' technology companies. Our role as coordinators in the project suits us particularly well, since it represents the highest technological challenges undertaken by the foremost in this field. It is also to record that with Great Britain the present undisputed leader in the field of international motorsport, my company will be directly involved in the first real effort to maintain that position in the future".

For further information contact Gerard Sauer on 071-7371769.

RESEARCH ON RSI

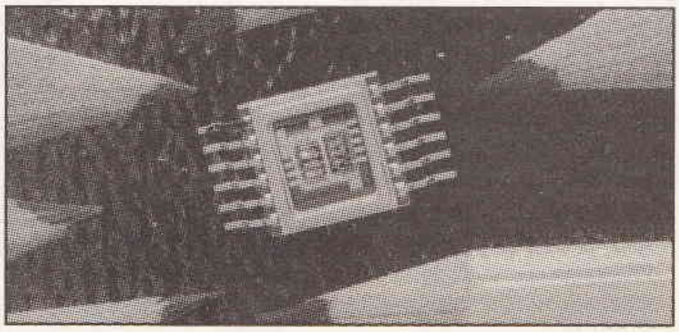
Simon Bull, Final year student at Leeds Polytechnic has carried out his research project on the subject of 'Repetitive Strain Injury' or RSI. This subject is something on which information is extremely limited, and yet the consequences of misinformation or even ignorance could be devastating to many companies. Repetetive Strain Injury' or 'RSI' is just one of the many names used to describe damage to the soft tissue of the hand, wrists, arms and neck. Although recognised in 'blue collar' workers since the 1970's (having been around for thousands of years) the cause of this damage is widely disputed where 'white collar' workers are concerned. It has recently manifested itself following the introduction of electronic keyboards in many industries. The arguments in industry, essentially lie along a Management/Union continuum. Blame is placed on 'poor ergonomics' and 'bad management' by the unions and on 'mass hysteria' or 'malingering' by management.

There are many weird and wonderful theories that can be placed along this continuum although, there are three main factors that tend to occur regularly, These are; poor ergonomics and posture, 'stress' (or other psychological factors such as dislike of the job) and social activities involving hand/ arms etc (e.g. sports such as squash). Until scientific evidence suggests otherwise, it is the authors belief that all of the above factors should be considered in any one case in order to establish the cause.The prevention and cure of RSI are also contested. Prevention is seen by some to simply mean a change of work systems job rotation and screen breaks, and by others to mean ergonomic assessment and the purchase of new equipment. What, clearly, is missing is an assessment of the whole situation before implementation of new technology is even considered.
RSI has past that stage in many cases and what is now needed is a change of attitude and increasing awareness so that all the parties involved can help their own situations. Journalists must use what equipment they have to the best of their ability by learning the importance of posture and sensible practice. Management must perform assessment of workstations (ergonomic and environmental), provide equipment where needed and most importantly of all make available information to it's employees.

Both common law and the impending EC legislation will probably have the most effect on both RSI and the companies involved although with a lead time of four years the legislation's impact is limited. The cost of the 'common law' approach will undoubtedly be very high with possible settlements of five times each persons salary, not to mention the time involved and wasted human resources.

NEW 1.9GHz POWER IC

Mitsubishi Electric UK is announcing the launch of its 1.9GHz power amplifier IC designed specifically for 1.9GHz cordless telephones presently used in Japan and soon to be adapted for the European DECT system.



MGF7121 GaAs power amplifier ICs are compact surface mounting devices. They measure 8.4 x 13.2 x 1.7mm and yet are capable of delivering a minimum power output of 22dBm.

The 1.9GHz ICs operate on a low supply voltage of just 4.6V. Operating frequency range is from 1.85 to 1.95GHz. Input power is -15dBm and typical efficiency is 30% for the Japanese digital cordless.

Typical efficiency will be 50% for the European DECT system as Class B or C operation is possible. Minimum power gain is quoted at 37dB with minimum control voltage rated at -4V.

CUT IN INTEREST RATES A START, BUT NOT ENOUGH SAYS IEE PRESIDENT

"The Government's cut in interest rates might provide a welcome boost in confidence for the business community, but it is not enough to solve the problems faced by Britain's manufacturing industry".

This was the warning given by Brian Manley, President of the Institution of Electrical Engineers (IEE). Mr Manley was addressing senior industrialists at the Annual Meeting of ERA Technology held at the Butchers Hall in the City of London.
"Unless the situation is resolved long-term" he said "high and fluctuating interest rates, the failing courage of the high street banks and the short-term approach of the City of London will continue to be the major causes of Britain's decline as a manufacturing nation".

Mr Manley criticised UK banks for their failure to support new industries.

"To be lending money like drunken sailors one minute and to be withdrawing support from viable businesses the next - not because the company has failed, but because the Bank has lost courage - is no recipe for the successful development of enterprise."

He also cited high real interest rates as being the dominant factor in discouraging manufacturing industry from investing in enterprise - as a result British companies are forced to achieve high short term results 'far beyond those that would be smiled upon in Japan'."There is no way that the British company can compete effectively" he said. "Opportunities for innovation that are grasped with enthusiasm overseas are passed over in Britain as offering insufficient return, or posing too high a financial risk".
Research and development, Mr Manley continued, is progressively viewed as an 'unaffordable luxury' engineering innovation is not seen as a priority and as a result the country is losing its ability to respond to international competition.

Although welcoming yesterday's cut in interest rates Brian Manley urged the government to establish a system of taxation which rewards long term investment and differentiates between capital growth and dividend distribution. He also stressed the need for new policies which would exploit the stability provided by the ERM and would lower interest rates to international levels as quickly as possible.

NEW WAY TO DECODE SIGNALS

Ampsys Ltd. - a small spin-off company of Paisley College, Scotland - has announced a new method of decoding narrow band FM signals. This patented technique removes the spikes (or spurs) from the radio signal just as the link is breaking down and also provides inaudible squelching. These two effects greatly reduce listener fatigue especially over sustained periods. Other advantages include a 25% increase in usable range for any FM PMR and/or a reduction of 50% transmitted power for the same reception range.

A further improvement is that co-channel interference is reduced and two FM channels on the same centre frequency can now be decoded without mutual destruction.

The circuitry is currently under evaluation by a number of multinationals. One evaluation engineer was heard to remark that "By the year 2000, every FM radio will be using this technique
The new decoder uses an Amplitude Locked Loop in combination with a Phase Locked Loop to achieve this breakthrough.

For further information contact Ampsys Ltd Tel: 041 848 3444

FRACTALS MAY OVERTURN EXISTING IMAGE COMPRESSION STANDARDS

The latest Gartner Group research report now gives Iterated Systems' Fractal Transform image compression technology a 50% probability of setting the de facto standard. This is an increase from their previous assessment late last year of a 20% probability. The ramifications could be severe for existing standards including JPEG, MPEG and H.261.

The report, from this respected strategic consultancy, stated "For the imaging and stored multimedia markets, we are increasing our probability to 0.5 that fractals will rule." It continued "We base this primarily on the significant improvements of fractals compared to anything else, including discrete cosine transform-based algorithms and the recognition of Microsoft for this technology."

The report concluded that Companies requiring very high compression ratios for high resolution images should investigate Iterated Systems' products today."

Bill Caffery, Vice President of Advanced Technologies at Gartner Group said "Multimedia is unquestionably an arena where the unrivalled superiority of fractal decompression is a boon to end users or consumers, and therefore to developers and publishers."

Alan McKeon, Managing Director of Iterated Systems' UK subsidiary, commented "This is further acknowledgement of the momentum which is establishing the Fractal Transform as the de facto standard for high performance, low cost still image compression."

Iterated Systems, Inc. a privately held company headquartered in Atlanta, provides fractal compression products that enable computers to handle pictures in a resolution independent manner at ultra.high compression ratios. The company is committed to fundamental mathematical research to establish and advance the Fractal Image Format. Iterated Systems Ltd is the Reading-based UK subsidiary.

BT WIDENS CASHLESS CALLING OPTIONS-PAYMENT FOR CALLS BY VISA CARDS

BT and Visa International today announced that more than 280 million Visa cards could soon be used to pay for calls from BT telephones in the UK. In addition, cardholders will be able to use their Visa cards to phone the UK from abroad using the UK Direct service.

Under the arrangement, BT and Visa will extend the VisaPhone service -which is currently available in the United States - to the UK by early 1993. 'This means that the UK, which currently has 28 million Visa card holders, will be the first European country to enrol in the service.

"The service will appeal particularly to visitors to the UK, who will not need a BT telephone account as the cost of the calls will be billed to their Visa card in their local currency.

The new facility will complement BT's own Chargecard, and will use the same calling method. Users will dial 144 in the UK or, when abroad, special freephone numbers from over 60 countries.

For security, callers will have to provide their card number plus a telecode - a four-digit personal identification number (PIN). This number will be used for making telephone calls only and will be different from the PIN used for cash dispensers.

Tony Vardy, BT's Director of Service Development, said: "Cashless calling is becoming vital to our customers in the UK and to travellers all over the world.

"At BT we strive to ensure that everyone who needs to use our network, and that includes every visitor to this country, is given the best service and the most convenient method of payment.

"With VisaPhone, which complements our existing Chargecard services - plus our facilities for using commercial credit cards and pre-paid Phonecards at selected Payphones - BT will offer customers the most flexible range of telephone card services in Europe."

WORLD'S FIRST 1.3" DISK DRIVE MODULE

Hewlett-Packard has introduced the HP Kittyhawk Personal Storage Module - the world's first 1.3in disk drive, capable of storing 21.4 Mbytes of information in a device the size of a small matchbox. It combines the miniaturisation, durability and removability of solid-state technology with the cost and power advantages of disk-drive technology.

Though the HP Kittyhawk PSM uses much of the same technology as larger disk drives in desktop and portable computers, it is less than 0.4 inches high, measures 2 inches by 1.44 inches and weighs less than one ounce.

Aimed at highly mobile applications, the HP Kittyhawk PSM was designed to be rugged using techniques learned at HP's calculator division. As a result, the PSM can withstand an

operating shock of 100g (non-operating of 225g) to survive a system drop of approximately three feet. In comparison, today's 1.8-inch and 2.5-inch disk drives can handle operating shocks between 10-20g.

Our customers demanded a no-compromise storage design," explained Anna Tunnicliff, 11P Product Manager. "They needed high resilience and dependability - to provide solid computing devices for their customers.

To achieve this, the drive incorporates a special, HP-designed integrated circuit based on the same technology used in cars to sense collisions and trigger airbags. As a result, the HP Kittyhawk PSM can sense and brace for a shock before it occurs, keeping data safe.

The dimensions of today's 2.5-in and 1.8-in drives, as well as their sensitivity to shock, limits them as storage solutions for the next generation of mobile computers," continued Tunnicliff.

She pointed out that solid-state technology is approximately five times as expensive as the rotating disk technology used in the HP Kittyhawk PSM. 20 Mbytes of solid-state memory currently costs about $1,000, or $50 per Mbyte, compared to about $12 per Mbyte for the HP Kittyhawk PSM. HP storage devices are already being designed into mobile computing devices, digital imaging peripherals and consumer electronics.

Designed for long life, the storage device features a mean-time-between-failures (MTBF) of 300,000 hours. It offers an 18 millisecond average seek time and supports active, idle, standby and sleep modes.

Further information can be obtained from the Customer Information Centre Tel:0344369222

ETI

NEWS

...Stateside...

Preventing electrical surges

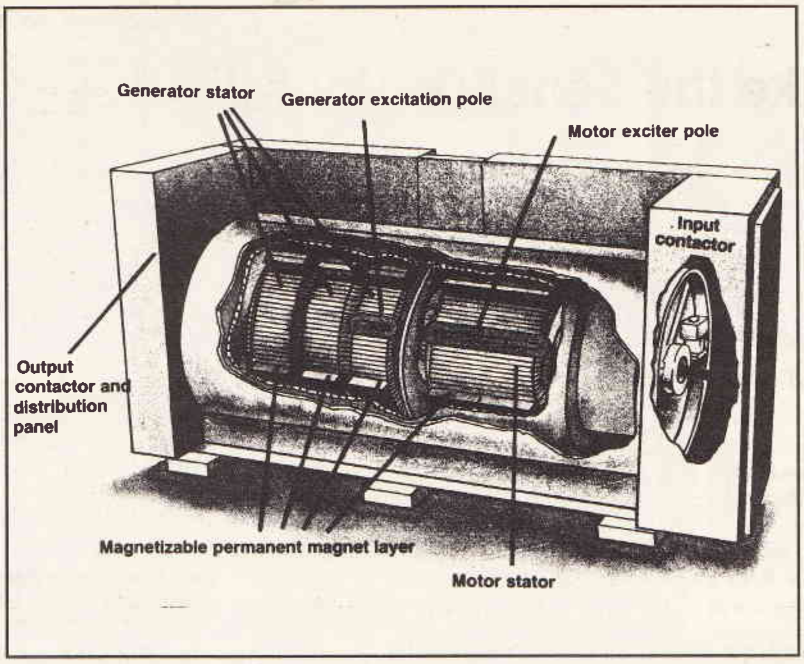
An optical fibre manufacturer found that outside lightning storms and other weather conditions, combined with a large number of DC motors inside the plant, continually affected the frequency of plant power by creating harmonic distortions. Upsets in process equipment used in continuous production steps required the production to stop, or could cause irregularities in the final product.

By installing a motor-generator set from Precise Power Corp. of Bradenton, Florida, most of these problems have cleared up, according to a plant engineer at Alcoa-Fujikura Ltd. of Duncan, South Carolina.

Power conditioning or uninterruptible supply can be accomplished in a variety of ways. Traditionally, suppliers have provided a motor-generator set, in which incoming power energized an electric motor, which in turn could be used to run the generator. Interposing a flywheel or other device enabled the motor generator set to overcome momentary outages. Later systems employed inverters powered by lead-acid batteries to get through the outage.

These setups have a number of limitations. The flywheel-based motor generator set provides power for only about 0.3s before frequency drops below 59.5Hz, unless substantial oversizing of the equipment is specified. The battery-based systems require constant monitoring and frequent maintenance, and if heavily used, must have the batteries changed out.

The Precise Power system motor generator set emphasises constant frequency for a minimum of 15 seconds at full-rated load, and 45 seconds at no load. This is often long enough to ride through momentary outages, or to get a backup generator online. Instead of a conventional generator field coil, the device has an excitation coil that 'writes' magnetic field poles into a layer of high-coercivity magnetic material on the inner surface of a rotating annulus core. As rotor speed changes (due to power fluctuations), the number of written poles varies such that a constant frequency is induced in the stator windings.



The advantage of the pole-writing motor over an induction-motor driver are that it provides the efficiency and constant-speed characteristics of a synchronous motor, and that it has the high-starting torque of a high resistance squirrel-cage motor. Squirrel-cage windings embedded in the annular rotor cause the motor to start like an induction motor. When it reaches about 80% of operating speed, the stationary excitation coil is energized at line frequency. This means that a high-inertia rotor can be used (it weighs nearly 1 ton), to provide high ride-through capability.

Improved electroluminescent lamp

Electroluminescent lamps contain a powdered phosphor - usually zinc sulphide - between two electrodes. When alternating current passes between the electrodes, the phosphor emits light.

Although these lamps are used primarily for backlighting, the range of applications has been limited because the phosphors degrade rapidly when exposed to moisture and high humidity. Conventional EL lamps use a desiccant layer and a thick co-polymer outer cover, which is sealed at the edges to protect the phosphor. While this packaging increases lamp life, it does not completely eliminate phosphor degradation because the polymers are permeable. The relatively thick package and edge seals also limit applications.

Durel Corp. of Tempe, Arizona, has recently made advances in EL technology that increase lamp life, improve reliability, and reduce the bulk and cost of the lamp system. Instead of the multilayer approach, each individual phosphor granule is encapsulated in glass to protect it from moisture. Encapsulated phosphors allow engineers to make lamps much thinner and more flexible. These can be cut into intricate shapes since the edges do not need to be sealed. The Durel lamps are also less expensive than conventional EL lamps because the outer plastic film, some components, and manufcturing steps required to produce the lamps are eliminated.

World record critical current density

A superconducting polycrystalline film prepared at General Electric Company, Schenectady, exhibits world-record critical current density (CCD) for thick films, and maintains high CCD in a magnetic field. At 4.2K and zero magnetic field, the CCD of the 3.2m film of thallium-barium-calcium-copper oxide (TBCCO) exceeded 500,000 A/cm^2. At 82K, CCD was 100,000 A/cm^2. The previous high was 60,000 A/cm^2 at 77K and zero magnetic field.

While the CCD of previous superconductors falls rapidly in the presence of magnetic fields due to weak-link coupling of grains in the polycrystalline structure, the TBCCO film retained CCD in elevated magnetic fields, indicating strong grain linkage. This is especially critical for power applications such as motor windings, generators, and magnetic resonance imagers which operate in the presence of magnetic fields.

In the film-making process, a heated polycrystalline yttria-stabilized zirconia substrate is spray-coated with an aqueous solution of barium, calcium, and copper nitrates, and oxidized to produce a Ba-Ca-CuO film. This is reacted with thallium-oxide vapour to erect a superconducting film of the composition $TlBa_2Ca_2Cu_3O_x$ in the predominant phase. This proprietary process lends itself to scale-up in thicknesses and coating surfaces of extended lengths. To date the longest length of oxide produced is only 100m.

READ/WRITE ETI Letters

Hazadous Waste Disposal Suggestions

The letter in Read/Write from A Ward caught my attention in the June issue as I have had the problem of disposal of hazadous waste myself in the past.

Having bought a pack of unmarked regulator ICs from a Tandy store and started to search for data, I found a similar toxic substances warning on the data sheet for the TO3 style five terminal regulators. Fortunately, at the time a relative was employed by a fire exstinguisher company, who happened to have a toxic waste container on sight, and permission was obained to dispose of the devices in this way.

It is an intersting point, that there are a number of electronic components which contain toxic substances, and the labelling is not all it could be, good examples of this being Ni-Cd rechargeable cells, which contain large quantities of easily disposable cadmium, also fluorescent lighting tubes which rarely if ever carry any cautionary markings about the Mercury vapour and rare earth elements they contain., not to mention smoke detectors which contain a small quantity of radioactive material.

In recent years civic ammenity sites (refuse container sites) have become better equipped with in addition to household and garden refuse compactors and scrap metal skips as standard have been augmented by the provision of waste clothing containers for charities and waste glass and paper containers for recycling. Unfortunately the nearest toxic waste disposal is the provision of waste engine oil containers. I suppose that if you had no other choice, then throwing such devices in this container, whilst probably illegal, would at least prevent them from corroding, and there is always the possiblity that the people who process waste oil would be better equiped to deal with the hazadous materials than any other sectors of the salvage industry.

There are companies which specialise in scrapping computer junk, who are mainly interested in the gold used on the boards. Inevitably they will handle quantities of components containing toxic materials, whether they realise this fact is another matter.

Most scrap merchants will have information on companies which specialise in computer scrap as due to the difficulties involved in recovering the valuable metals.

Perhaps the local fire officer could help as it is his job to know of any premises in his area.

An excellent starting point for enquiries of this nature is the Environment Department in the FAX directory. Friends Of The Earth might also tell you.

I hope this research is useful.

I Field,
Letchworth,Herts

Fog Factor Not So Clear

Regarding the fog index talked about last month.

There have been many other tests of readability. In 'The Art of Plain Talk,' Dr Rudolf Flesch argues that the closer a word is to its root the easier it is to understand. This implies that prefixes and suffives added to words will 'decrease' their readability.

Unfortunately, readability, like style, is not simple to classify or quantify. Sentence lengths and syllable counts do not provide a reliable guide. Syntactic analysis alone, especially at word level,is inadequate because it ignores semantic and pragmatic considerations.

For example, in the two 11-word sentences below, the first is classified by the Fog approach as difficult to read since it has 11 words and some 27 syllables.

1. Marilyn was undoubtedly realizing that individual protests had become totally ineffective.

2 The girl saw the man the boy saw riding his bicycle.

However, the second sentence, also 11 words, has mostly single syllable words except for two, yet is considerably more difficult to read and is highly ambiguous, for example:

a)Did the boy see the man while he (the boy) was riding his bicycle?

b) Did the boy see the man riding his (the boy's) bicycle?

c)Did the boy see the man riding his (the man's) bicycle?

Various parts of written English have been counted. Back in 1852 Augustus de Morgan (Professor of Mathematics at the University of London) suggested word length might be a distinctive trait of a writer's style. In 1887, Mendenhall (Professor of Mathematics at Ohio State University) was counting the absolute word lengths of some 5000 words from "Oliver Twist" Yule (1938 and 1944) studied sentence lengths and counted nouns. The Waring-Herdan formula (1969) attempted to analyze and account for style by looking at ascending and descending features in concordances of texts. The publications, "Computing in the Humanities", and "Literary and Linguistic computing", have also carried numerous mathematical approaches; all complex and unconvincing.

I am presently conducting post graduate research into the style and readability of written English prose, hence my interest. Believe me - volumes have been written on the subject: I know, I've had to read the damn things!

My own approach is multi-level . I have written a suite of computer programs which will take any English text and provide the following analysis:

1 Number of sentences
2 Number of words
3 Number of syllables
4 Number of letters
5 Average sentence length
6 Average syllables per sentence
7 Average syllables per word
8 Average letters per sentence
9 Average letters per word
10 Average letters per syllable
11 Total number of words (Tokens)
12 Total number of different words (Types)
13 Type Token ratio
14 Full concordance listing of all words (user may choose the concordance to be listed
a) unsorted
b) alphabetically ranked
c) frequency ranked
15 Frequency of each word in descending order
16 Percentage of the total text accounted for by each word
17 List of all auxiliary and modal verbs in various forms
a) full - positive
b) full - negative
c) contracted - positive
d) contracted negative
18 The 30 most commonly used words in English (according to the LOB corpus of 1 million words) and a direct numerical comparison of the usage of those words in the text under analysis
19 Individual count of each letter used
20 An individual count of letter pairs printed out as a two dimensional table in a 26x26 matrix

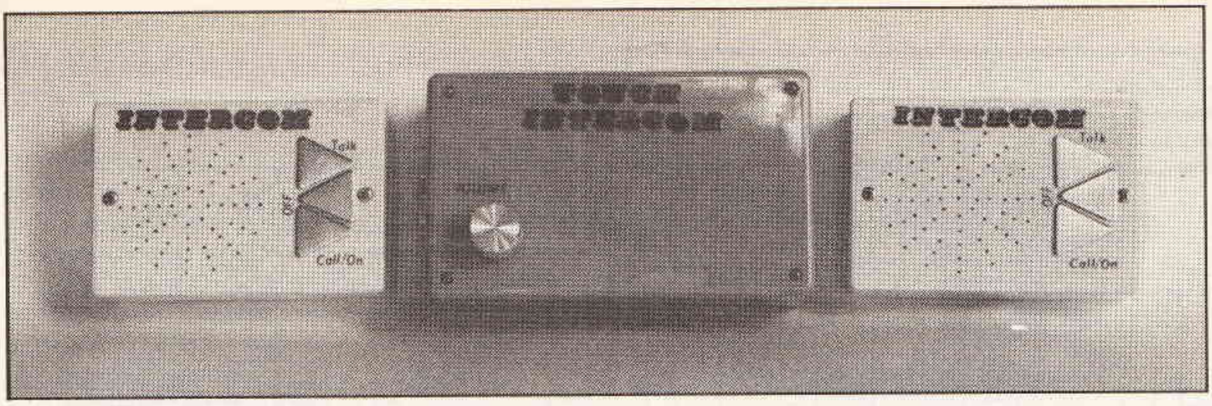
The resultant analyses, together with the statistical analyses of them, are of some complexity yet these are only one level of the total analysis; there are another three levels of equal complexity.

The few words above barely scratches the surface of the subject but it can be seen that it is unwise to classify a given text merely on the basis of a couple of counts.

English just cannot be reduced to a simple mathematical equation.

David Gibson,
Little Gaddesdon,Herts

PROJECT



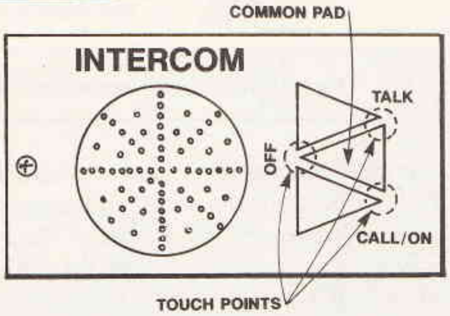


Fig.1 A remote station showing touch points

Touch Controlled Intercom

by Colin Meikle

This is an intercom with some interesting differences from most common intercom designs. It was designed as a simple two way intercom for use around the home or between outhouses and the home. As with most intercoms only one person can speak at a time but this intercom has two way control i.e. either person can control it. Other features include touch

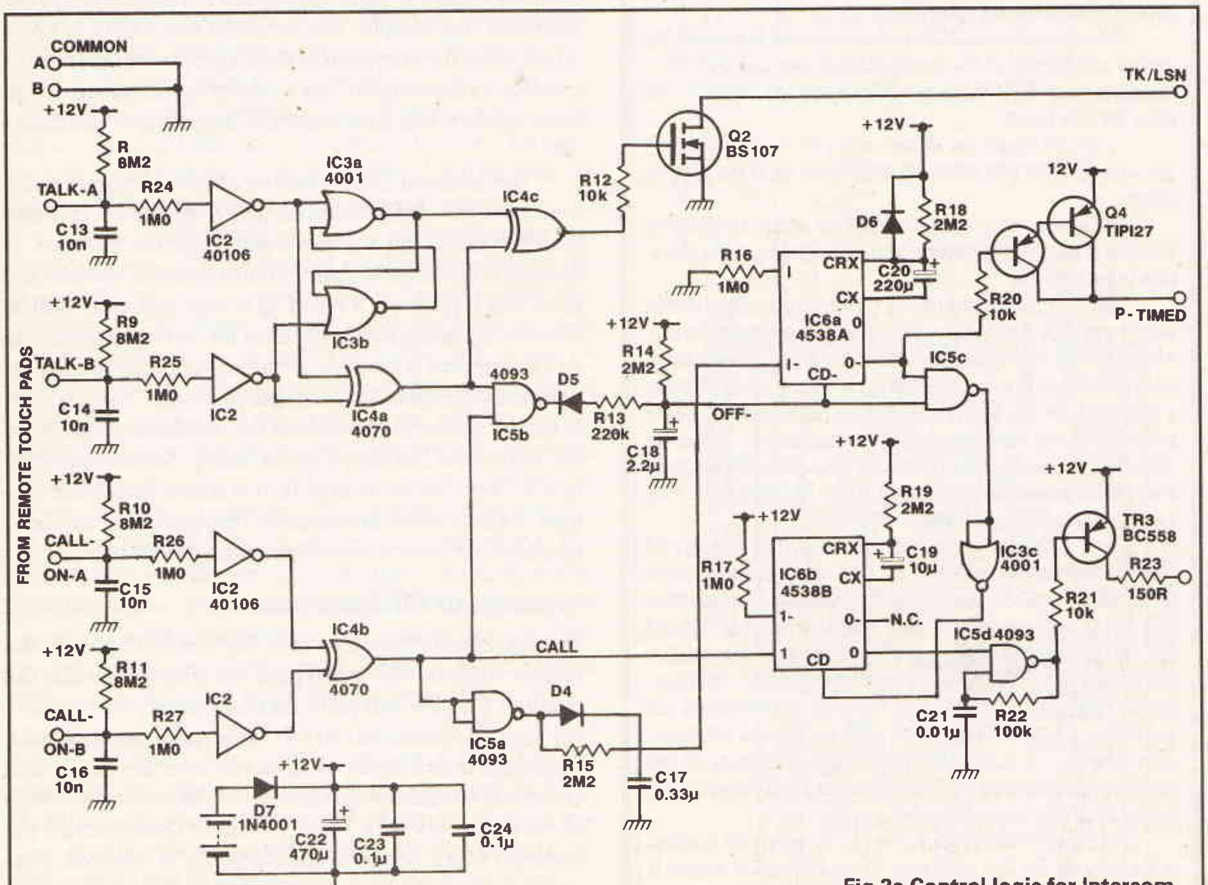


Fig.2a Control logic for Intercom

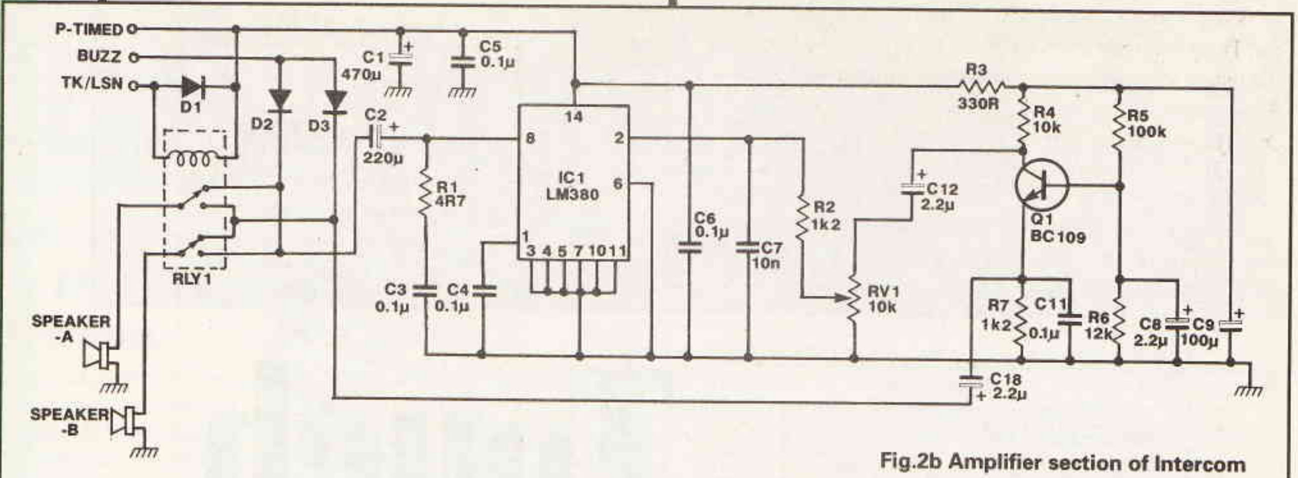


Fig.2b Amplifier section of Intercom

HOW IT WORKS

The intercom consists of two distinct parts, the control logic and the amplifier section each has its own PCB. The circuit diagram is shown in Figure 2.

The amplifier section consists of two parts, a common-base pre-amplifer followed by a single IC power amplifier. The speakers in each of the remote stations act as a microphone as well as a loud speaker. The function of the speaker is determined by the position of the relay RLY1, which is under the control of the control logic.

The input signal from the relay is fed into a common base preamplifier, which consists of Q1 and associated components. The amplifier increases the input signal to a level which can drive the power amp, a gain of approximately 100. This amplifier configuration is suitable due to the low input impedance of the speaker (8 ohms). The amplified signal is fed into RV1 which allows the volume to be controlled. The resultant signal from RV1 drives the input of the power amplifier IC1. The output from IC1 drives the appropriate speaker via the relay.

The potentially long cables which are used to connect the control unit to each of the remote stations can pick-up a lot of electrical noise (50Hz mains and RF). C7 and C11 decouple this noise from the signal.

D2 and D3 couple the 'Buzzer' tone from the control circuit into each speaker line, when the amplifier is on, these have no effect.

The control logic turns the amplifier on/off, controls the direction of speech (by switching a relay) and produces the buzzer tone to each station.

The touch pads on the remote stations are arranged so that when a pad is touched the user 'shorts' a signal pad to ground with a finger tip (the middle pad is ground, the two other pads are signal pads – see Figure 1). Each of the four signal lines are pulled high via a high value resistor (R8-R11), when a pad is touched the line is pulled low (due to the skins relatively low resistance – 100k or so) IC2 buffers the signal to give a good logic level. C13 to C16 decouple any noise to prevent false triggering, R24 to R27 protect IC2 from damage due to ESD (Electrostatic Discharge).

When a Call/On pad is touched, timer IC6b is triggered enabling the oscillator formed by IC5d, this oscillator drives QR3 which produces the Buzzer tone. If the call/on pad is held timer IC6a will be activated, after a short time determined by R15 and C17. This timer will turn QR3 on and therefore enable the amplifier and disable timer IC6b, therefore stopping the buzzer. The amplifier will be shut off after timer IC6a times out (determined by C20 and R18) or if both talk and call/on pads are touched simultaneously. This causes the OFF line (reset signal to IC6a) to go low therefore resetting the timer. This signal has small delays built-in to stop accidental triggering.

The logic formed by IC3a, IC3b and IC4c control the direction of the relay via Q2. This logic ensures the direction of speech is always correct no matter which station is controlling the intercom.

control and automatic turn off. The remote stations have been designed to be suitable for use in harsh environments.

Most simple intercom systems are of the Master/Slave type i.e. one unit, the master turns the units on/off and controls the direction of speech. The slave can only call the master, alerting someone to turn the intercom on. This setup is often very inconvenient, especially for the person at the master end who has to answer the intercom when some one calls.

This design overcomes the above problem by effectively making both units physically and functionally identical. Either unit can call the other unit, turn the intercom on/off or control the direction of speech. All controls (except volume) are via three touch pads mounted on each unit, see Figure 1; these touch pads make controlling the intercom very simple. The intercom has a built in timer which turns the intercom off after a predetermined time (5 minutes on prototype). This is a very useful feature if you have children who have a habit of forgetting to turn things off.

The intercom itself consists of three separate units, two identical remote stations and a central control unit which contains all the electronics and the batteries to supply it. The remote stations only contain a speaker and three touch pads. As a result, it is very robust as well as reliable. By using mylar speakers the remote stations can be used almost anywhere including damp environments like sheds, greenhouses, even the bathroom. The intercom is battery powered, this avoids the problems of having to site the control box near a mains outlet. Current consumption is very low in standby (a few micro-amps) and less than 100mA when in operation (the prototype has been operating for over a year on the same batteries).

Operation Of Intercom

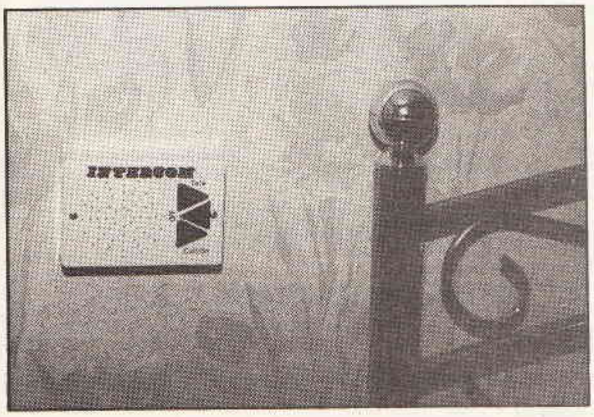
The operation of the intercom is done via touch pads on the remote stations. The touch pads are arranged so that the operator touches two pads simultaneously, see Figure 1 (to turn the intercom off all three pads are touched). Operation is as follows: A brief touch on the call/on and common pads causes the intercom to put out a tone to each of the remote stations. This tone can be used to attract the attention of the person at the other end, if required. If no one turns the intercom on then the tone will terminate after 30 seconds.

A sustained touch on the call/on and common pads turns the intercom on, after a two second tone has been put out. This short tone stops someone turning on the intercom and listening in on your conversation without your knowledge.

The direction of speech is controlled by touching the talk and common pad; touch to talk release to listen.

The intercom is turned off by touching all three pads simultaneously. The intercom will automatically turn off after a predetermined time (approximately 5 minutes), if not turned off manually.

Both units are controlled in the same way, either remote station can control the intercom, no matter who called or answered.



Construction And Installation

The intercom consists of two PCBs, one for the control logic and one for the amplifier. Construction of each is straight forward, the component overlay is shown in Figure 3. Note the polarities of the polarized components, especially the capacitors on the amplifier board. All the ICs can be mounted in sockets if desired except for IC1, where it is better to solder it directly to the board as the tracks act as a heatsink. The interconnections between boards is shown on the component overlay and the connections to the remote stations is shown in Figure 4.

Figure 4 shows the suggested dimensions for the remote stations. To obtain neat results when drilling the array of holes for the speakers, cover the front panel with masking tape then draw six concentric circles 5mm apart (starting with a 5mm radius). The circles should now be divided into eight equal segments and 1.2mm holes drilled at the interceptions of the straight lines and the circles. When the boxes have been drilled the lettering should be added (dry lettering covered with a coat of lacquer to protect it). When completed the speaker can be glued in place and the wiring to it and the touch pads added. The remote stations can now be connected to the control box, standard 4-core telephone cable is suitable. You may wish to add suitable connectors to the control box (eg DIN sockets/plugs) rather than hard wiring it to the remote stations. A suggested layout for the control box is shown in Figure 5. Try to keep the wires going to RV1 short (use screened cable if the wires are longer than 1") and if possible solder the ground wire to the metal can on RV1, as well as to the tag. Be careful of ground loops if you deviate from the suggested connections. For example do not connect the ground returns from the remote stations (common wires) together, at any point.

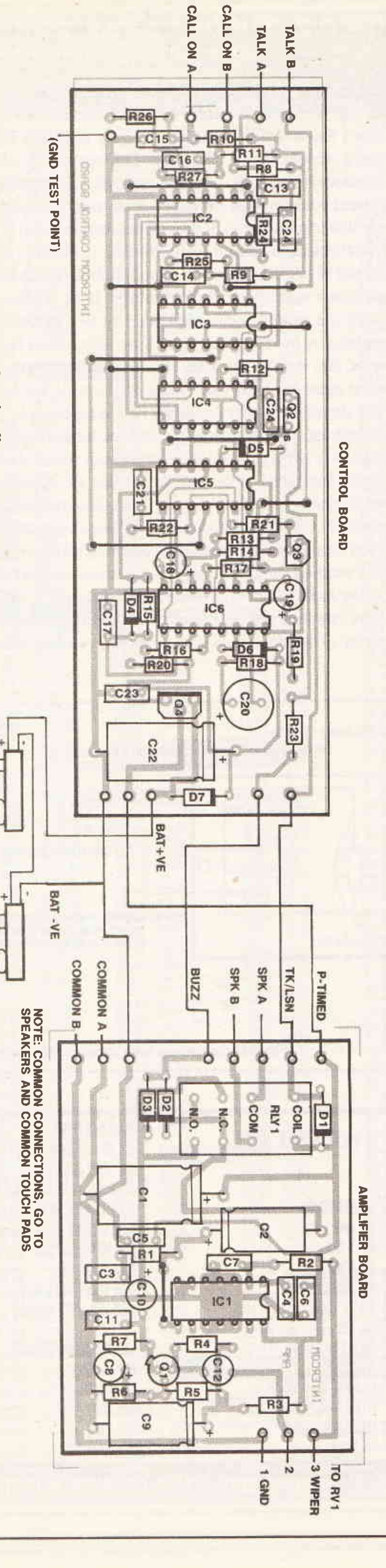


Fig.3 Component overlay and connection detail

Testing Operation

When complete, place both remote units near each other and turn the volume control down. Note: if you are testing the circuit before you have mounted the speakers in their boxes the sensitivity will be adversely affected. The first check should be to check the current drawn with the intercom in standby. With the intercom off (standby) the current consumption should be a few micro-amps (if you are able to measure it), if the current consumption is more, first check the intercom is off – turn the volume up, nothing should be heard from either remote station (touch all three pads on a remote station to turn off). If all is OK briefly touch the call/on – common pads on one of the remote stations. A buzzing tone should be heard from both stations, this should stop after 30 seconds or so. Now do the same again but hold your finger on the pads, the buzzing tone should stop after 2 seconds (if it is replaced by a loud screeching sound you have the volume set too high). If the volume is slowly turned up a screeching sound should be heard (feedback), if the remote stations are moved apart, considerably, this should stop. If the talk/common pad is touched a few times the direction of speech should change – you should hear the relay click on/off. If all three pads are now touched simultaneously the intercom should turn off. Do the same checks for both remote stations, if all is OK move one station to another room and get someone to talk to you. (The intercom is very sensitive, you do not have to be standing next to a remote station or shout at it to be heard).

When in operation there should be no problems, however a few precautions should be noted. If any of the remote stations are to be used in very damp environments (e.g. the bathroom) you may need to move the touch pads further apart to stop condensation activating the intercom. When routing the cables for the intercom try and avoid routing them beside long runs of mains cables. Also try and avoid positioning any of the boxes, or cables next to electrical equipment that generate a lot of electrical noise for example televisions.

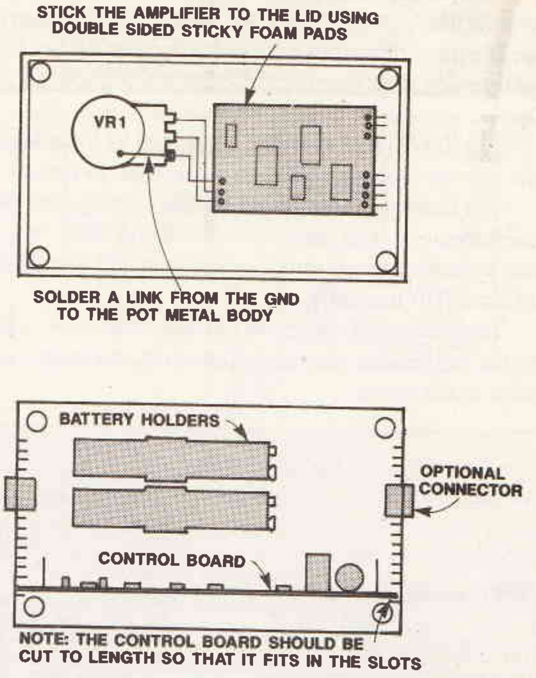


Fig.5 Layout of the control box

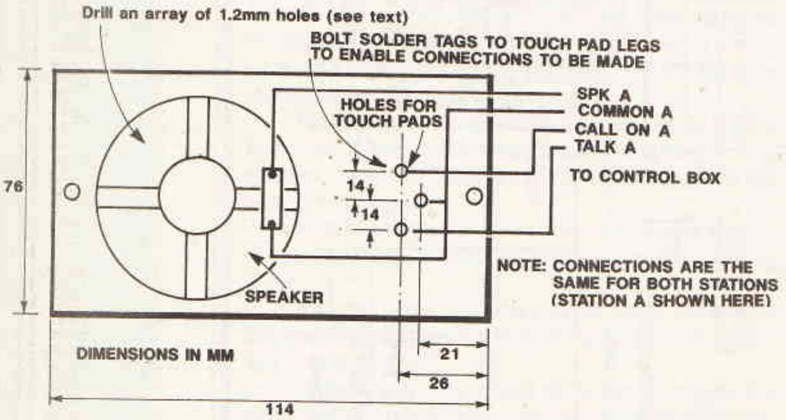


Fig.4 Dimensions and connection details

BUYLINES

The mylar speakers and relay can be obtained from Rapid Electronics (Tel: 0206 751166)

PARTS LIST

RESISTORS
R1 4R7
R2,7 1k2
R3 330R
R4,12,20,21 10k
R5,22 100k
R6 12k
R8,9,10,11 8M2
R13 220k
R14,15,18,19 2M2
R16,17,24,25,26,27 1M
R23 150R 0.6W
RV1 10k

CAPACITORS
C1,22 470 axial µ/16V
C2 220 axial µ/16V
C3,4,5,6,11,23,24 100n poly or ceramic
C7,13,14,15,16,21 10n ceramic
C8,10,12,18 22/16V tant
C9 100 axial/16V
C17 330n poly
C19 10µ/16V radial
C20 220µ/16V radial

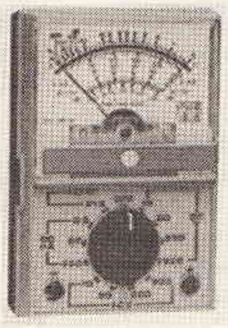
SEMICONDUCTORS
IC1 LM380
IC2 40106
IC3 4001
IC4 4070
IC5 4093
IC6 4538
D1,2,3,6,7 1N4001
D4,5 1N4148
Q1 BC109C
Q2 BS170
Q3 BC558
Q4 TIP127

MISCELLANEOUS
RLY1 12V DPDT relay (ormron G12V or similar)
2 x 8R Mylar speakers (66mm dia)
6 x triangular touch pads
2 x 6V battery holders (4AA type)
2 x 'PP9' type battery connectors if required
8 x AA alkaline batteries
2 x box for remote stations 114x76x38mm
1 x box for control station 150x90x54mm approx

ESR ELECTRONIC COMPONENTS

Station Road, Cullercoats,
Tyne & Wear NE30 4PQ
Tel: 091 251 4363 Fax: 091 252 2296

TEST AND MEASUREMENT



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Dim. 90 × 60 × 30mm £5.96

HM103S ANALOGUE METER
19 ranges (inc 10A dc(, fuse and diode protection, battery test, shock resistant tilted case, mirrored scale, supplied with battery, leads and instructions.
Dim. 154 × 77 × 43mm £11.47

HC2020S ANALOGUE METER
20 Ranges (inc 10A dc), fuse and diode protection, transistor and diode testing, polarity reverse switch, high impact shock resistant case. Supplied with battery, leads, stand and instructions.
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M2315B DIGITAL METER
17 ranges (inc 10A dc), 3.5 digit 12mm LCD, diode test, buzzer, auto polarity and zero, over-range ad low battery indication, supplied with battery, leads and instructions.
Dim. 130 × 72 × 33mm £23.40

TL34 DIGITAL METER
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1.0	—	—	£0.05	£0.06
2.2	—	—	£0.05	£0.06
4.7	—	—	£0.05	£0.08
10	£0.05	£0.05	£0.06	£0.08
22	£0.05	£0.05	£0.09	—
47	£0.06	£0.06	£0.11	—
100	£0.06	£0.09	£0.11	—
220	£0.09	£0.12	£0.31	—
470	£0.15	£0.19	£0.57	—
1000	£0.22	£0.29	—	—
2200	£0.37	£0.57	—	—
4700	—	£1.11	—	—

ELECTROLYTIC AXIAL CAPACITORS

uF	16V	25V	63V	100V
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1.0	—	—	£0.10	£0.10
2.2	—	—	£0.10	£0.10
4.7	—	£0.09	£0.10	£0.10
10	—	£0.12	£0.12	£0.12
22	—	£0.09	£0.13	£0.17
47	£0.10	£0.11	£0.16	£0.20
100	£0.10	£0.13	£0.21	—
220	£0.13	£0.18	£0.42	—
470	£0.21	£0.20	£0.69	—
1000	£0.33	£0.40	£1.05	—
2200	£0.52	£0.64	—	—
4700	£0.90	—	—	—

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0.5W 5% CF E12 Series	£0.95/100
0.25W 1% MF E12 Series	£1.72/100
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5mm Yellow LED	£0.10
5mm Orange LED	£0.10
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common cathode	£1.14

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BNC Chassis Skt	£0.78
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PL259 11mm	£0.58
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SQR UHF socket	£0.40
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F Plug RG6	£0.27
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25 Pin	£0.48	£0.50
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LM381	£2.70
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LM387	£1.60
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LM393N	£0.28
LM394	£3.98
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NE531	£1.56
NE5534	£0.66
NE556N	£0.36
NE567N	£0.36
UA733	£0.64
CA741CE	£0.18
CA747CE	£0.39
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74LS113	£0.21
74LS114	£0.21
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74LS165	£0.48
74LS179	£0.30
74LS173	£0.24
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74LS175	£0.24
74LS190	£0.25
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74LS193	£0.24
74LS195	£0.24
74LS196	£0.24
74LS197	£0.24
74LS20	£0.16
74LS21	£0.14
74LS22	£0.14
74LS221	£0.36
74LS240	£0.32
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74LS245	£0.33
74LS247	£0.32
74LS251	£0.24
74LS257	£0.24
74LS258	£0.24
74LS26	£0.14
74LS266	£0.14
74LS27	£0.14
74LS273	£0.32
74LS279	£0.25
74LS30	£0.14
74LS32	£0.14
74LS365	£0.21
74LS367	£0.21
74LS368	£0.21
74LS37	£0.14
74LS373	£0.32
74LS374	£0.32
74LS375	£0.34
74LS377	£0.32
74LS378	£0.62
74LS38	£0.14
74LS390	£0.25
74LS393	£0.24
74LS395	£0.26
74LS399	£0.62
74LS40	£0.14
74LS42	£0.25
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4013	£0.17
4014	£0.30
4015	£0.31
4016	£0.18
4017	£0.25
4018	£0.27
4019	£0.19
4020	£0.31
4021	£0.31
4022	£0.32
4023	£0.16
4024	£0.21
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4026	£0.40
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4053	£0.25
4054	£0.56
4055	£0.30
4060	£0.31
4063	£0.29
4066	£0.18
4067	£1.91
4068	£0.16
4069	£0.20
4070	£0.17
4071	£0.20
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4073	£0.17
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4077	£0.17
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4085	£0.28
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Digital Television

A view of the future by James Archer

3

We have seen how digital techniques effectively allow a television picture to be broken into individual picture elements, pixels, and the information sent along the microwave links or cables to the transmitters is a sort of complex morse coded message which effectively says 'picture element number x has brightness level y and colour z'. At the far end of the link the message can be decoded, and a completely new picture can be reconstructed, a pixel at a time, by rebuilding each pixel according to the brightness and colour levels specified in the coded message. The received picture should therefore be a perfect recreation of the original.

Surely the next step should be to leave the picture in digital form, to transmit the digital bit stream over the air , and to receive it in your home, where a perfect picture could be recreated in the digital receiver? There would be no need for analogue to digital conversion before transmission or for digital to analogue conversion right until the display stages of the receiver, so any unavoidable degradations could be kept to a minimum.

Now all this is very fine in theory, but the actual amounts of information required to convey a moving television picture are absolutely huge, and it is this high bit rate requirement, and the corresponding wide bandwidth requirement, that proves to be the problem.

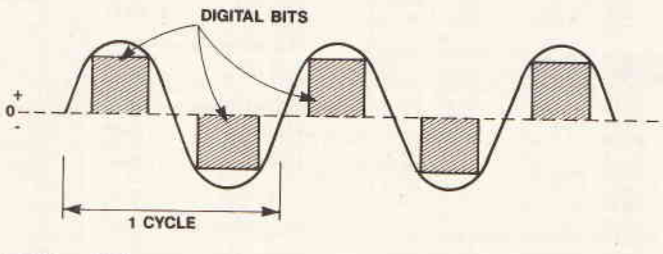


Fig.1 Simplistic representation of how a sine wave might carry two bits per cycle of PCM

As we saw in the first part of this series, CCIR Rec.601, the world standard for digital television studios, says that to turn a standard studio signal into digital form, each of the three component signals Y, BY, and R-Y must be sampled at the following rates:

Y (luminance) at 13.5MHz
BY (colour difference) at 6.75MHz
R-Y (colour difference) at 6.75MHz

Thus we must take a total of 13.5 + 6.75 + 6.75
27 Million samples per second.
Since eight bits per sample are required to provide the necessary 256 grey levels, we thus need:

8 x 27 Million bits per second
= 216Mbit/sec.

Using simple pulse-code modulation we can carry no more than about two bits per cycle of the radio frequency carrier, so we would need a bandwidth of around 108MHz (i.e. 216/2) to convey our digital signal, whereas our initial baseband signal took up only 5.5MHz.

So there is the main problem: you never get anything for nothing in engineering, and digital signals have to pay a price for their ruggedness - bandwidth. In practical terms a DBS satellite could transmit one digital signal with a bandwidth of 100MHz rather than four analogue ones each with a bandwidth of 27MHz, which is not much of a bargain when you are a broadcaster. There are very real costs associated with this bandwidth, of course - with individual satellite transponder leasing costs of several million pounds per year, you would have to be a very altruistic broadcaster to decide that you will use and pay for four transponders rather than one, just to give your customers a potential improvement in technical quality!

One beauty of digital electronics is that it is possible for the engineers in the research laboratories to make reasonably accurate assumptions as to what will be possible one year, two years, and even five years ahead, and this is important, since it means that we can now see the way ahead in the digital transmission field.

Since our major transmission problem is that we currently need to transmit too many bits, it will come as no great surprise that the solution we are seeking lies in the adoption of techniques called BIT RATE REDUCTION.

Bit Rate Reduction

The last few years have seen great strides in the development of bit-rate reduction techniques for television pictures, these usually being based on the well-known fact that most TV pictures carry a great deal of redundant information, since the picture occuring at any instant is generally very similar to the one that came 1/50th of a second before it. Initially the bit rate reduction work was carried out with the aim of finding practical ways of transmitting digital signals between studio centres along links provided by PTTs (post & telegraph authorities, the national & international carriers), but it now looks as though the work will eventually lead to methods of transmitting signals directly to the home. Most of the transmission circuits around the world comply with standards laid down by the CCITT, an organ of the International Telecommunications Union (ITU), which is the international body responsible for regulating the world's telecommunications systems. CCITT standard digital transmission circuits offer nominal capacities of 140Mbit/s, 68Mbit/s, 45Mbit/s, or 30Mbit/s, so if full studio quality 216Mbit/s digital television are to be sent along these circuits, some form of digital bit rate reduction is a must. In addition to the technical difficulties there are often severe cost disadvantages to using these digital circuits, and the cost of using a 140Mbit/s circuit to carry a single digital video picture signal compares unfavourably

TV

with the cost of sending an analogue video signal. In reality, then, these digital circuits can only be used economically if several digital picture signals can be squeezed down one 140Mbit/s circuit; this requires significant amounts of bit rate reduction.

Many different techniques have been examined in the search for the optimum method of bit rate reduction, the aim being to produce pictures that show only the smallest possible loss of quality from the original, whilst using the lowest possible bit rate. The two main techniques currently in vogue for bit rate reduction are DPCM and DCT. Differential Pulse Code Modulation is a technique in which oniy the difference between a predicted sample value and its actual value is transmitted. The value of the predicted sample can be obtained by intrafield sampling, i.e. looking at the adjacent samples in the same field, or by interfield sampling, where the prediction is based on adjacent sample values from adjacent fields. DPCM has been used for some time now to carry out the relatively simple reduction from 216Mbit/s to the 140Mbit/s required for digital contribution circuits, and intrafield predictions have been shown to give excellent picture quality, making the processed pictures indistinguishable from the originals.

Discrete Cosine Transformation

A second technique, Discrete Cosine Transform coding, takes the data from a complete block of picture elements and uses mathematical processes known as orthogonal transforms, which are similar to the better known Discrete Fourier Transforms, to convert this data into an equivalent form from which any redundant information can be removed before transmission.

With both of these techniques some additional degree of bit rate reduction can be achieved by removing the main horizontal and vertical blanking periods before the main coding is begun. Another useful technique, variable length coding, can be used to supplement both DPCM and DCT. Whereas in normal coding methods a fixed number of bits is allocated to each sample, this method gives the most commonly occurring sample values the shortest code words, the longer words being allocated to less frequent sampled values.

Unfortunately there is no one technique that gives the best results on all pictures; different types of picture respond best to different bit rate reduction methods, and it always seems possible to find some exceptional picture material which can be used to 'crack' even the most sophisticated bit rate reduction system. In spite of this,tests in 1989 and 1990 by a working party of CCIR and CCITT members showed that it is possible to obtain very good picture quality on nearly all types of material at the lowest bit rates currently used by broadcasters, 34Mbit/s in Europe and about 45Mbit/s for the USA. The work in the various research laboratories around the world has now crystallised in a general agreement that the way forward is to use a complex technique called motion-compensated hybrid Discrete Cosine Transformation (DCT) of the original TV picture signal information. This is not to say, that other techniques could not also be used, and we shall also take a quick look at the exciting possibility of using fractal mathematics and vector quantisation processing as alternative bit rate reduction techniques.

The last few years have seen significant strides in the development of rugged and practical methods for redundancy reduction coding of television pictures, and as mentioned earlier, most research projects in this field have converged on the technique known as motion-compensated hybrid discrete cosine transform (DCT) coding. Successive television images nearly always contain a high degree of correlation, and by cutting out the redundant or repeated information from one image to the next, only essential data need be sent. Because some parts of pictures contain less redundancy than others, the result is a variable data rate which must be smoothed by a buffer store. As the buffer store is limited in size, pictures which have little redundancy for the coder to exploit could be subjected to some distortion, and while typical pictures will easily be accommodated, the chosen processing system must ensure that any unwanted effects on critical programme material are kept to a minimum, and any such effects must be insignificant if the quality of the broadcast service is to be maintained. This variable demand for data capacity in the buffer circuitry is normally considered to be something of a disadvantage, but any surplus capacity could be put to good use for carrying additional data such as teletext or over-air addressing for conditional access purposes.

When an analogue signal is sampled at regular time intervals, using the quantising process described in the first part of this series, the result is a signal made up of a series of these analogue samples, sometimes known as a discrete-time signal, and more generally called a digital signal. For digital signals we usually understand that both the time and the amplitude of the signal are quantised.

Any such signal can be examined in at least two ways, in the time domain, which has been the traditional approach, or in the frequency domain, depending upon the kind of analysis to which we need to subject the signal. As an example, filter design can be carried out in either the time or the frequency domain. The introduction of modern high speed computers in the form of digital signal processing chips has given engineers the capability of mapping signals into the frequency domain reasonably easily, and the mathematical processes used to carry this out are known as Discrete Fourier Transforms (DFT). A fast Fourier Transform (FFT) is merely a computer algorithm (program) which enables the DFT to be calculated hundreds of times faster than using standard techniques, and FFFs now enable fairly modest computers to carry out the calculations in real time. A particular type of DFT found to be especially suited to image processing is the Discrete Cosine Transform.

The essence of the Discrete Cosine Transform technique when it used for bit rate reduction of video images is that the differences between two successive video frames are calculated; since it would involve too much data to do this for every individual picture element, or pixel, blocks of eight by eight pixels are generally used. The DCT mathematically transforms each 8x8 block of pixels into blocks of 8x8 numerical coefficients, and the statistical properties of these coefficients are different from those of the parts of the television image from which they were derived. Many of the coefficients turn out to be effectively zero, since no change has occurred between frames, and these coefficients do not therefore need to be transmitted. The system is active in nature, adaptively choosing the most suitable processing mode for the picture with which it is dealing. Sometimes it will decide to process the coefficients representing the actual blocks of pixels with which it is dealing at a particular instant, whereas at other times it will take into account and process the information which it has stored from the equivalent blocks of pixels from

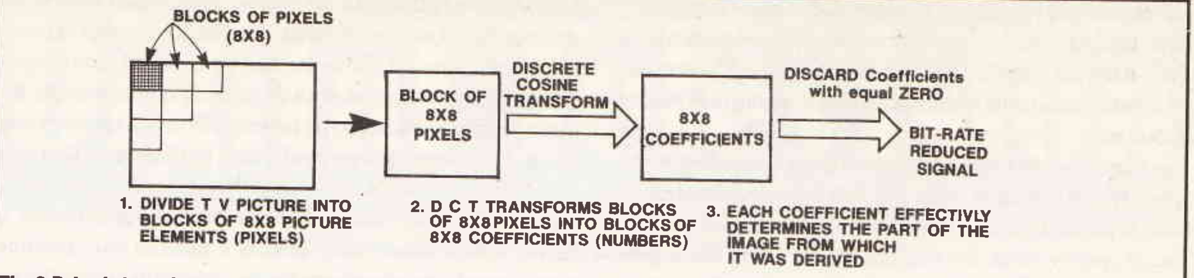


Fig.2 Principles of DCT image coding for bit rate reduction

previous television fields.

The current state of the art allows perfect decoding of even the most critical picture material with a bit-rate of about 30Mbit/s, a big reduction from the original 216Mbit/s, and by accepting that it would be practicable in the domestic TV market to accept some reduction in vertical chrominance resolution, and by using improved motion compensation techniques, engineers are confident that high quality pictures can be transmitted and received at a bit rate of as little as 12Mbit/s. This was demonstrated at the IBC exhibition in Brighton in September 1990.

Alternative digital compression methods.

Vector Quantisation (VQ).

Unlike the DCT based systems, Vector Quantisation does not rely on coding differences between successive frames of the picture, and its exponents claim that it is therefore much better at dealing with the fast moving parts of pictures, which can give rise to problems, known as motion artefacts, on other systems. The intraframe compression technique used by the VQ system is said to require only low-cost read-only memory (ROM) rather than the higher cost dynamic random-access memory (DRAM) used by other systems. To make use of the VQ technique a wide range of different pictures, representative of those to be transmitted, is processed, and the result is a large number of 'image vectors', which represent blocks of pixels, rather like those produced by the DCT process. These vectors are then compared with a set of vect has been stored in a standard vector codebook, looking for the stored vector which gives the closest match to the original. Compression of the data before transmission is then achieved by transmitting only the address of the particular vector, rather than the actual image data.

The receiver then takes the decoded addresses, and uses a carefully selected set of the corresponding image vectors to rebuild the original images, with only a limited amount of degradation.

Scientific Atlanta has demonstrated NTSC quality pictures, compressed using VQ techniques, at a bit rate of only 3.9Mbit/s, and there is still some considerable doubt as to whether the VQ system could be extended to deal with HDTV pictures. For VQ to work satisfactorily, very large code books are needed, and the complexities of searching these for the wanted vectors have only recently made it realistic to use such techniques.

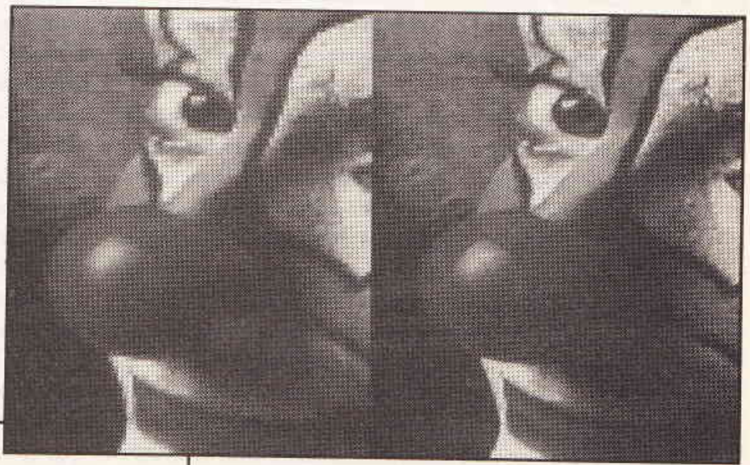


Photo courtesy of Iterated Systems Ltd

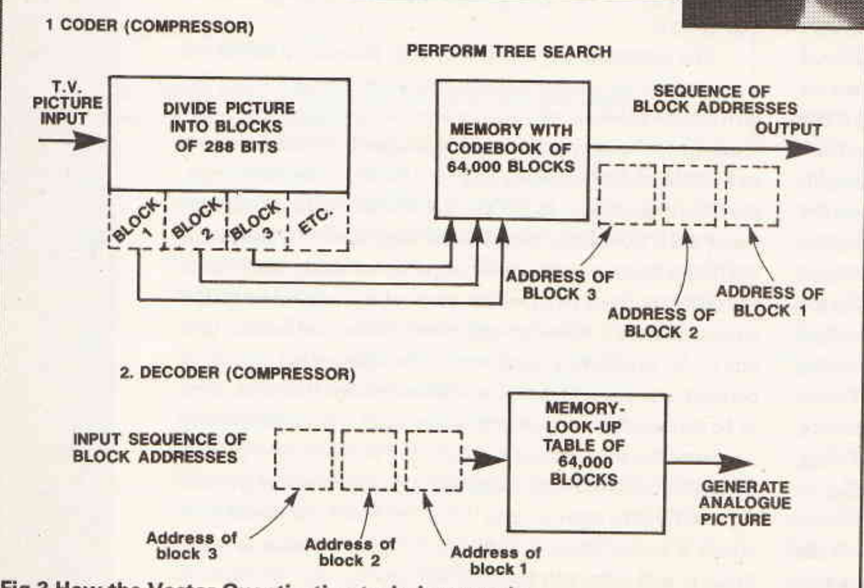


Fig.3 How the Vector Quantisation technique works.

Fractal Transform Image Compression

Microsoft, best known for its work in the personal computer field,has recently announced that it intends to licence its Fractal Coding Image Compression technology, claiming that it offers the highest usable image compression on the market. In parallel with this, a company called Iterated Systems Inc., led by a former professor of mathematics at the Institute of Technology in Georgia, is claiming that fractal technology will revolutionise image processing. Whether or not these American claims can yet be justified, there is no doubt that fractal transforms can play a useful part in this work. Fractals are associated in most peoples' minds with clever computer generated displays that seem to take ages to build up, and become more and more complex as they do so. The mathematician Mandelbrot brought his ideas to the attention of the public in the late 1970s, showing patterns that appear equally complex no matter how much they are magnified, and it was shown that this type of pattern could be created from relatively simple computer programmes - I remember generating such images on a BBC microcomputer, and the patterns kept on growing for hour after hour as the iterative programme continued on its way; although the pattern grew larger and appeared more complex, a keen examination of the pattern would show that it was made up of a relatively small number of similar elements, no matter how big it became. Although much interest was generated when it was shown that many naturally occurring but complex features, such as vein patterns and the edges of tree leaves could be shown to follow fractal formations, it is only in very recent times that TV researchers have taken an in-depth look at this area, to see whether there are any techniques used in fractal mathematics that could be applied to television images.

The two features that seem likely to appeal to TV image processing researchers are that fractal images can effectively be magnified or zoomed into 'ad infinitum' without losing their basic structure, and that since such images can be produced from fairly simple computer programmes, they may have only a small essential information content, which could make them suitable for use in compressed form. The generation of fractal images from computer programmes essentially starts from a chosen place on the screen, 'the origin', and then merely tells the screen dot to move and turn in a regular series of movements, and then to repeat the process. After thousands of repeated passes through this instruction set, a pattern builds up on the screen, and it is fascinating to see how just a few bytes of information can build up a complex pattern.

In a similar way to that used for image processing using the Discrete Cosine Transform,it is now possible to mathematically process images using a 'fractal transform', an invention claimed by Dr Barnsley, who now heads Iterated Systems Inc., which has actually registered the 'fractal transform' name. The fractal transform can take an image and decide which fractal elements can be used to make up that image. It is then possible to generate computer algorithms which can create each fractal element, and by running these algorithms the image can be re-created. As in the DCT analysis of an image, the fractal analysis first divides the image into small blocks, groups of pixels. Having done this, the fractal process begins to search for similarities in the image; what is being sought are similar blocks elsewhere which make up the image - they may not be exactly the same as the block with which they are being compared, but they will have the same basic information, perhaps moved in position and twisted angularly from the original. The process is an iterative one, so once some similar blocks have been found another set will be sought for that are similar to the second set, and so on.

Fractal images can currently be compressed by as much as 77:1, and the mathematical processes are still being refined. The compression process, whereby fractals are generated from moving images, is currently very computer intensive, requiring much time and fast computer hardware, but the decompression algorithms are very much simpler. This should make it possible to regenerate the compressed pictures with simple equipment, and a plug-in PC card is currentlyavailable.

Although this new technique promises much in the television field, once again I have to caution against over-optimism in the short term. A typical computer VGA image currently now takes a few seconds to decompress, and video frames, which can be compressed to take up only a few kilobytes, take around a second. There is therefore quite a long way to go before this technique can be used for real time moving TV pictures at 50 fields per second. Nevertheless, it is possible to see where these techniques could lead, and to predict that advances in computer processing power will have an enormous impact on the application of fractals to TV image processing.

Fractal techniques need not be used by themselves, and the image compression that they can provide could readily be combined with techniques such as differential coding and motion compensation, which have been developed for use with other systems. It does seem that for any particular quality of image, fractal transforms will continue to be able to give greater compression ratios than DCT, the problem being that in order to increase the compression available from DCT it is necessary to split the image into bigger and bigger blocks of pixels, which means that the resolution of the image must inevitably become coarser and coarser. The fractal compressed image theoretically still keeps the same resolution, no matter how great the compression used, since you are effectively reconstructing the picture from a mathematical series of instructions. At the limit, it seems likely that an image which has been subjected to too much fractal compression would start to lose some detail. Fractal techniques, as newcomers to the television field, could have much to offer, and I shall be keeping a close watch on their development over the next year or two.

Transmitting the bit-rate reduced pictures

Now that we are able to see a possible way forward for the generation of bit-rate reduced digital television pictures, ways in which we could possibly transmit these pictures have to be found. In an ideal world the TV broadcasters would simply be allocated a portion of unused radio frequency spectrum for these new services, but in real life the lack of spectrum availability means that this is most unlikely to happen. A most useful aim would be to transmit this digital information, that is, extra TV programme channels, in the existing UHF band, sharing it with the existing analogue broadcast programmes. As can be imagined, this is by no means simple.

Digital Television – Developments in transmission

Although we have so far intimated that there is no real possibility of transmitting television pictures digitally for years to come, because of the high bit rates and wide bandwidths required, IBA UK research laboratories revealed in September 1990 that new techniques that they have developed might bring digital TV forward by several years.

Using the acronym SPECTRE

Special
Purpose
Extra
Channels for
Terrestrial
Radio-communication
Enhancements

A project is currently being undertaken in order to try to see whether it might be possible to further develop the already extensive use which is made of the UHF spectrum for carrying television signals. The present situation is that the 44.8MHz wide channels that make up the current UHF TV broadcast band in the UK are actually used and reused by something like 3400 different television transmitters, most, of course, being low power relay stations.

The idea of SPECTRE is to investigate the feasibility of additionally squeezing a large number of low-power digital signals into this same chunk of bandwidth, without affecting the existing analogue signals.

What makes such an idea possible is that although the present UHF band was well planned back in the early 1960s, making various assumptions about receiver technology and transmitter performance capabilities, we can now see, with our more up to date knowledge of digital signal performance, that there is in fact some spare capacity, some redundancy in this band that we could now perhaps make use of.

The first problem to be solved is how to transmit the digital signals without causing interference to the existing analogue PAL pictures.

This can be done, provided that the level of the digital signals is kept extremely low, compared with that of the analogue signals. It is possible to provide digital TV pictures over the same coverage area as the analogue signals even with the digital signals at very low power, because the carrier to noise ratios required for the satisfactory reception of digital signals are very much less than those for the existing analogue signals.

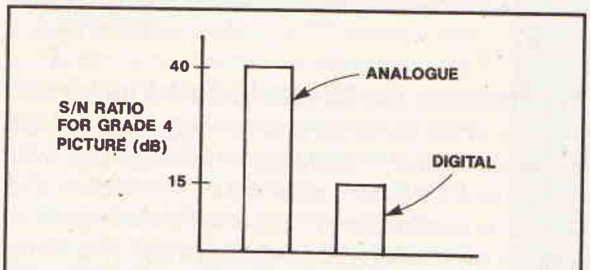


Fig.4 Comparison of Signal to noise ratios required for analogue and digital transmissions

Typical C/N figures for a grade four picture, almost perfect, are about 40dB for our existing analogue PAL service, but only 15dB suffices for a digital service. With modern receiver front ends and the improved noise figures that have become realistic over the past couple of years, it should be possible to obtain matching coverage areas from digital transmitters radiating signals with 30dB less power, that is one-thousandth of the power of the equivalent analogue transmitters. This low power digital operation is of great benefit in reducing potential interference to the existing analogue PAL signals.

As I never tire of reminding people, however, we never get something for nothing in engineering, and one disadvantage of the low power operation is that the digital signal will have to operate in a very hostile, noisy and interference-prone environment. As well as this, there may well be problems with multipath interference, where signals are reflected from nearby buildings or hills and arrive at the receiving aerial at various different times.

Whereas this would cause 'ghost images' on an analogue signal, a receiver that it is trying to pick up a digital transmission will effectively have to sort out the direct pulses from the reflected 'ghost' pulses that arrive at different times, a very difficult task under normal circumstances - how can a receiver distinguish between a 'wanted' pulse and a reflected one? Readers who have tried to receive teletext over difficult signal paths or without a proper aerial will know that the receiver often displays completely wrong characters when the digital signals are received with errors. For digital broadcasting, the choice of modulation method is therefore critical.

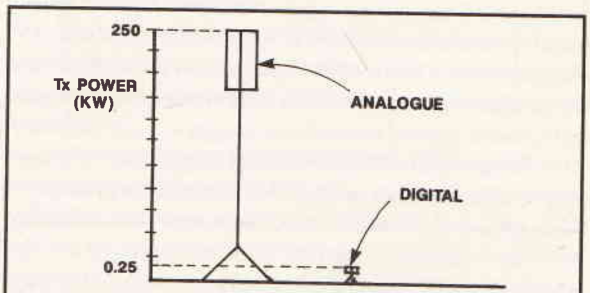


Fig.5 Comparison of transmitter powers required to provide similar coverage from analogue and digital transmissions.

Engineers from National Transcommunications Ltd., who have been at the forefront of this work, believe that the best idea could be to use a technique called Orthogonal Frequency Division Multiplexing modulation, using a large number of overlapping digitally modulated carriers, some form of QPSK (Quadrature Phase Shift Keying) coding then being used. QPSK is a well-tried modulation method in which the phase of the radio frequency carrier is altered between four different states; very low error rates can be achieved, and the decoding circuitry is simple. The rule of thumb for QPSK modulation says that the bandwidth required is approximately equal to the bit rate, so that a 34Mbit/s signal can comfortably be carried on a 34MHz wide satellite transponder.

Orthogonal
Frequency
Division
Multiplexing

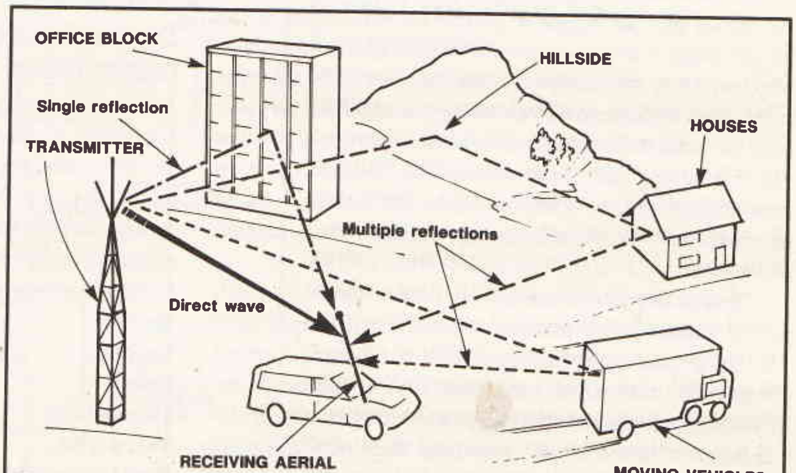


Fig.6 Multipath interference

We are familiar with the idea of FDM, frequency division multiplexing, where several different signals are modulated onto a group of adjacent carrier frequencies which are then sent along a common channel. OFDM takes this a stage further, and the diagram illustrates what happens.

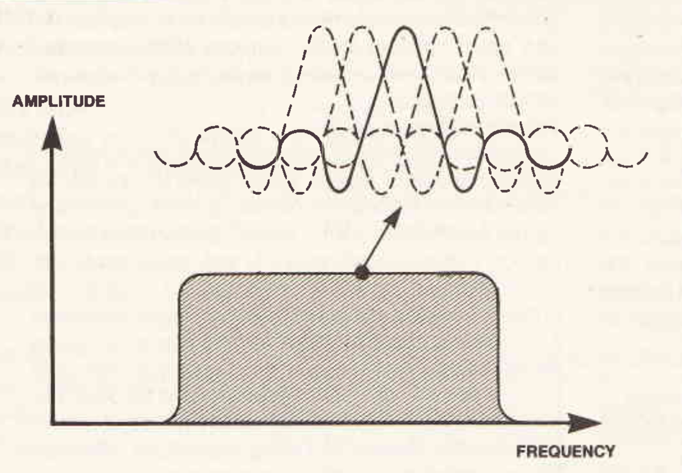


Fig.7: OFDM Waveforms

The OFDM technique breaks the digital data stream into sections and uses these multiple signals to modulate a large number of simple carriers, and it is the total of all these modulated carriers that makes up the OFDM signal. The individual carriers are derived from the Discrete Fourier Transform of the original signal.

The essential feature of the OFDM technique is that the frequency spectrum of each modulated carrier is deliberately allowed to overlap its neighbour, and careful selection of the carrier frequencies, of the phasing of each of the signals, and of the digital coding system that is used, allows the individual carriers to be separated out again by the application of the Fast Fourier Transform as the signal is demodulated. The same circuitry that can distinguish between the individual carriers is able to take account of out of phase reflected signals, such as those that are produced by multipath interference.

The spectrum that is produced by the tightly packed overlapping spectra of each of the individual carriers gives an effect like that shown on the bottom of the diagram, from which we can see that the system makes very efficient use of the available radiofrequency bandwidth.

In practice there is some intersymbol interference caused by the inevitable multipath interference, but the problem is reduced or eliminated by leaving a guard band interval between each symbol; provided that this is long enough to exceed the delay suffered by any of the reflected signals all the orthogonal carriers can be demodulated without difficulty.

The OFDM spectrum, gives excellent results in the presence of multipath interference, and it also turns out to be ideal for use in a hostile interference environment.

The main interference from existing analogue TV transmissions will take the form of two high power carriers at the sound and vision frequencies. OFDM can cope well with this because the large number of carriers used means that it is possible to decide not to transmit any information on some of them, if we wish. For our purposes this means that we can cut out those carriers that fall in the portions of the spectrum near to the analogue vision and sound carriers. Effectively, then, we can define a spectral template for the digital signal, and get our receiver to look only at the information within that template, ignoring the interference that would be caused by the powerful sound and vision carriers.

UHF TV channels in the UK are spaced 8Mhz apart, so that the frequency spectrum of transmissions from a single transmitting station will, to a first approximation, look like four equally spaced pairs of sound and vision carriers. Whilst it is true that there will also be energy in the form of subcarriers present at the colour subcarrier frequency and at the frequency of the NICAM digital sound carrier, the effects of these are reduced by the dispersal effect of their modulating signals, so that they have much less peak power than the sound and vision carriers. The regular spacing of TV channels therefore makes it possible to fit in extra low-level digital signals, between the existing signals, as the diagram shows.

Theory suggests that we could actually provide eight new digital channels for each transmitting station, but in practice transmitters are not sited on an ideal frequency lattice, and there are various other problems that restrict the frequencies that can be used.

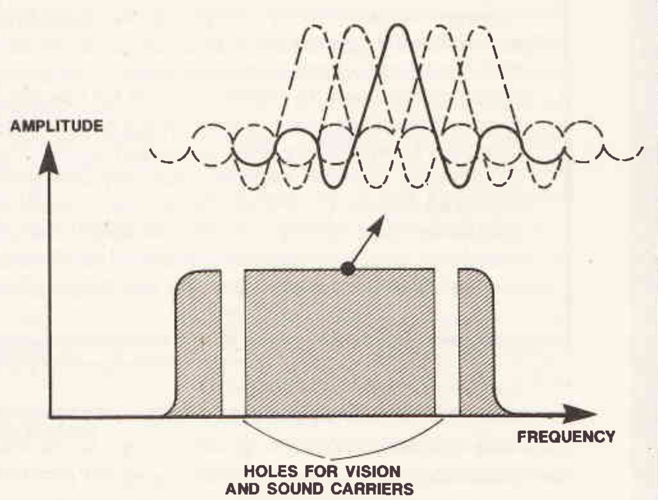


Fig.8 Spectral template to cut out vision and sound carriers

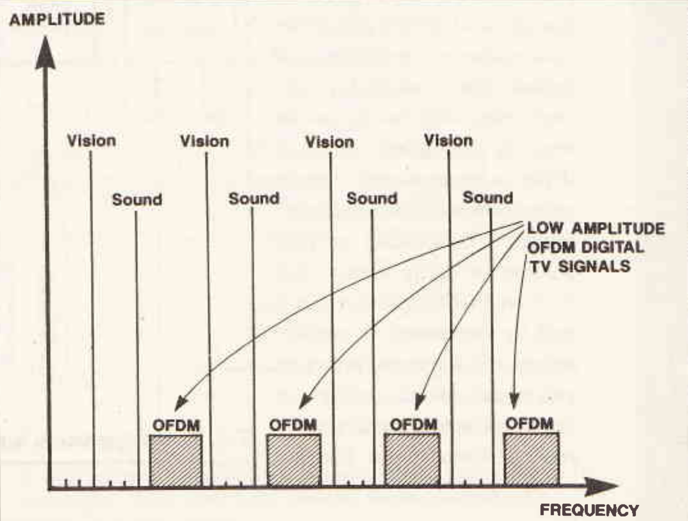


Fig.9 Spectrum with digital signals inserted

'Taboo' channels affecting an analogue TV network

A television transmitter network generally consists of a large number of different transmitting stations situated all around the country, and since the amount of radiofrequency spectrum that is allocated to television broadcasting is invariably restricted, several transmitters in different locations will have to share the same channels. If the transmitters are far enough apart then channel sharing should generally be possible without interference, but in many practical situations transmitters using the same channels will not be as far apart as would be ideal, and there is always the possibility of so-called 'co-channel interference'. Unfortunately, because domestic receivers are generally built down to a price, rather than to the highest possible technical standards, it is found that there are other channels which could cause interference if they were to be used within a certain distance of another transmitter.

Adjacent channel interference

let us assume that the channel to which the receiver is currently tuned in order to view the programme is channel 'n'.

Any nearby transmitter radiating signals on the adjacent channels, i.e. n + 1 and n - 1 would cause interference in the form of patterning or buzz on sound, simply because the selectivity of the ordinary receiver is not good enough for it to separate two adjacent signals.

Local Oscillator Interference

All modern receivers are of the superheterodyne type which use a local oscillator to beat with the incoming signals in order to produce a standard intermediate-frequency signal that can be readily amplified and demodulated. This local oscillator (LO) can be considered as a small transmitter, and in UK television receivers the LO is usually 39.5MHz above the incoming frequency to which the receiver is tuned, channel n. The LO therefore generally radiates potentially interfering signals on channel n + 5 when the receiver is tuned to channel n, since 39.5MHz is close to 40MHz, which represents five 8MHz-wide TV channels. The affected channel obviously depends upon the intermediate frequency used, and is different in different countries.

This means that if a nearby receiver were to be tuned to channel n + 5 it would suffer interference from its neighbouring set tuned to channel n, and in order to prevent this from happening we must treat channel n + 5 as a 'taboo' channel. Similar problems apply in reverse if the n - 5 channels are used, so these too must be regarded as 'taboo'.

Intermodulation products

Certain combinations of input signals can rrbt together to produce spurious signals which generally appear as patterning on the wanted picture. If there are several carrier frequencies, say 'a' and 'b', then various combinations of these, such as (2a-b) can cause problems. Sometimes a similar phenomenon known as cross-modulation can occur, where the modulation from one carrier is effectively transferred to the other, giving the annoying effect of two pictures superimposed, or of the sync pulses of one picture showing up on the other. The planners of the television spectrum usage therefore have to ensure that transmitters using frequencies which are likely to generate these spurious signals are kept well separated.

Intermediate-Frequency Beat Interference

When two television signals from two different transmitters are separated by the intermediate frequency of the receiver the two signals can combine in such a way that they produce a beat signal at the difference between the two frequencies, and this signal can be picked up by the receiver's intermediate frequency amplifier. Such an arrangement of channels must therefore be considered as taboo, providing frequency planners with yet another restriction.

Image interference

Since the front-end selectivity of the average domestic television receiver is poor, signals at twice the intermediate frequency of the receiver from the wanted channel can pass through the receiver. This applies to both sound and vision signals, and in UK practice means that signals on n + 9 must also be considered taboo.

Many of these taboos are really only applicable to the current situation where other analogue transmissions could interfere with other analogue transmissions, and much work remains to be done to see what the practical effects of various combinations of analogue and digital signals interfering with each other will be. Since the level of the digital signals is deliberately kept very low in order to reduce potential interference to the existing services, we can be fairly certain that problems will not arise, but it will be interesting to see the effects of high power analogue vision and sound carriers on low-level digital signals closely packed between them. The first

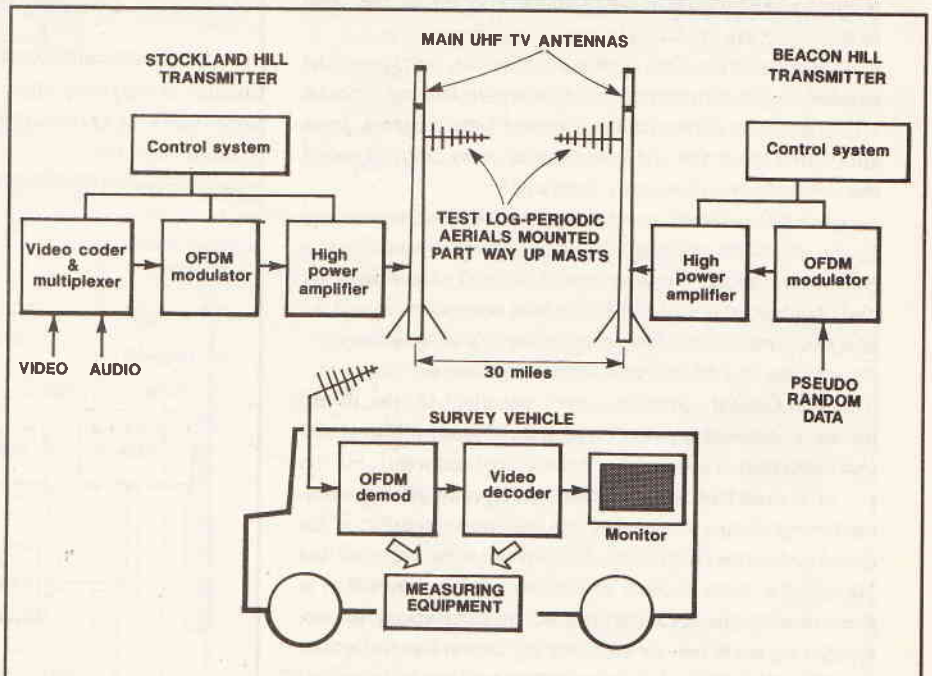


Fig.10.Arrangements for field trials in Devon (ITC/NTL IBC 92 paper)

TV

tests of SCEPTRE took place in the west of England during the spring of this year, but there is a lot more work to do before a usable system is developed.

At the International Broadcasting Convention in Holland at the beginning of July 1992, details were given of the arrangements for field trials of the SPECTRE system that are being carried out using the Stockland Hill and Beacon Hill transmitters in Devon.

The transmitters are about 30 miles apart, and at each site a log periodic aerial has been erected half way up the existing mast; the aerials are directed at each other, making possible a range of different experiments. Normally one transmitter will broadcast a digital television signal using the OFDM modulation system, and the other will radiate an interfering signal.

The arrangement shown in the diagram indicates that Stockland Hill is configured as the provider of the wanted digital signal, and Beacon Hill as the source of potential interference. The Stockland Hill transmission is a compressed digital video signal modulated using OFDM, whilst the interfering OFDM signal from Beacon Hill is modulated by a pseudo-random data sequence. A mobile field strength measuring vehicle can drive around the service areas, checking on received picture quality and measuring both field strengths and error rates. The OFDM transmissions will take place at the same time as the two stations are transmitting their normal four PAL UHF transmissions. The Stockland Hill transmitter is well sited for tests of the effects of SECAM transmissions from France, and of co-channel interference from the main Rowridge UHF TV transmitter on the Isle of Wight.

The equipment has been arranged so that it is possible to transmit signals from both Stockland Hill and Beacon Hill on different 'taboo' channels, and with both OFDM transmitters working on the same frequency it is will be possible to investigate the use of single frequency working, a technique that could have enormous implications for future television networks. If a method could be devised of operating dozens of television transmitters throughout the country on the same frequency channel, without mutual interference, as has already been tried for digital radio broadcasts, then the use of the existing frequency spectrum could be multiplied many times.

All sorts of claims have been made in the technical press about SPECTRE being able to carry as many as forty simulcast channels, but those engineers closest to this work dismiss such talk as nonsense, and taking into account all the various factors that we have discussed above, a more realistic outcome might be to be able to provide four analogue and four digital TV signals from each transmitting station, thus doubling the number of possible terrestrial TV programmes.

Whilst this would be some considerable achievement, I must stress that it will not happen overnight. Much work still has to be done before we can know whether a digital TV service to the home would be feasible and economic, but the rewards would be so great that it certainly seems worth continuing with the research. As the same techniques are developed further it should one day prove possible to use the digital bit streams to carry HDTV signals, or, eventually, to carry tens of extra channels within the same spectrum. Research engineers are already looking forward even further, towards a time when it may be possible to clear the terrestrial spectrum of the existing old-fashioned PAL transmissions, thus freeing all the bandwidth to be used for dozens of different digital HDTV transmissions. This is the basis of some of the rather way out reports and predictions that have appeared in the press, suggesting that digital television will bring all this and more - it may well do so, but the likelihood is that some of the predictions will not be fulfilled for many years to come.

One key question as to whether such ideas can be turned into reality will centre around the ability of receiver manufacturers to provide low-cost digital receivers which can process Fourier transforms. Initial discussions suggest that it will be the late 1990s before such receivers could be available at realistic prices. Although it is not yet possible to buy off the shelf large scale integrated circuits to carry out these processes for television images, new digital signal processor chips are starting to appear in the manufacturers catalogues at reasonable prices, and one can easily extrapolate from the performance of today's chips to a time just a few years hence when high quality TV pictures could be processed. An example of such technologies is the PDSP16510 stand-alone FFT processor chip that is currently being made by GEC Plessey Semiconductors in the UK. The PDSP16510 performs forwardor inverse Fast Fourier Transforms on data sets containing up to 1024 points. The input data and the numerical coefficients are each represented by 16 bits. The 1024 point complex transform can currently be completed in 96us, which means that it can be configured to carry out continuous transforms in real time. Now although the processing powers of the current chip are still nowhere near good enough for our projected TV use, even though it is possible to improve system performance by connecting chips in parallel, it does not take much imagination to see that future developments of this sort of digital signal processing device could lead to full function TV image processors becoming available in the not too distant future.

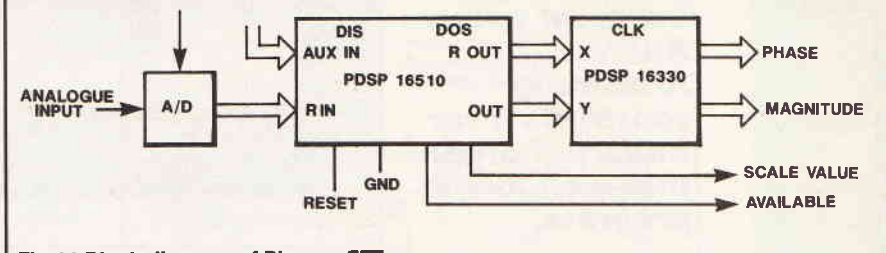


Fig.11 Block diagram of Plessey FFT processor

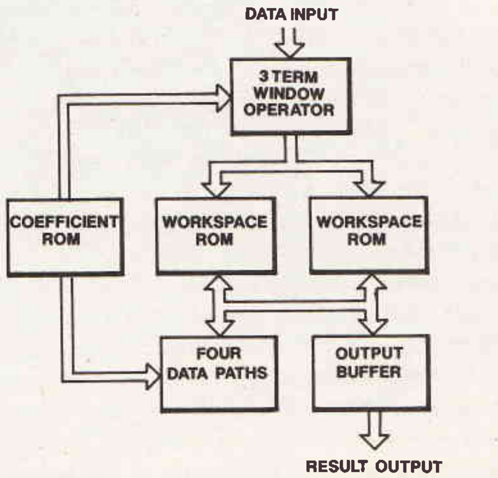
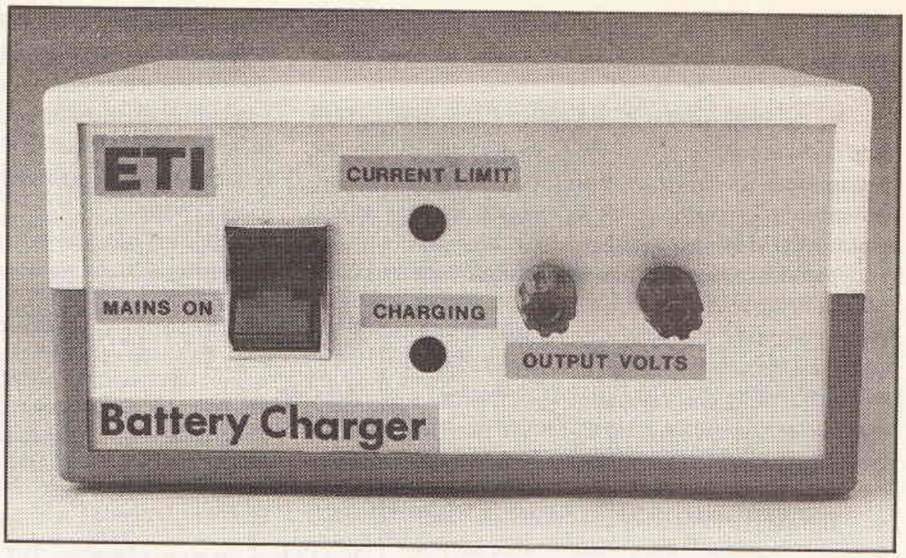


Fig. 12 Typical configuration for perfornnng continuous transforms. (Courtesy GEC Plessey Semiconductors) Digital TV transmission

ETI Cover PCB Project

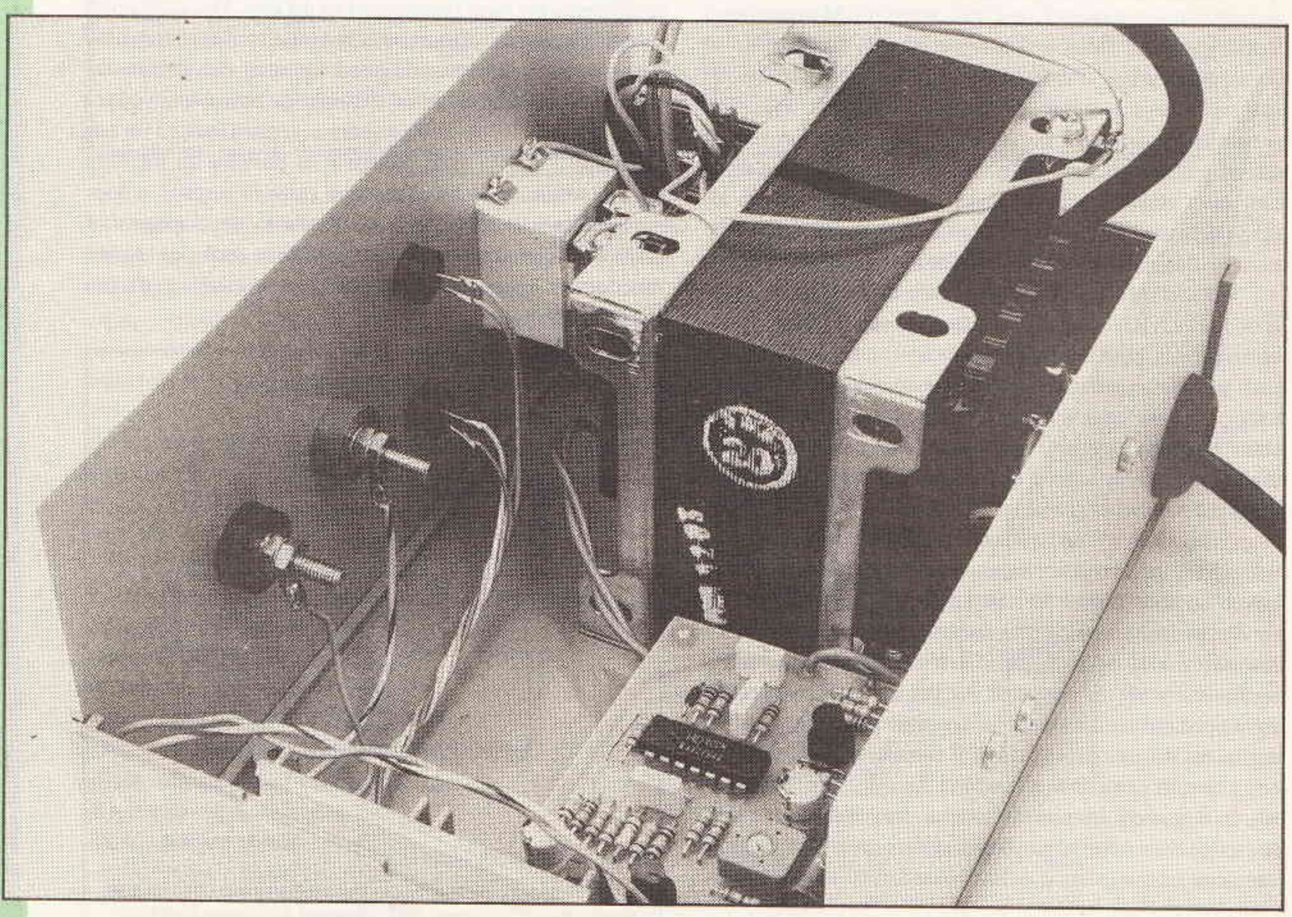
Using the PCB given away on our front cover, Andrew Armstrong constructs a car battery charger that monitors its progress.



Smart Charger

Now that Summer is here, it is a good time to consider the problem of flat batteries in motor cars – before it catches you out next winter. This battery charger project was inspired by a problem my neighbour had. The battery on her car was so flat that the courtesy light only glowed dimly, and when she tried to start the engine, the solenoid made a loud clatter, which was what alerted them to the problem. What was happening was that the battery, though at only about 9V, had enough power to pull in the solenoid, but as soon as the solenoid made contact the starter-motor pulled the battery voltage down almost to zero, causing the solenoid to drop out again. Apparently a fault in the car's security alarm had quietly (for a wonder) flattened the battery overnight.

An attempt to charge the battery using an ordinary battery charger resulted in a blown fuse as the car battery attempted to draw an excessive current. I solved the problem at the time using the gross technological overkill of a 12V 15A short-circuit proof power supply which I built to run my amateur radio equipment. It occurred to me



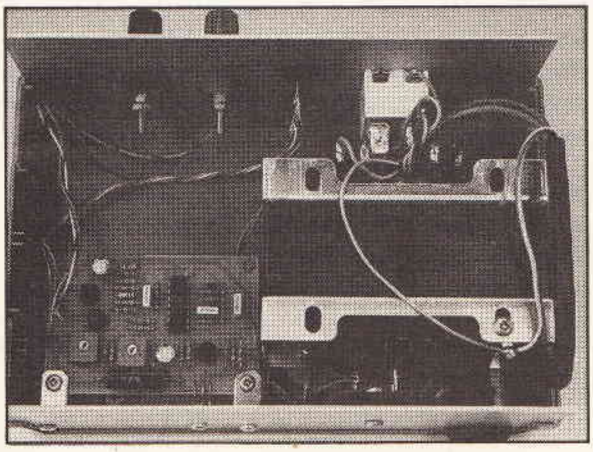
PROJECT

that a thyristor-controlled car battery charger might be a more appropriate solution.

Control method

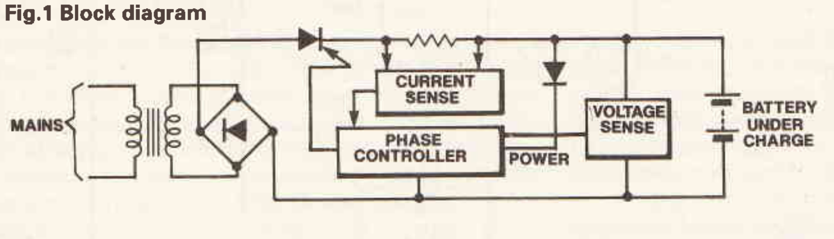
There are two obvious approaches to controlling charging current by means of a thyristor: one is to phase control the rectified unsmoothed output from the transformer; the other is to use a burst-fire technique to regulate the average current. The burst-fire technique can result in higher peak currents, and, unless it is very cleverly controlled, can produce a net DC in the secondary winding of the transformer. It does however typically generate less radiated interference. Despite the question of interference, the decision was made to use phase control which seemed likely to work better in every other respect.

Since the charger was to be electronically controlled anyway, it should limit voltage as well as current, so that it could be safely left connected to a battery without the risk of overcharging it. This might be valuable if the family car is to be left in the garage unused for several weeks, but needs to be started without trouble at the end of that time. It also seemed sensible to use LEDs to indicate the charging status rather than to use an ammeter (which seems to get broken very easily in a garage environment).



The block diagram shown in Figure. 1 illustrates the general principles by which current and voltage measurements are used to control the charging current. Current is sensed across a low value resistor, while the terminal voltage of the charger is also measured. The average charging current is not allowed to exceed a pre-set level (normally this would be set to 4A) no matter how low the battery voltage, but the charging current will decline towards zero as the battery approaches the preset charging voltage, which would normally be set to 14V.

Fig.1 Block diagram



Power to run the phase control and firing circuit is drawn from the battery. One effect of this is that, if the battery is connected the wrong way round, the thyristor will not be fired and no current will flow.

Construction

This unit is intended to be constructed on the free PCB supplied with this magazine. Assembly of the PCB should present no problems, and none of the components are especially static-sensitive. Depending on how the PCB is to be mounted, any or none of the thyristor and the two indicator LEDs may be mounted on the board. On the prototype unit, the thyristor was board-mounted, while the LEDs were connected with wires and mounted on the front panel, as shown in the case plan, Figure 2. Wherever the board is mounted, the thyristor must be on a heatsink or a suitable piece of metal which will act as a heatsink. The prototype unit was built into a clamshell-type plastic case, with aluminium front and rear panels, so the rear panel forms a suitable heatsink in this unit. The PCB is mounted onto the rear panel, using small steel right-angled brackets, and M3 bolts. The triac is also mounted on the rear panel using an insulating kit, so that the panel is not connected to any part of the electrical circuit. The right angled brackets should be bolted onto the back panel before the triac is bolted on, in order to minimise stress on the soldered joints.

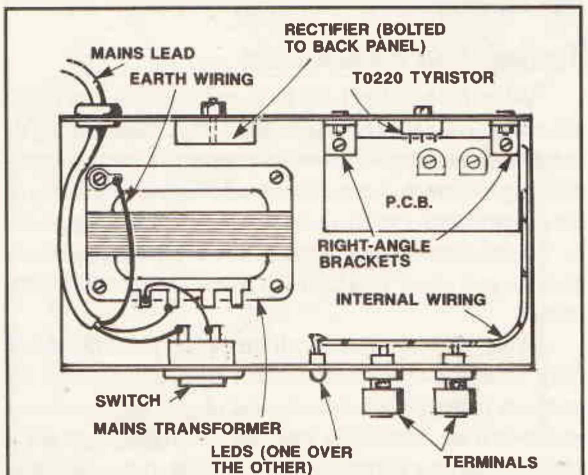


Fig.2 Plan of case

It is also necessary to mount the bridge rectifier on a heatsink, because when 4A is flowing it will dissipate approximately 6W. It is recommended that heatsink compound is used to assist heat transfer from the body of the bridge rectifier to the metal.

The mains transformer was bolted to the plastic base of the case, using M4 bolts. It had to be positioned very carefully, to maintain a safe spacing from the switch terminals while allowing room for the bridge rectifier behind it. The switch was mounted at such a level that its terminals were below the level of the transformer terminals. Mains wiring enters at a retaining grommet on the rear panel and goes straight to the switch. The switch is then connected to the transformer with short wires.

In line with what is currently believed to be safe practice, the metal panels at front and rear of the case were connected to mains earth: the front one via the metal switch bezel, and the rear one via one of the mounting bolts, retaining the PCB. To prevent strain on the earth-wiring, it was star-connected to a tag secured to one of the unused upper mounting holes of the transformer.

Because the current is limited in this design of charger, and because it is protected against reverse battery connection, an easily replaceable output fuse is unnecessary. Nevertheless, it is probably advisable to fit an output fuse internal to the unit to protect against catastrophic failures; for example, the thyristor failing short-circuit. A 6A 1.25-inch fuse was used in the prototype, though this had not been fitted at the time it was photographed. (The prototype was made in accordance with the most advanced modern manufacturing control theory: Just In Time.)

Testing And Adjustment

When the unit has been built and inspected for obvious errors, connect it to a car battery via a 100R resistor to check that nothing is seriously wrong; if something is drawing too much current in the unit, the resistor will get hot or even burn out. This is a fail-safe. If it burns out, look for solder-blobs on the back of the PCB causing short circuits, and check that the IC is inserted the correct way round.

When all is well, set both the preset potentiometers fully anticlockwise, then connect up the unit with an ammeter instead of the resistor and plug it into the mains and switch on. Gradually increase the voltage pot RV2 until the charging current stops increasing. If the current is noticeably less than 4A, increase the voltage setting further and adjust the current set potentiometer RV1 until 4A is flowing. Then connect a voltmeter to the battery, and adjust RV2 so that the current is approximately 0.5A when the battery voltage is 13.8V. It may be necessary to wait a while until the battery charges to this voltage in order to complete the adjustment of the unit.

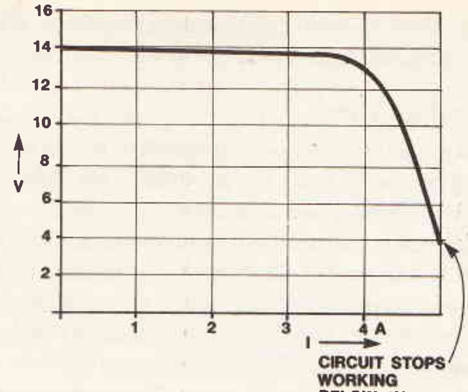


Fig.3 Charging characteristics

The LED indicators show the status of the battery charging process. LED 2 illuminates when the current limit is coming into effect. The current limit is progressive rather than abrupt, giving rise to the curved current/voltage characteristic shown in Figure 3. When this LED glows brightly, it is a reasonable deduction that the battery is deeply discharged.

LED 1 brightens in proportion to the phase angle for which the triac is triggered. When the current limit, LED2, is on, LED1 will be at its brightest. As the battery charges up past the point at which current limiting occurs, LED1 will decrease in brightness. When the charging current has

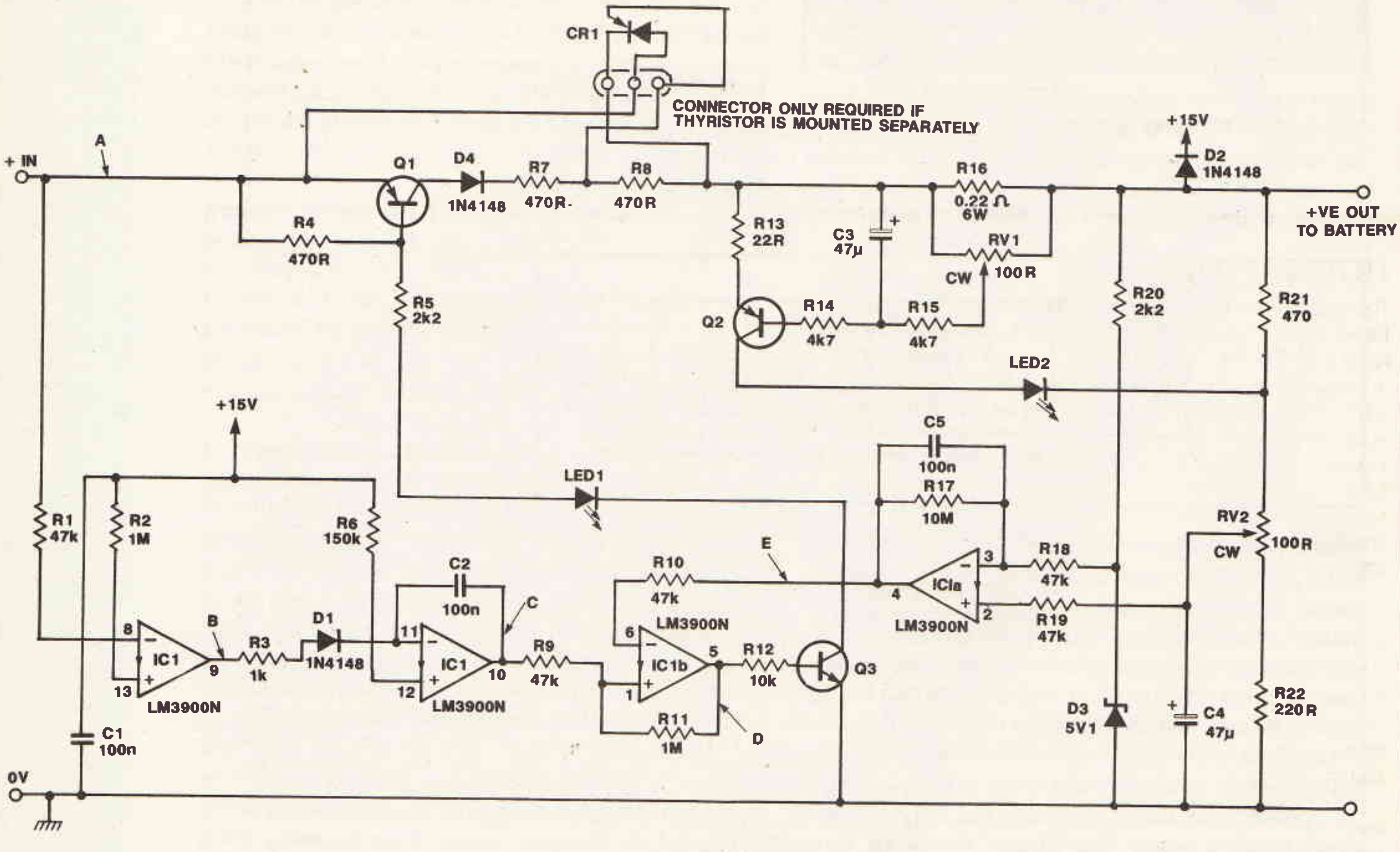


Fig.4 Circuit diagram of Smart Charger

reduced to a very low level indicating that the battery is properly charged, LED1 will be quite dim. It will remains bright enough to see, however, because a significant phase angle of triggering is required before any current at all flows. If triggering occurs after the point on the output sine wave at which the transformer voltage falls below the battery voltage, no current can flow.

The current limit of this charger can be connected to even the flattest battery with no risk of damage, while the voltage limit makes it similarly safe to leave connected for weeks if necessary. The only drawback, if such it is, is that if the battery voltage is below about 4V, the circuit may not work, in which case no charging current will flow. On the other hand, car batteries will very very rarely become this flat, and a battery which is this flat may well be past salvage in any case.

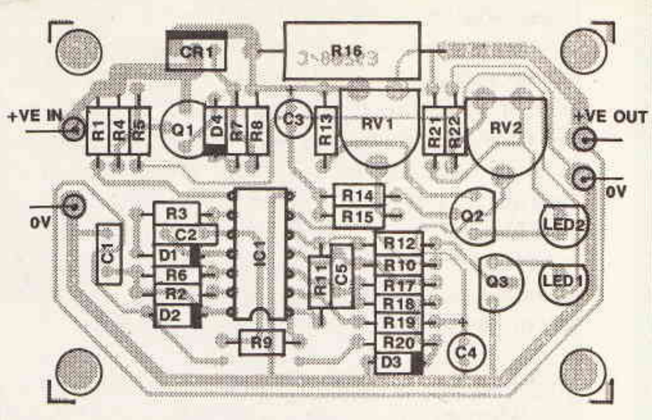


Fig.6 Component Overlay

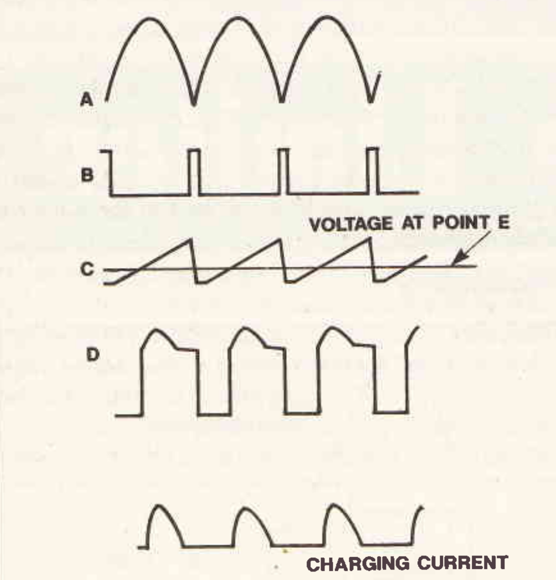


Fig.5 Waveform diagrams

PARTS LIST

RESISTORS

R1,9,10,18,19	47k
R2,11	1M
R3	1k
R4,7,8,21	470R
R5,20	2k2
R6	150k
R12	10k
R13	22R
R14,15	4k7
R16	0R22 6W (Welwyn W22 or similar)
R17	10M
R22	220R
RV1,2	100R

CAPACITORS

C1,2,5	100n
C3,4	47μ/16V radial electro 0.1" pin spacing

SEMICONDUCTORS

D1,2,4	1N4148
D3	5V1
Q1	BC182
Q2,3	BC212
IC1	LM3900N
T1	Thyristor 100V 10A or better eg C126D
LED1,2	red LEDs

HOW IT WORKS

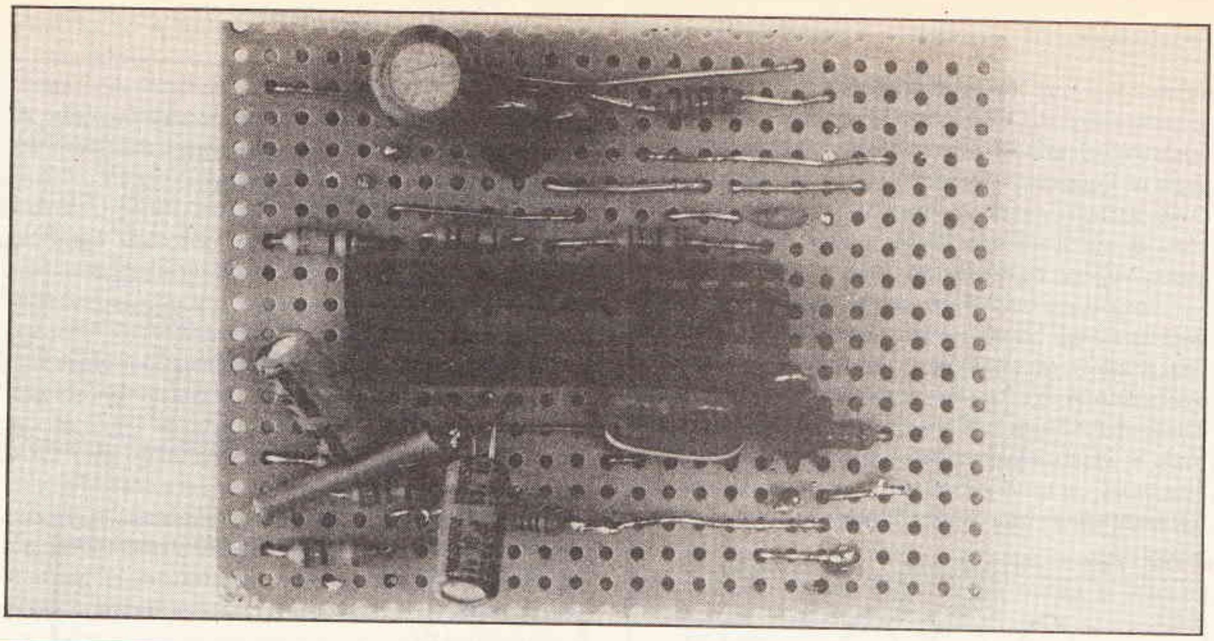
The rectified by unsmoothed output from the mains transformer is fed to the input of IC1c, which is a Norton op-amp. A Norton op-amp is current rather than voltage operated. The output voltage of the opamp is equal to the gain multiplied by the difference in currents between the two inputs. As long as positive input current is flowing, the inputs remain at one diode-drop above the negative supply voltage. IC1c therefore forms a comparator, arranged so that when the input voltage to R1 is high enough to make approximately 12A flow in R1, the output of IC1c switches from high to low. This occurs when the input voltage to R1 is approximately 1.2V (remembering the one diode-drop voltage on the op-amp input). This is illustrated in the waveform diagram, Figure 5.

IC1d forms a ramp generator which is reset by the pulses from IC1c. This part of the circuit generates a linear ramp by using the op-amp to maintain a constant charging current to C2. The op-amp output voltage rises at such a rate that the input current to pin 11 is the same as the constant current fed into pin 12. The reset pulse is of such a high current that the output voltage falls very rapidly, but the capacitor does not charge up in the reverse direction, because it is clamped by the diode characteristic of the op-amp.

The battery voltage is potted down by R21, R22 and RV2, and compared with a reference voltage on a zener diode. IC1a forms an error-amplifier with a DC gain set by the ratio of R17 to R18. The frequency response is limited by C5 in order to keep noise out of the system. The resulting error voltage is compared with the ramp in IC1b to generate pulses which trigger the thyristor via Q3 and Q1. Because the battery voltage as measured at the terminals of the charger will fluctuate substantially as a result of the charging current, the battery voltage measurement is averaged by C4.

The trigger pulses have a peak superimposed on what would be a squarewave, because the battery voltage (which supplies the electronics) rises during the charging pulse. Note that the peak on the top of the trigger waveform ceases before the end of the high part of the waveform – because the charging current ceases to flow when the mains sinewave is close to its zero crossing.

The average current is measured by the circuit including Q2. Again, the measurement is averaged by an RC time constant, because the charging current is pulsating, and it is the average current which must be controlled. As the current limit cuts in, current is fed via LED2 into the battery voltage comparison chain, reducing the error voltage proportionally to the amount of overcurrent. This reduces the firing angle, so regulating the current. The gain of the current limiting control loop has been limited to avoid the possibility of oscillation. The resistance of the voltage measurement chain was deliberately chosen to be low enough that LED2 would light visibly when current limit was in operation.



Dynamic Noise Limiter

A Hi-fi noise reducing system by Jeff Macaulay

Although recently, signal sources have become progressively less noisy it's still an unfortunate fact of life that the problem still raises it's head. In particular radio suffers from noise, exacerbated when listening in stereo. Records also suffer from low level noise, and despite Dolby systems so do tapes. As a hi-fi nut of the first order I find this particularly exasperating and I long felt the need for a simple in-line noise eliminator that could be used when required and bypassed when not.

Most of the noise reduction systems rely on encoding the material first and producing noise reduction on decoding. The two major noise limiters work on this principle. Dolby using treble boost and cut whilst the DBX system uses a compander. Of the two systems the DBX is more complex but is able to do more in the way of noise reduction.

To explain the compander principle one has to consider the dynamic range of an audio signal. The dynamic range is essentially the difference between the loudest and softest notes in a performance. Live music is capable of a dynamic range exceeding 10,000:1, or 80dB. It follows that in order to completely cope with a musical performance any recording medium has to capable of a similar range.

The best records can just manage 1000:1 range whilst a good studio quality reel to reel recorder can manage a little better. The only widely available medium that can pack the range is the CD with better than 90dB signal/noise ratio. It's noise that ultimately limits the available range.

If the softest passages are likely to be lost in the noise level some method must be found to prevent this. The result is that most recorded music is dynamically limited, usually the peaks are compressed.

As there are no standards laid down for limiting it's a pretty futile excercise trying to restore the original range of the program electronically. Heres where a compander comes in. The word is an amalgam of compresssor expander. If the compression is carried out in a certain manner and the expansion in exact reverse the dynamic range of the signal can be halved. We now only require 40dB dynamic range to encode the signal, a feat that can be accomplished by a cheap cassette recorder! On playback the noise level is expanded downward to become inaudible whilst the full range of the signal is presented intact.

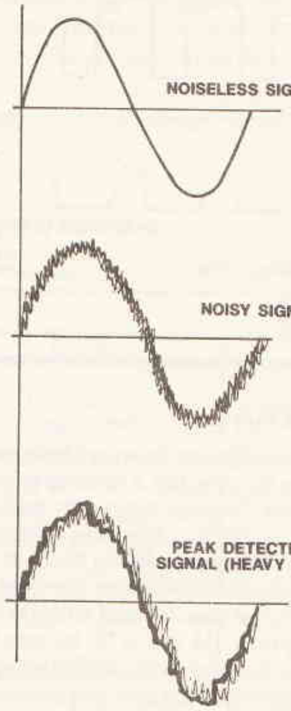


Fig.1 The noiseless, noisy and peak detected signal

In the early eighties I had the opportunity to use such a system. It gave CD quality from specially recorded records with a background of complete silence. However back here in the real world no ammount of fidlling with signals dynamics

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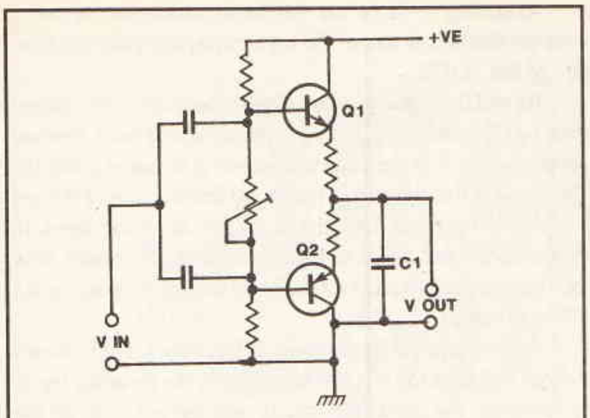


Fig.2 Basic Push Pull peak detector

will help matters with an already noisy signal. However there are ways to alleviate noise problems and these should be investigated before using noise reduction.

Noise comes from two major sources. One is the electrical environment of the equipment and the second due to components within the equipment itself. A major source of noise in an FM system is due to low signal levels. This can usually be traced to an insuficient aerial or one incorrectly positioned. However an aerial can only be pointed in one direction at a time and is usually fixed.

With vinyl records there is usually noise produced from tracing the groove as well as the natural noise level generated by the resistive part of the cartridge impedance. Unfortunately no two records seem to be recorded at the same level and so some are noisy from the first playing.

The noise problems of tape recorders, especially cassette players are to well known to require further elucidation. This brings us back to the original problem, designing a device to reduce the noise to acceptable levels without destroying the quality of the recording. At this stage I ought to issue a disclaimer. As far as I am aware there has yet to be a noise reduction system that doesn't affect the quality of the input signal. To remove noise also implies that some signal will also be adversely affected.

Nevertheless a very worthwhile subjective improvement can be had by fairly simple means. There are two main ways of reducing pre-existing noise. The first is to use filtering of the high frequency content of the signal. This works because subjectively mid to high frequency noise is the most noticiable. The filtering is most efficient, if you like your music sounding as if it were recorded through a thick pillowcase! A better refinement of the system is to make the filtering signal dependent. Here high frequencies are detected and used to alter the filter's turnover frequency.

A well designed system of this kind works well enough. Luckily softly played instruments generate few hamonics so the full advantage of filtering can be obtained where noise problems are most prominent.

On loud passages when the filter bandwidth is widened the signal effectively masks the noise. However most have subjective problems due to the audible opening and closing of the filter's bandwidth.

The other alternative is to use a noise gate. Here the signal is muted until it exceeds a certain amplitude. The gate allows high level signals through whilst supressing low levels. As you can imagine such on-off action is not very hi-fi and so it's rarely encountered in practise.

Figure 1 shows the difference between a clean and noisy signal. As you will notice the noise 'rides' on the back of the wanted signal and is usually of much lower amplitude. If you were to draw a line through the centre of the trace you would have a good approximation to a noiseless signal. My first attempts to produce such a circuit were fruitless and it took many months before a solution was forthcoming. Figure lb shows it.

What I've shown here is a peak detector circuit. Whilst the signal is positive going the sucessive positive peaks are held by the circuit which traces the average waveform. On negative going excursions a negative peak detector provides a complementary output. Sticking both together provides a good copy of the original. In fact the noise can be considered as a form of uncertainty which the circuit reduces. Having worked out how it might be done another few months went by before I could shout Eureka! This occured when I realised that a complementary pair of transistors and a capacitor could provide the results I was looking for.

Figure 2 shows the circuit in it's simplest form. On positive going signals Q1 charges up C1 to the positive peak level. The transistor acts as an emitter follower. When the signal goes negative Q2 conducts charging C1 to the negative peak value. By adjusting the bias voltage between Q1 and Q2's bases the operating level can be adjusted between letting 100% of the signal through to just the peaks.

With a pot to adjust the bias the amount of signal can be adjusted to allow the noise to be partially or completely eliminated. Push pull act ion of the output stage also means even harmonic distortion is eliminated. Some distortion remains however due to the switching action of the circuit. This can be ameliorated by proper selection of C1 to produce a ramping action between samples.

Although the circuit worked well as described it soon became obvious that dynamic control would be better. Essentially I needed a circuit that would automatically adjust the bias with signal level.

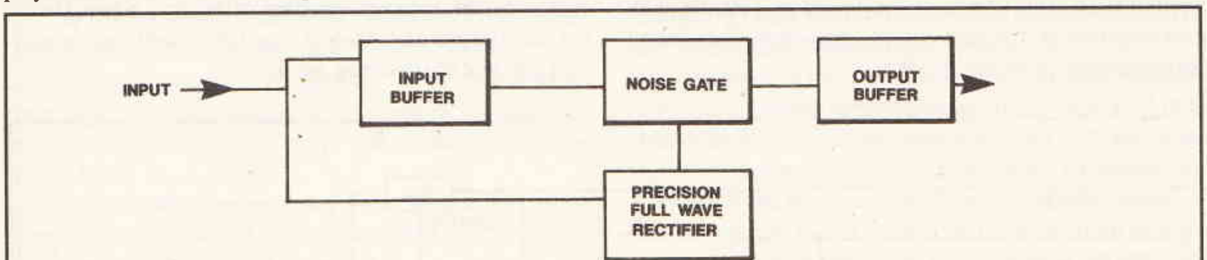


Fig.3 Block diagram

The Circuit

The circuit is best explained by reference to the block diagram. Here the input signal is first buffered and applied to the noise gate. Simultaneously the input signal is precisely full wave rectified by a pair of op-amps used in a appropiate circuit. At the input of the buffer the input signal applied to the rectifier can be attenuated by RVl. This allows the amount of 'active' noise reduction to be easily controlled.

The buffer always receives the full input signal since it's input is derived from the hot side of RV1.

The input buffer drives the two transistor noise gate. The amount of static noise reduction set by RV2. In operation RV1 is turned so that it's slider is at earth potential and the static noise reduction set by RV2. Then RV2 is advanced until the best sound is achieved. An instant comparison with the input is available via SW1.

Finally the signal is buffered out to the rest of your equipment. Having discussed the generalities lets look more closely at the actual circuit. The buffer and rectifying devices are good old TL072's. These are ideal for the purpose having the virtues of high input impedance,low noise and high slew rate. Q1 and Q2 don't have to be anything fancy either. I used BC327 and BC337's here.

Input signals are fed into IC1a's non inverting input via the DC blocking capacitor C1. RV1 provides input bias for IC1a via it's connection to the voltage divider comprising R1 and R2. Because these two resistors have the same value the voltage at their junction is half the supply voltage. C2 has a large enough value to bypass any audio at the junction to ground whilst the required DC level remains unaffected by it's presence.

The output of IC1a is directly coupled to the base of Q1. Q1 in conjunction with Q2 form the noise gate proper. Biasing is set by R3/4/5 and RV1. The latter component sets the static noise threshold. Bias current is provided by Q3 and R8 biased by the potential divider comprising R9 and R10.

C3 provides the sample and hold facility whilst the output from the noise gate is fed directly into the non inverting input of IC1b. This op-amp is configured as a simple unity gain output buffer. The unprocessed signal can also be selected for comparison purposes by SW1.

Returning to IC1a the feedback connection is rather unusual and makes use of the high open loop gain and slew rate of the TL072.

R3 and R4 have the same value so, because of the current sunk by Q3 an identical voltage appears across each. Normal op-amp action is to maintain both inputs at the same potential. The result is that the output rises until the junction of R3 and R4 'sits' at the same potential as the non inverting input. In operation the full wave rectifier modulates the curent flow through the bias circuit. As a result the voltage drop across R3 will also vary.

Suppose current flow through Q3 increases. The collector voltage will drop and this will be sensed by the inverting input. In response the op-amp's output will go positive. In the process the audio signal at the junction of R3 and R4 will be maintained the same as the input. Similarly the DC signals on the bases of Q1 and Q2 will cancel out across C3. This allows the circuit to operate in real time without needing turn on or turn off time constants. Another advantage of using Q3 is that the rest of the circuit 'sees' a high impedance looking into the collector. The result is that the AC signals at the base of Q1 and Q2 will remain identical.

The full wave rectifier circuit also needs some discussion. Although there is nothing particularly new about it it is not often seen in audio circuitry. The non inverting inputs of both parts of IC2 are held at half supply volts by the junction of R1/2. Both amplifiers are used as inverters. The particular configuration used is the virtual earth amp. The reason is that the aforementioned habit of an op-amp to keep both it's inputs at the same level. Because the non inverting input is sitting at 0V the inverting try to do likewise. The high open loop gain of the device drives the output up and down in such a way as to ensure that this happens regardless of the input signal. Hence the name virtual earth there is very little, ideally no measurable signal at the inverting input.

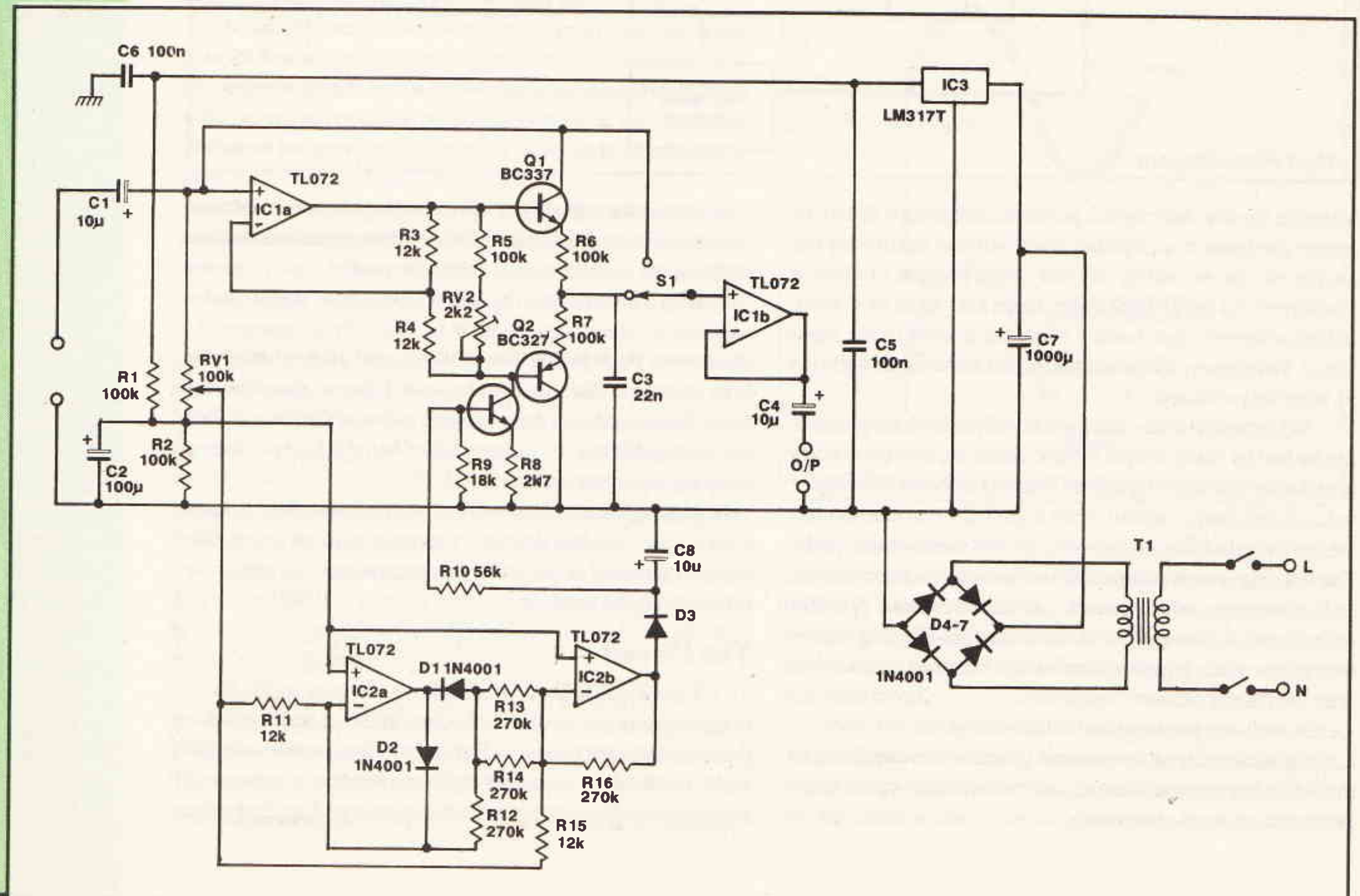


Fig.4 Circuit diagram of Dynamic Noise Limiter

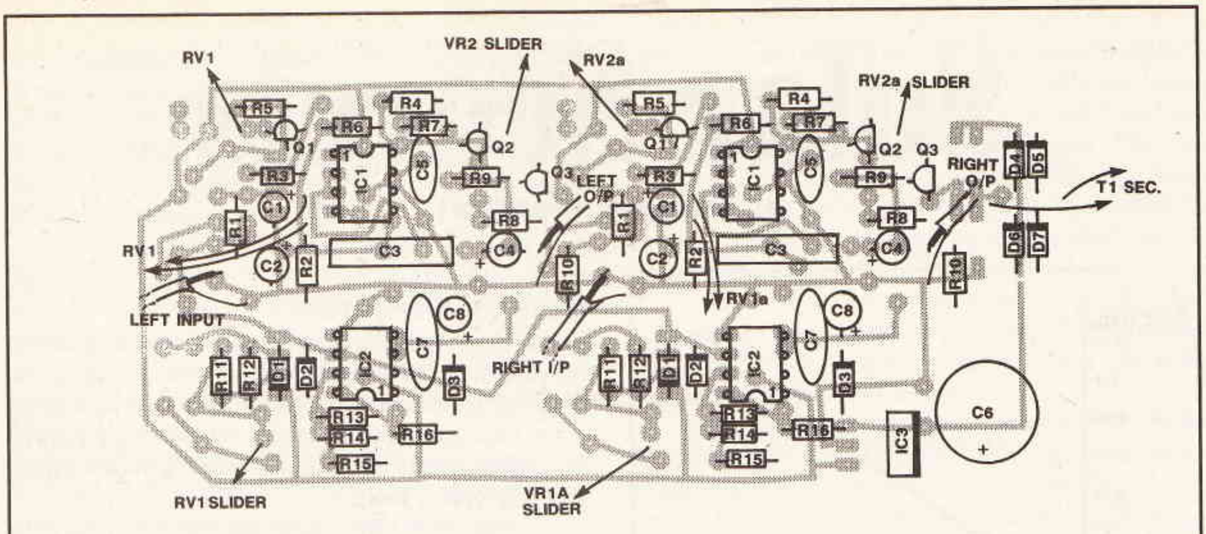


Fig. 5 Component overlay of DNL circuit

This configuration is ideal for mixer stages. Several inputs can be mixed by feeding them to the inverting input via input resistors. There is no interaction between them at the input end. The output is the sum of the inputs.

With this in mind the action of the full wave rectifier is more readily understood. Input signals are simultaneously applied to the inputs of IC2a and IC2b via the input resistors R11 and R15. IC2a is a precision half wave rectifier circuit. D1 is in series with the feedback resistor R12. When the input signal is positive going the op-amp output will swing negative. The high open loop gain will compensate for the forward voltage drop across D1 and the output will consist of negative going half cycles.

When the input signal is positive going D1 becomes reverse biased by the op-amp and effectively disconnects any output from across R12. D2 prevents the op-amp's output from going any more than about 0.6V positive. The high slew rate of the opamp ensures rapid and precise rectification across R12.

A4 adds the negative going output across R12 to the input signal fed through R15. The sum output produced by the voltage across the feedback resistor R16.

If the parallel combination of R13 and R14 has half the resistance of R15 then the whole circuit acts as a precision full wave rectifier with a positive going output signal. The output of the rectifier is half the supply voltage under quiescent conditions and this is used to complete the biasing loop via R10 and R9.

Finally the power supply. This is very conventional. The mains voltage is stepped down by T1, rectified by D4-7. The raw DC is then smoothed by C6 before being fed to the regulator chip, IC3. This provides a stable 12V supply for the circuit.

Construction

Most of the circuit is mounted upon the PCB, see layout diagram. Very little comment is required about this. Just ensure that the overlay is followed and that all the polarised components are correctly orientated. The input output pads need to be fitted with Veropins to ensure easy connection between the various sections of the circuit.

Having soldered up the board check your work thoughly to ensure that there are no dry joints or unwanted solder blobs. When you are satisfied that all is well put the board on one side. Attention can now be turned to the mechanical construction of the project.

The drilling detail is shown in the Figure. Drill out the holes for the pots ,switches etc. At this stage the case can be finished as required. The prototype was sprayed matt black. Legends were then applied with rub down lettering. These were then fixed with a clear spray varnish.

Mount the panel components and the power transformer. The board can now also be mounted on short spacers. Once all the parts are in place the final stanza can commence. Start with the power supply. Thread the mains lead through the appopriate hole. Remember to fit a grommet into the hole first!

Work your way through the wiring systematically. Remembering to use screened leads where indicated. Once you have finished the time has come to test it out.

Connect up a suitable noisy source, a distant FM station is ideal and of course your amp to the output. Wind RV1 fully counterclockwise. Similarly wind RV2 clockwise. The signal will sound the same as before. Now adjust RV1 in a clockwise direction. If you have constructed the circuit properly you will find that you are able to remove the noise but at the expense of some signal loss. Set RV2 so that the noise is just removed. Now adjust RV1.

You should soon find a point where the music seems to restore itself as if by magic. By adjusting these two controls you can set your own compromises between signal loss and noise reduction.

PARTS LIST

RESISTORS

R1,2 100k
R3,4 12k
R5,6,7 100R
R8 2k7
R9 18k
R10 56k
R11,15 12k
R12,13,14,16 270k
RV1 100k dual log
RV2 2k2 dual log

CAPACITORS

C1,4,8 10μ/25V
C2 100μ/25V
C3 22n poly
C5,6 100n ceramic
C7 1000μ/16V

SEMICONDUCTORS

IC1,2 TL072
IC3,4 TL072
Q1 BC337
Q2 BC327
D1,2,3,4,5,6,7 1N4001
IC5 LM317T

MISCELLANEOUS

Case
12V AC 1A sec transformer
2 Dual phono socket
Knobs
2 DPDT switch

Solar Secrets

by Douglas Clarkson

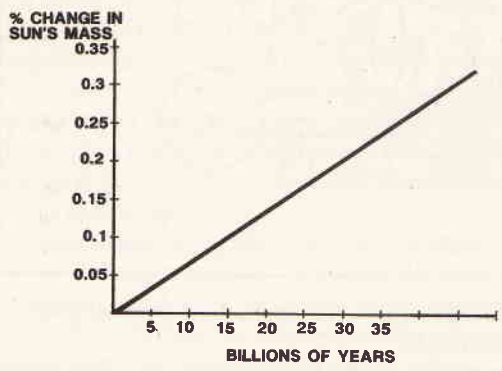


Fig.1 The figure illustrates the percentage reduction in the total mass of the sun as a function of time (millions of years).

In the structures of the ancient world, of Egypt, Babylon and the Mayas, the sun was central to human outlook both material and spiritual. Just as information is being discovered about the remotest corners of the known universe and the life and times of distant stars is being unravelled in the time of its youth, so also new insight is being gained into the processes at work in our local star. These new insights have an obvious relevance for the continued existence of life as we know it on earth and also for understanding how processes throughout the universe take place.

At a time when Nature is increasingly being watched for even the most minute of changes, it is important to understand the mechanisms which lead towards the apparent constancy of the sun. The current awareness of the Greenhouse effect, for example, needs to be clearly distinguished from any effects which derive primarily from variations of the energy output of the sun.

There is considerable yet 'unfashionable' evidence which suggests that as well as the sun having an 11 year sunspot cycle, it also has a 76 year cycle in which it changes its size very slightly. This results in an associated detectable output change in solar radiation. It is naturally important to fully understand this latter phenomenon if it exists and account for it in global warming theory. The historically developed theories of the sun's evolution and function have largely been derived from terrestrial observation. New observing satellites, however, have been able to identify startling new mechanisms at work within the sun's structures. Also, some of the great puzzles in Cosmology such as 'Dark Matter' are thought to involve in some way the function of the sun. This consideration arises out of the fact that scientists now think that only some 10% of the physical matter of the universe can actually be detected. In what form the missing mass of the universe exists is still unknown though it may modify the functioning of the sun at its core.

Historical Perspective

It is a common observation that any observation of Science is placed within a framework of contemporary thinking. As additional mechanisms become understood, then new theories can be formulated. This is exactly what has taken place with the understanding of the mechanism whereby the sun radiates its vast amount of energy.

Early ideas suggested the sun achieved its energy by means of gravitational contraction. Where a mass m_1 falls into the sun from 'infinity' to its external diameter d, an amount of energy E is given up equal to:

$$\frac{G\, m_1\, M}{d}$$

where G is the gravitational constant 6.7×10^{11} Nm^2kg^{-2}. In the example of a 1 kg mass falling into the sun this would release an energy of nearly 10^{11} Joules or the energy output of a 1000 Megawatt power station for nearly 2 minutes.

If the sun had formed from the attraction of a cloud of particles, debris and dust, then the initial release of kinetic energy would have been able to heat up the component atoms to temperatures of millions of degrees.

If the sun was contracting, it would release large amounts of energy. Noted 19th century scientists such as Lord Kelvin using the laws of physics then known estimated that the age of the sun was between 20 million and 200 million years old. Moreover if the sun continued to cool gracefully then in this simple model the sun could contract at a rate of 50 metres per year. This posed a point of great difficulty to Evolutionists who from geological evidence were advocating ages of the earth in several hundreds of millions of years. As the orthodoxy of long term evolution became established, the lack of understanding of the sun's vast energy output became one of the irritating problems of science which refused to go away.

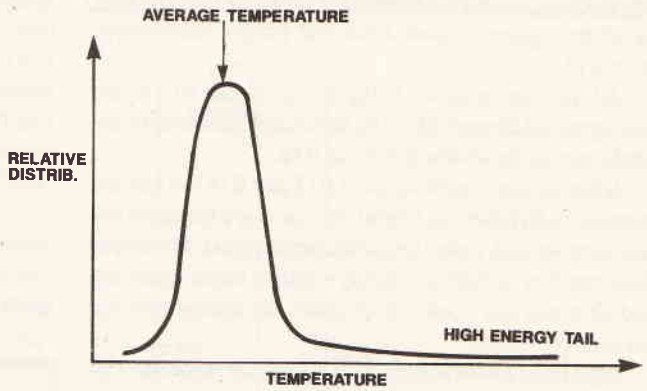


Fig.2 Typical distribution of energies of protons about a mean temperature value. Only protons on the very extreme right can take part in fusion reactions.

In trying to assess the colossal power output of the sun it is perhaps relevant to consider that each square centimetre of the sun radiates approximately 6kW and that the sun has a diameter of 1,392,000km (864,950 miles).

It was not until the scientific horizons were lifted to discover radioactivity, that other possible mechanisms to explain the solar power source emerged. Theories were advanced that even a relatively small amount of Radium in the sun could release large amounts of radiative energy. Also, it was realised that the earth will be releasing heat from radioactive decay of elements within its structure and this too allowed the age of the earth to be extended considerably. Longer time scales for the evolution of life within the solar system became more respectable.

It also becomes apparent that in the life cycles of stars and the evolution of life on planets such as the earth, the timescale of solar 'constancy' needs to be sufficiently long to give time for life to evolve and develop from very humble beginnings. On the face of it therefore, the processes of stars such as our sun appear to be in the main extremely predictable and constant.

The revolutionary theories of Einstein were the building blocks on which all modern theories of the mechanisms of the sun were founded. Whilst the equivalence of mass and energy was firmly accepted by the scientific community, it was not until Arthur Eddington published his acclaimed 'The Internal Constitution of the Stars' in 1926 that the subject of Astrophysics was born and a new set of theories were applied to the mystery of how the sun can sustain such vast levels of energy loss.

In simple terms, Einstein has shown that there was an equivalence between mass and energy and that mass could be transformed into energy via the equation:

$$E = m c^2$$

The total energy output of the sun is estimated to be 3.86×10^{26} W and this corresponds to the rate of loss of mass 4 million tonnes (1tonne = 1000kg) per second. Figure 1 shows the future estimated percentage change in the mass of the sun with time assuming this will be constant over time in the future.

The ability of the sun to sustain its energy output in this calculation shows that in the next 5 billion years if the sun continues to radiate at the same rate then its mass will only have decreased by about 0.03%. In the next 15 billion years which is considered the present age of the universe since the 'Big Bang', it will only apparently decrease by about 0.1 %.

The mechanisms of energy production rely upon streams of 'cycling' elements within the sun's core. These mechanisms have a more finite life so the relative change of the mass of the sun compared to its total mass does not give an accurate assessment of the lifetime of the sun.

In the formulation of a theory of a star, Eddington explained how stars were basically influenced by two opposing forces gravitational attraction and radiation pressure. Stars can be considered in Eddington's analysis to exist have masses expressed in the form 10^n grammes where n = 1,2 3 and upwards. The release of energy following the gravitational collapse of material into the star results in an increase in its temperature. The larger the star the higher will be the temperature of its core. Stars however, have to be sufficiently large to raise their temperature to levels at which nuclear process can be triggered. This defines the lower value of n.

Where photons 'bounce off' hot star particles, they exert a radiation pressure on the particles, so the energy radiated from the star will tend to push the material of the star outwards. There is a point at which the radiation pressure can be sufficiently great to literally blow the star apart. The lower limit of the mass of the star and the higher limit of the mass of the star according to Eddington defined the limits within which stars could function. The lower value of n was identified as 32 and the higher 35. Our own sun fits nicely into the picture with its value of n of 33.

Thus larger stars should be hotter since they have released more gravitational energy initially from the material which gravitated inwards to form them. Smaller stars should be cooler and release less energy.

The concept of radiation pressure is not one which is frequently referenced. It arises by the change in momentum of photons of light as they, for example, are absorbed by a surface. If radiation of intensity I (Joules/m^2) is incident upon area A, the force experienced by an absorbing surface is given by:

$$F = \frac{I}{c}$$

where c is the speed of light. At the surface of the sun where

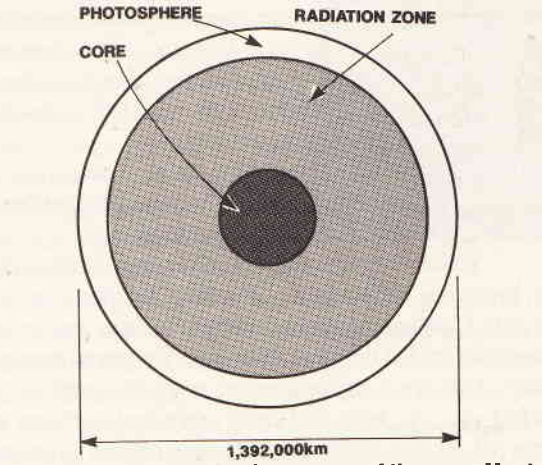


Fig.3 Diagram of internal structure of the sun. Most of the sun's mass is located in the core. Radiation eventually escapes from the radiation zone and passes upwards through the convection zone.

the intensity is 6.35kW/cm^2, the force on a 1m^2 area absorbing surface is some 0.2 Newtons. This is approximately equivalent to the force on a mass of 20 grammes in the earth's gravitational field.

It was during the late 30's that progress was made in putting forward mechanisms for the production of the sun's vast energy output. While initially it was the CNO cycle (Carbon, Nitrogen, Oxygen) which was 'discovered', this mechanism is more important in hotter stars than in our sun. The important mechanism for us is the Hydrogen to Helium chain one which on earth has been demonstrated in all too graphically in the form of the Hydrogen bomb.

It is said the sun's core has a 'temperature' of around 15 million degrees Celcius. This is in fact a description of the most likely kinetic energy of a hydrogen nucleus. In fact the distribution of energies looks like the curve in Figure 2, where there is a 'majority' temperature but also a so called 'high speed tail' where a small number of nuclei are travelling at much higher temperatures.

The 'temperature' of such a proton can be expressed in terms of a value of kinetic energy using the formula

$$\frac{3kT}{2}$$

where k is Boltzman's constant and T is the absolute temperature of the proton.

Table 1 describes the values of energy in so called keV (electron volts x 1000).

It is only the very few atoms in the 'tail' of the curve which will have sufficient energy to take part in proton-proton reactions.

Millions degrees	keV value
50.0	6.5
100.0	12.9
150.0	19.4
200.0	25.9
250.0	32.3
300.0	38.8
350.0	45.3
400.0	51.8
450.0	58.2
500.0	64.7
550.0	71.2
600.0	77.6
650.0	84.1
700.0	90.6
750.0	97.0
800.0	103.5
850.0	110.0
900.0	116.4
950.0	122.9
1000.0	129.4

Table 1

Table 1: Values of proton kinetic energy (keV) as a function of proton 'temperature'. Only protons in the very high energy tail of the distribution have the necessary energy to take part in fusion reactions.

The first confirmation of the processes of transmutation of elements was achieved in 1932 at the Cavendish Laboratory at Cambridge under the direction of Cockcroft and Walton. Lithium atoms were bombarded by high energy protons and the resulting pair of alpha particles (Helium nuclei) which were released could be easily detected. The composition of the sun was assumed to be unsuitable for sustaining fusion reactions involving hydrogen.

In the proton-proton interaction, the process of energy release is thought to take place in three distinct phases. Initially two hydrogen nuclei fuse to form a Deuterium nucleus consisting of a proton and a neutron. The Deuterium nucleus then interacts with another high energy proton to create a He^3 nucleus. When two of these He^3 nuclei fuse together, high energy protons are released.

At the estimated mean temperature of the core of the sun of 15 million degrees only one proton in 100 million is travelling fast enough to initiate this reaction.

While the proton-proton cycle is largely responsible for generating the energy of the sun, the transition of 4 discrete hydrogen nuclei into a single Helium nucleus has consequences for the size of the sun. The super energetic nuclei in the sun's core behave very much like particles of a perfect gas. As more and more individual particles are removed, the size of the sun will decrease slowly. As the core shrinks in size, gravitational energy released will increase its temperature and the proton-proton fusion process will have more high energy protons available. The increased level of radiation will tend, however, to expand the sun's outer layer as additional energy is radiated outwards from the core.

It is thought the warming of the sun's core over the past 4.5 billion years has increased its brightness by about 40%. This amounts to a percentage change on average of 0.00000001 % per year which probably cannot be detected. In a further 6 billion years, the mass of hydrogen in the core could well be depleted and the fuel essentially will have run out.

When the process of hydrogen burning slackens, the radiation pressure will fall allowing the star to contract and in so doing release gravitational energy. Hydrogen will then be able to 'burn' in the outer layers of the sun and it will as a consequence expand out these layers of hydrogen rich material.

During this phase, the core of the sun will increase in temperature and a point will be reached at which Helium 'burning' can take place. Sufficient energy will be released to blow off most of the outer layers of the expanded sun. When it settles down, the sun will become a hot Helium burning star.

It is estimated that large stars experience more rapid transitions from one phase to another. Many stars in our galaxy which were created at approximately the same time as our own have long ago undergone their various phases of youth, maturity and old age. Can this indicate that life as we know it is more favoured in planetary systems supported by suns of the size of our own?

One byproduct of the phase of Helium burning when it does take place is Carbon 12 a key element in all forms of life on planet Earth. When much more massive stars cycle through phases of higher and higher core temperatures, additional 'burning' mechanisms become possible. Carbon and Helium can fuse together to form Oxygen. Carbon can interact with carbon to create a range of products including neon, magnesium and sodium. At higher temperatures still 'oxygen burning' takes place creating elements such as silicon and sulphur. The final end point element in this cycle is the creation of Iron 56. Our sun is not likely to enter into this dynamic phase of element production, it simply is not massive enough.

Figure 3 shows the internal structure of the sun as estimated from current models and theories. The core is considered to extend to 25% of its total diameter and while occupying some 2% of the sun's total volume (proportional to the cube of its radius), well over half of the sun's mass is concentrated in its core.

Energetic gamma ray photons released by nuclear fusion reactions pass outwards into the radiation zone where countless interactions take place with energetic nuclei. The level of interaction is so high that it is estimated that it takes on average 10 million years for a photon to make its way through the extensive layers of the radiation zone. This is all to do with concepts of mean free path between collisions and the associated 'random walk' of photons.

By the time photons reach the upper edge of the radiation zone, they have sufficiently lowered their energy to interact with the material in the convection zone to be able to pass on their energy in the form of kinetic energy and transfer heat to the outer structures. The sun's energy is now streaming outwards, partly escaping through the convection zone and out into space and partly into establishing convection currents where waves of heated gas rise up to the surface, cool and fall back.

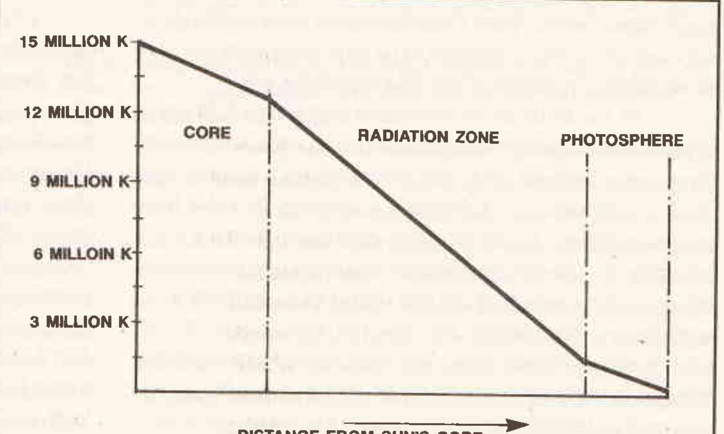


Fig.4 Diagram of temperature profile within the various regions of the sun.

Figure 4 shows the variation of temperature within the sun ranging from the core where the process of hydrogen burning takes place to the visible surface of the photosphere.

The bright disc of the sun is called the photosphere. Careful (and safe) photography of its surface reveals it to be a patchwork of granules which change patterns as new volumes of heated material rise up from below and spread out over its surface. Sunspots are in fact areas which are slightly cooler than the areas around them.

The Neutrino Safari

The models of the solar nuclear processes predict the release of large numbers of particles called neutrinos meaning in the singular 'little neutral one'. After having first been postulated by Wolfgang Pauli in 1931, Enrico Fermi 'adopted' the concept into the official world of nuclear physics in 1932. The neutrino has an extremely low cross section of interaction it very rarely interacts with matter and so for many years remained undetectable. The scenario of physics has on numerous occasions been the process of detecting the undetectable. In 1956, Reines and Cowan detected neutrinos generated by a nuclear reactor where the flux of particles was estimated to be some 30 times higher than that received at the earth from the sun. It was not until 1964, however, than an attempt was made to determine the levels of solar neutrinos. Figure 5 shows the giant tank of perchloroethylene constructed in the Homestake gold mine in South Dakota.

A neutrino which interacts with the isotope of Chlorine Cl 37, causes a neutron to change to a proton, forming an atom of Ar 37. This in turn decays with a half life of 34 days releasing in the process a characteristic gamma ray which can readily be detected. Typically as the entire contents of the tank are processed each month to release the key Ar 37 atoms, only 12 counts are detected.

This poses a problem for the theorists who predict that levels should be at least three times higher. Where have all the solar neutrinos gone? Can this one experiment indicate that the detailed solar models are wrong? Some recent developments in Cosmology may throw new light on the problem.

The Search for Dark Matter

One of the more startling thoughts to strike cosmologists in recent years is that if the gravitational behaviour of the galaxies is observed, then at least 90% of the mass which makes up the matter in galaxies which takes part in gravitational attraction is 'missing'. Thus even accounting for stars, black holes, planets, dust clouds, comets and gas clouds in a typical galaxy there is considerably more matter which cannot be detected this is the so called dark matter.

Astrophysicists have considered how such dark matter could relate to the function of the sun. One theory is that the sun has through time attracted a considerable amount of dark matter to its core and this acts to cool the core slightly and hence reduce the rate of production of neutrinos. This is still an open chapter as preliminary experiments are only now being set up to try and detect dark matter particle interactions with the 'visible' particles of our physical world.

One of the current 'in' terms is WIMP meaning Weakly Interacting Massive Particle. This term can be applied to particles which may make up such dark matter.

As ever, theorists are never slow to come up with possible solutions to the low number of solar neutrinos observed. One theory considers that it could be possible for neutrinos to 'flip' from one type to another in the time of their flight to the earth. It is likely that more sophisticated neutrino detectors which are planned or in the process of being completed during the 1990's will resolve the solar neutrino mystery.

Solar Cycles

Much attention is given to the 11 year sun spot activity of the sun. The number of sunspots visible has been observed to increase and decrease in a predictable way. The sun reached a peak of such activity early in 1990 and is therefore at present declining in its activity. Significant interest is now being shown in making accurate measurements of the SIZE of the sun following indications that the sun experiences phases where its apparent size increases and decreases. Following contradictory observations by separate groups, Ron Gilliland took data from a broad range of sources and concluded that the sun could be shrinking at an overall rate of 0.2 of an arc per century. Superimposed on this long term reduction was a 76 year cycle with the sun passing through a minimum point in 1911. Between then and 1987 the sun was expanding and therefore cooling slightly so opposing the greenhouse effect. As the sun begins to shrink in size it will tend to increase in temperature and add to the greenhouse effect contributing an estimated 0.3% of energy output and increasing the average temperature over the globe by 0.25° Celcius. There is not sufficient data, however, to determine if the overall reduction in diameter since 1700 is part of a cycle with a much longer period.

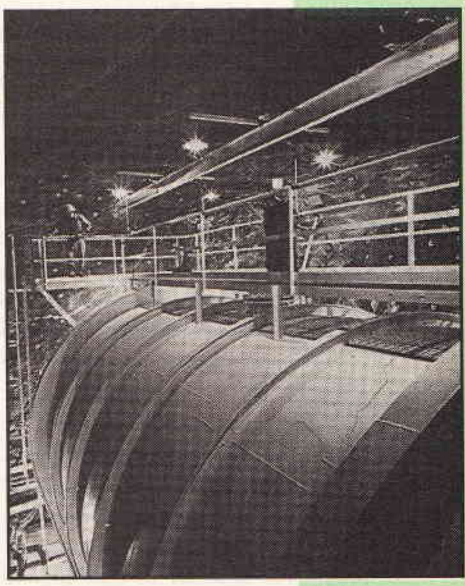


Fig.5 Illustration of the neutrino detector in the Holmstake gold mine of South Dakota.

This is why groups of solar scientists have started to undertake some basic research into measuring the size of the sun.

Solar Resonances

More detailed observations of small sections of the sun's surface reveal that they oscillate vertically in timescales of 5 minutes with displacements of about 50 km. From observations of Doppler shifts in the radiation emitted from specific locations, it has been determined that such disturbances involve the entire sun's surface. Present theories indicate that the sun is behaving like a spherical 'gong' in as much as its surface is vibrating as a result of complex modes of surface acoustic resonance.

Just as the surface of a drum can resonate in specific resonant patterns, so the solar surface can resonate with a multiplicity of resonances which involve the entire solar surface.

Figure 6 shows what is considered to take place with a shock wave of vibrational energy which passes from the surface of the sun towards its core. The direction of propagation of the waves are bent into circular paths due to the changing

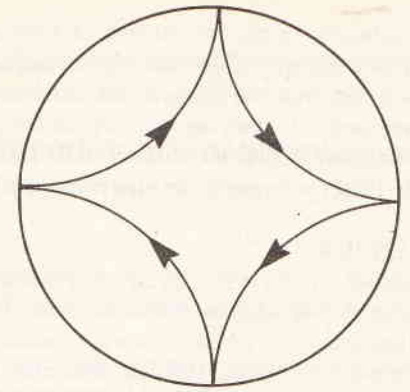


Fig.6 Section through sun showing contours of resonant modes of vibration within the sun which result in patterns of surface oscillation

densities of material in the sun. Modes of vibration which are encouraged are those which 'complete the circuit' of a vibrational path through the sun's structure. The frequencies of these waves is quite slow about 0.03Hz.

While the Solar Heloiosphere Observatory is planned to be launched in 1995 and will include an experiment to record movements of the solar surface, an earth based research group GONG Global Oscillation Network Group is already collecting sets of data.

While a specific satellite can maintain a constant contact with the area it is observing, the ground based GONG programme requires at least six sites to preserve continuity of measurement on a rotating earth.

In many ways the real complexity of such a project lies largely in the extensive data processing required. Vast amounts of data require to be captured, catalogued and processed to work out the various modes of resonance which are active upon the sun's surface.

One of the interesting concepts which such work throws up is about the nature and extent of the radiating surface of the sun. If the sun's surface can be imagined to be 'smooth' in a nonresonating system or in a state of resonant agitation , it could give rise to variations in its effective radiating surface area an important aspect for its radiative cooling mechanisms.

Understanding of the sun's cycles of 'size' variation could involve appreciating how its surface resonant energy is distributed within the sun's structure. There must, for example, be a significant amount of energy 'locked up' in this mode of resonance.

Conclusion

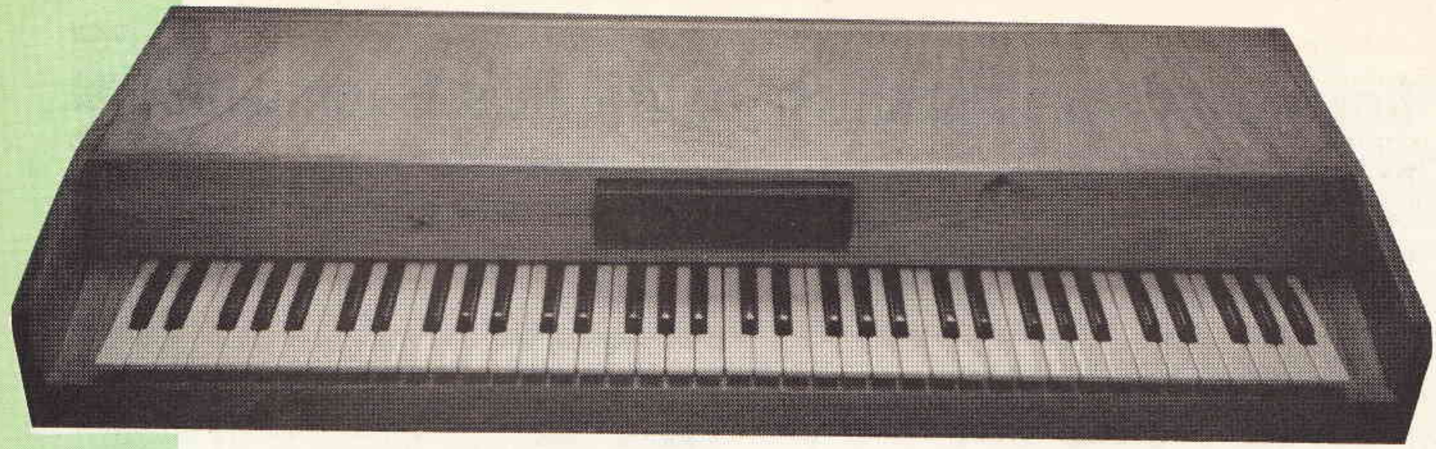
While it is good to be aware of details of the remotest corners of the known universe, some surprises have recently come to light 'in our own backyard' as aspects of the sun become better understood. During 1990's much additional scientific information will become available from new research programmes and enhance significantly our appreciation of the sun's inner workings.

Further Reading:

Blinded by the Light: John Gribben, Black Swan Publications 1991.

The Particle Explosion, Frank Close, Michael Marten and Christine Sutton, Oxford University Press, 1987.

Sun and Earth, Herbert Friedman, Scientific American/W.H. Freeman, new York, 1986.



MIDI Keyboard

A quick and economical way to turn an ordinary keyboard in to a synthesiser. Stephen Lenham tells us how it can be done.

How faithfully can electronics simulate a piano? That was the question which inspired this project. I needed an instrument with all the character of the real thing, but without the size or the price tag.

With the availability of inexpensive MIDl synthesizer modules, it seemed foolish to 'reinvent the wheel', even if such sound quality could be achieved by an anateur design. So attention was turned to creating some electronics that would generate the MIDI codes to drive such a module.

ETI's 'How to MIDI a Piano' article provided a suitable circuit, but the double–sided PCB, microprocessor, and the need to program an EPROM

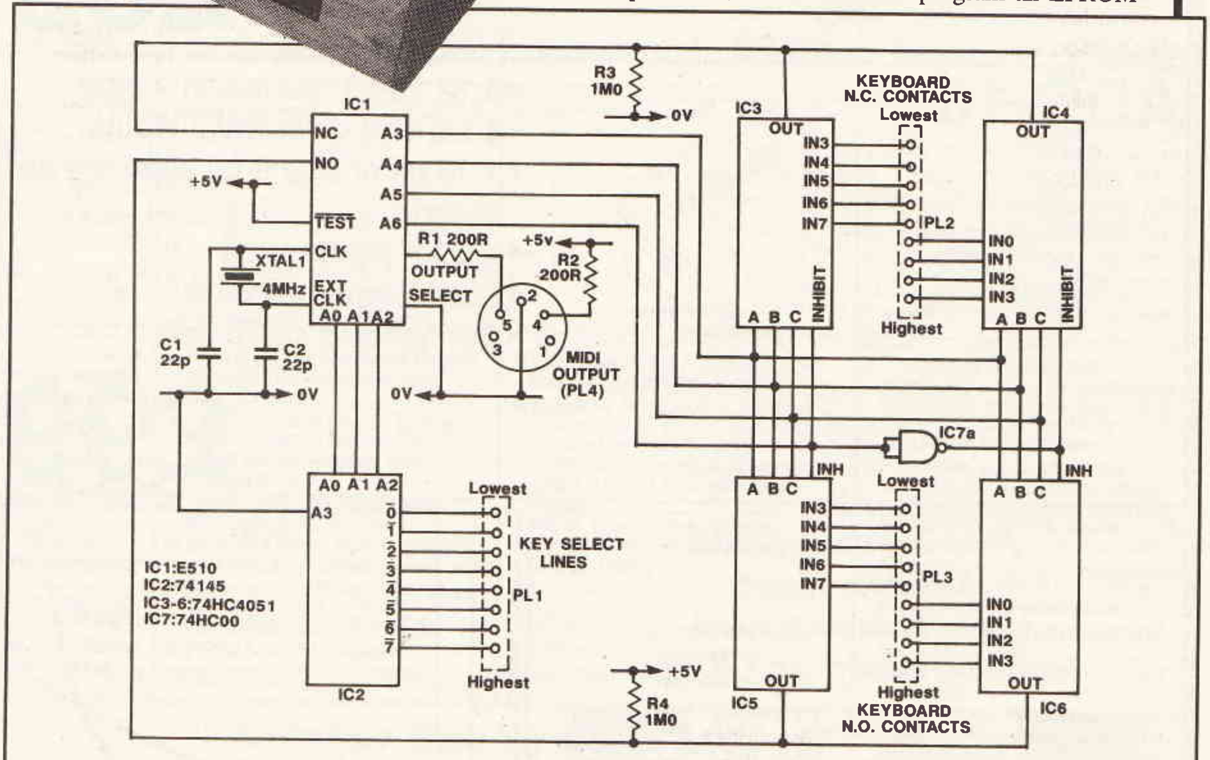


Fig.1a Main circuit diagram of MIDI control

PROJECT

were daunting. The E510 IC, recently introduced by Maplin Electronics, came to the rescue. All the functions required to scan a keyboard of virtually any size, including full polyphony and velocity sensing (whereby the volume of each note depends upon how hard the key was played) have been integrated onto this IC.

Costing about £25, it is possibly one of the most expensive 16–pin ICs many people will have come across, but the fact that only a few extra ICs for address decoding and a handful of passive components are required to forn a conplete keyboard scanner and MIDI code generator makes this small component very attractive.

Here, a circuit including the E510, a six–octave keyboard, and a Yamaha EMT10 sound expander module are brought together within a specially made wooden box to create an instrument which, to answer my original question, sounds very much like a real piano.

Building The Circuit

Construction is made simple by the use of a single sided fibreglass PCB. Referring to the component overlay in Figure 2, insert and solder the components, starting with

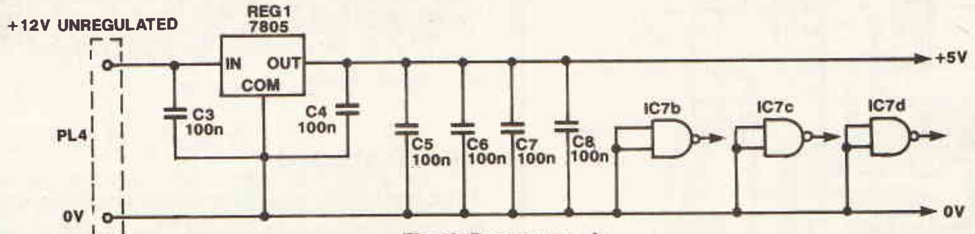


Fig.1b Power supply

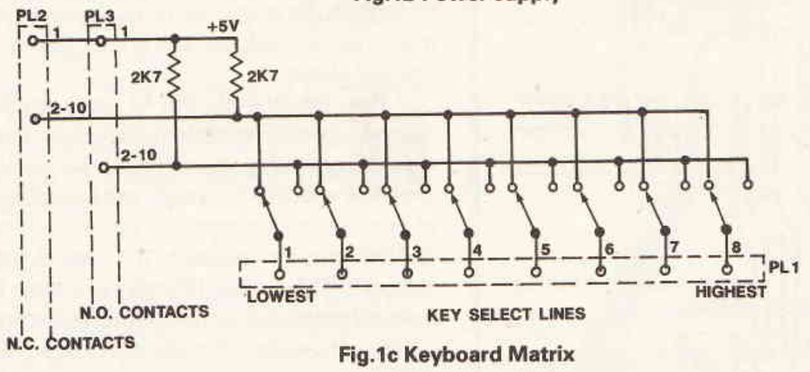


Fig.1c Keyboard Matrix

HOW IT WORKS

The E510, IC1 in this circuit, performs all the active operations necessary to read the keyboard. It is a high–speed, high–performance IC. fabricated in HCMOS, and this,is must be borne in mind when designing its support circuitry, as will be seen later. The system clock is derived from a 4MHz crystal, the internal oscillator of IC1 also dividing this frequency to produce the 31.25kHz (kbaud) MIDI data rate.

Essentially, the E510 cycles through a seven–bit address range (A0–A6), each location corresponding to an individual key on a keyboard. Seven–bits allow 27 or 128, keys to be scanned – ten octaves! Obviously this is unfeasible practically, and more importantly many MIDI instruments are physically unable to respond to pitch codes outside a certain range. So in this design a more sensible six–octave range has been chosen, located approximetely in the middle of the potential ten. That way, only extremely low and extremely high notes are lost.

Each key contact consists of an SPDT switch, with a normally closed and a normally open contact. Each of these is pulled–up to +5V with a resistor. The 'colon' terminal of the switch is connected to 0V, and will therefore pull one of the contacts low depending upon whether the key is depressed or not. By measuring the time between the normally closed contact opening and the normally open contact closing, which is what happens when a key is pressed, the E510 is able to deduce how hard the key was played and hence provide velocity information.

It would be mechanically complex to provide each key with a separate switch, so a multiplexed arrangement is used. The keys are organised in a 9 x 8 matrix, whereby each group of eight consecutive keys shares a normally closed and normally open contact, or busbar. (Please refer to the circuit diagram (Figure 1 for clarification.) Each key has a separate 'common' pin. With a little thought it can be seen that when we are reading from one set of busbars we are uninterested in what is happening on all the others and so every first 'common' in each group can be connected to all the others, as can every second, every third, etc. leaving only eight common lines. Hence the 9 x 8 matrix; nine sets of busbars, and eight 'common' lines.

The process of scanning therefore involves selecting one set of busbars and connecting them to the NC and NO inputs of the E510. Each common line is then taken low in, turn, and the state of the busbars read. When all eight keys have been read, the next busbar is selected, and so on. This can be done by decoding the lower three address lines to give the 'common' signals, and the upper four to control two 16–way selectors with only 9 inputs used on each.

The circuit realisation uses a 74145 4–to–10 line decoder to take the 'common' lines low. The open–collector outputs ensure no interaction between groups of keys. A standard TTL device was chosen because each output has to pull nine busbars low simultaneously – and quickly. HCMOS has the speed but not the drive capability, while the 74LS145 has an output saturation voltage of 2.3V; dangerously close to the HCMOS upper logic threshold.

Two pairs of 74HC4051 eight–way analogue switches form the 16 way selectors, with IC7a allowing the most significant address line to choose between the two IC's by manipulating the INHIBIT inputs. The 4067 sixteen–way switch, would have been ideal here, but is unavailable in HCMOS, the only CMOS family fast enough to keep up with the E510.

It is also interesting to note here that during prototyping the function of IC7a was performed by a simple transistor inverter, whose inability to perform fast enough manifested itself when several keys on the keyboard triggered more than one note each from the synthesizer. Traps for the unwary, as mentioned before!

The outputs of the selectors go straight to the appropriate E510 input. R3 and R4 simulate an unpressed key when an unused selector input is addressed, to avoid spurious responses.

The MIDI output comes direct from ICl, R1 and R2 limiting the maximum output current.

The 5V supply for the circuit is derived from the 12V supply to the synthesizer module by a standard 7805 regulator, and liberal use is made of small decoupling capacitors.

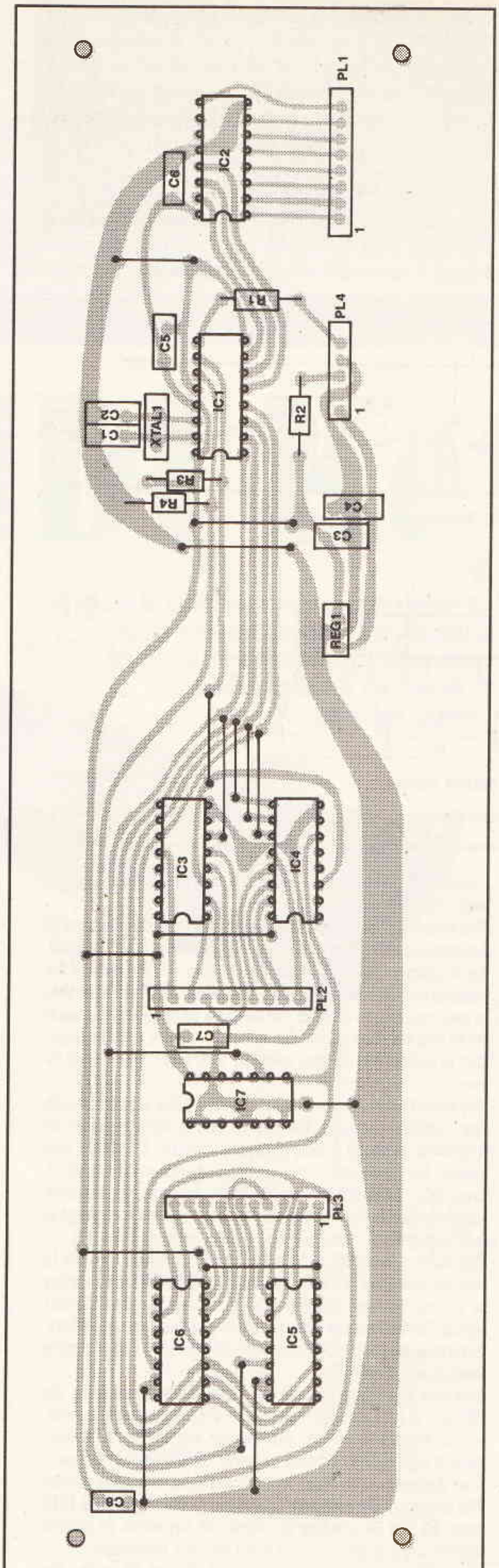


Fig.2 Component Overlay

the 17 wire links and continuing with resistors, capacitors, IC sockets and the crystal. Uniting the PCB plugs and sockets makes correct orientation easier – pin 1 is marked on the socket housing. Getting this right will simplify connections later on. A small piece of aluminium bracket, about one inch long, is ample heatsinking for the 7305, and this should be bolted on before soldering the regulator to the PCB. Heatsink compound is not necessary.

Do not insert the ICs at this stage. Thoroughly check for shorts, solder bridges and dry joints, then apply 8–15V DC. to PL4 (see wiring diagram for connection details) and check with a DVM that 5V appears at pin 16 of each IC socket (pin 14 of IC7). With this test completed, the ICs may be inserted, checking the polarity of each since some are 'upside down'.

The Keyboard

To be compatible with the scanning electronics, a keyboard must consist of approximately 72 keys with a common pair of busbars for every set of eight consecutive keys and every first, second, etc. key in each group commoned, as detailed above.

How much work will be necessary to achieve this depends upon the available mechanism. For instance, J & N Bull Electrical supply surplus three–octave mechanisms complete with SPST contacts and keys already configured in a matrix arrangement. The prototype used two of these mechanisms; unfortunately the need to modify the contacts to SPDT and the difficulty in linking the two halves make this approach far from ideal. Therefore I will assume a bare mechanism, bereft of contacts, and start from scratch.

Figure 4 shows the suggested approach. A long strip of Veroboard (or similar) is mounted securely along the bottom of the keyboard. (In practice, several lengths of board will be necessary to span the entire length.) For each key a length of gold–clad phosphor–bronze contact wire is soldered to the board, bent over parallel with the board, and passed through the hole in the key plunger such that the wire flexes in sympathy with the motion of the key.

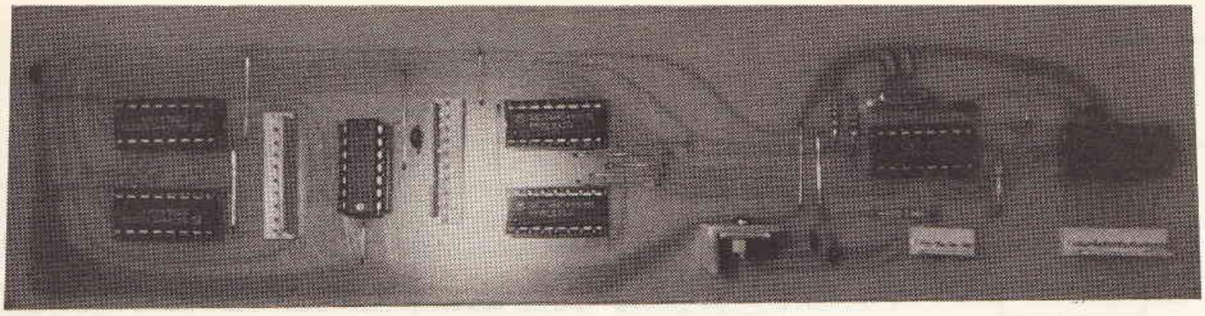
A busbar assembly consists of two support blocks as shown in the diagram and two lengths of 2.5mm diameter brass rod. The bars should be sufficiently long to allow contact with all eight contact wires but not so long as to cause shorting between adjacent assemblies. Push the rods through the holes in the support blocks, with one block at each end. The assembly is then glued to the board in accordance with the following necessary rules:

1 The key wires must be in contact with the NC busbar whenever their key is unpressed.

2 Each key wire must leave the NC busbar shortly after the relevant key is depressed, and make contact with the NO busbar slightly before the key reaches full travel. This ensures adequate contact pressure at all times.

3 Contact between the wires and the NO busbar must be maintained while the key remains depressed.

It therefore follows that since the busbar spacing is 3.5mm, the busbar assemblies should be attached at a point where the range of travel of the wires is greater than 4mm. It is *essential* that all nine busbar assemblies are consistent in their geometry and positioning if the keyboard is to have a constant velocity response along its length.



Epoxy resin is recommended for attaching the busbar assemblies to the board. While you are at it, make up a little extra resin and use it to secure the busbar rods in the support blocks. When this is set, any 'rogue' key wires that do not obey the above rules can be tweaked into conformity.

Wiring up the Keyboard

Connection between every first key in each block, etc, can be made on the Veroboard, as can the arrangement of pullup resistors between each busbar and the +5V supply. 5 volts is available at pin 1 of PL2 and PL3 for convenience. Figure. 4 c) shows the connections at keyboard level, while Figure. 5 gives all the necessary interconnections between the keyboard and the scanner PCB plus external inputs and outputs. With all this in place, you are in a position to see if the whole system works! But first let us consider how to house it.

Building the Box

Although the task may appear daunting, with just a little perseverance and care it is possible to produce an attractive wooden casing for a very modest price.

Figure 6 shows the timber used in the prototype. All dimensions are final and exact, so when cutting it is wise to be initially generous. The material is pine throughout, with just the large base and top panels in plywood. With luck, a good match between the surface of the plywood and the solid pine can be obtained, especially when both have been stained.

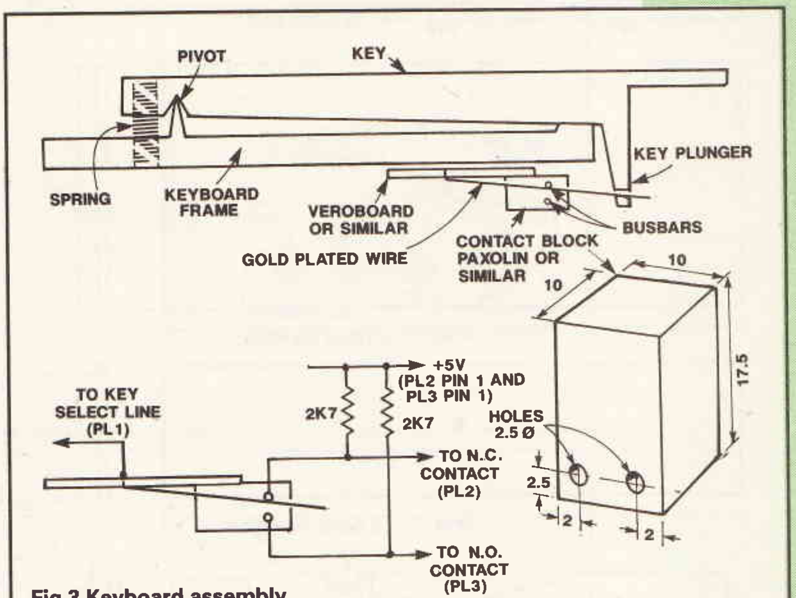


Fig.3 Keyboard assembly

Begin by cutting parts A, B and F (2 off). Part A has cutouts for the metal rear panels, the design of which is detailed in Figure 7. Also cut four pieces of scrap wood to form the corner supports, as per Figure 5. These pieces also support the top and bottom panels, so gauge their height accordingly. No's assemble these parts into a rectangular framework, glueing and screwing the lengths to the support blocks. Bear in mind that this exercise determines the

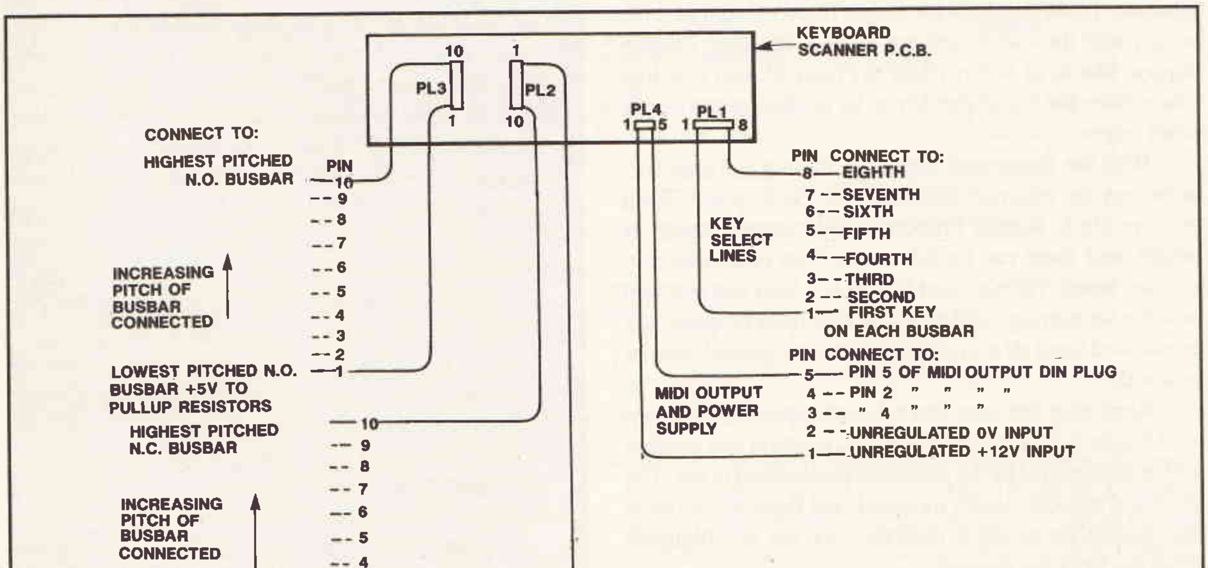


Fig.4 External Connections to PCB and Keyboard

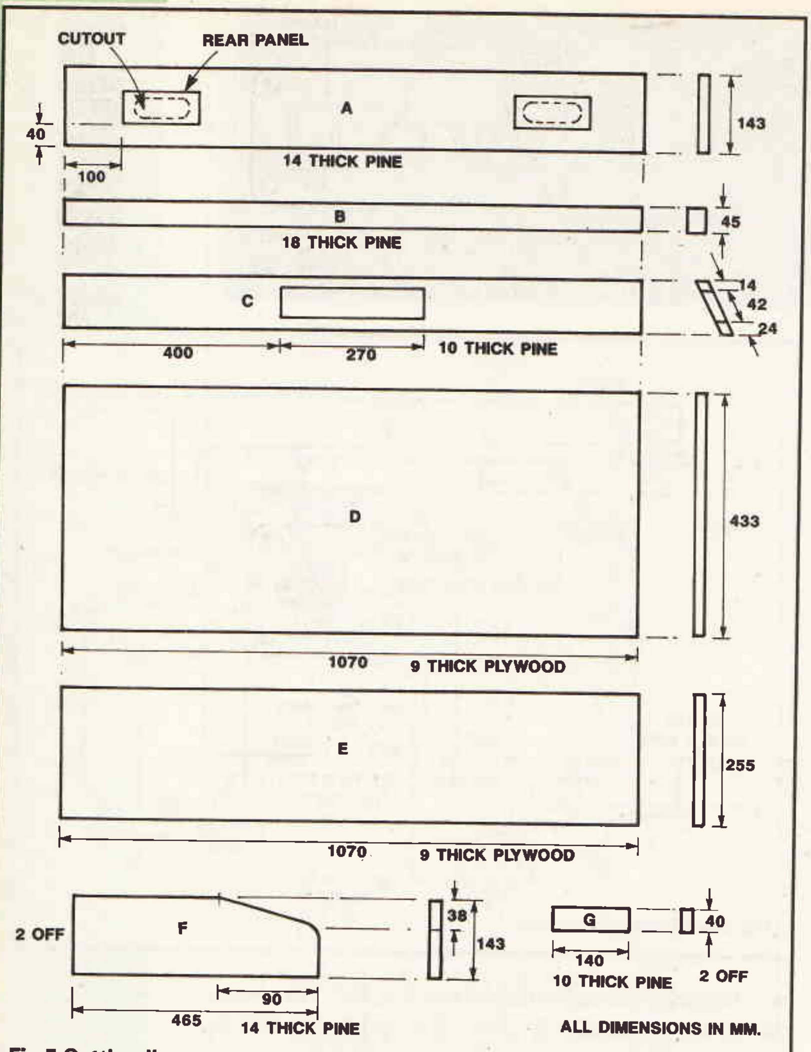


Fig.5 Cutting list

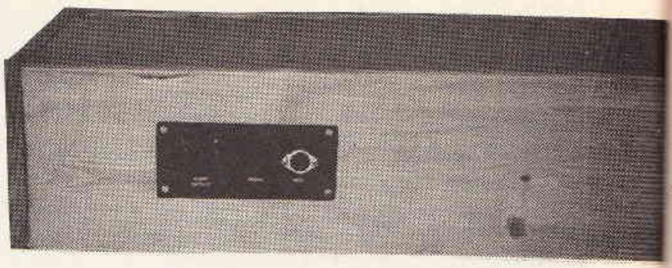
squareness of the entire instrument!

While this is setting, cut the other parts – again, making each slightly larger than necessary. Do not bevel the top edge of part C yet as this is best done with it in position. To do this, trim the length of part C so that it fits snugly into the framework but without bending. Prepare support blocks as before (refer to Figure 8), and glue into place. Now the top of part C can be planed level with the other edges.

With the framework complete, the top and base sections can be trimmed accurately to fit in place. Each requires six L–shaped brackets to add support along its length, and these can be fabricated from offcuts of aluminium sheet. The top (part E) can be glued and screwed onto the supporting blocks, but the base (part D) must only be screwed since all access to the interior is gained from its removal.

Removing the base, fit keyboard support blocks to it as in Figure 8. Their exact thickness, position and number will be determined by the particular mechanism in use. The keyboard must be rigidly mounted, and these blocks raise the mechanism to allow clearance for the key plungers when the keys are pressed.

When all necessary connections have been made to the keyboard, attach it securely to the blocks. A certain amount of trial and error will be necessary to get it in the correct place once the box is assembled, so do a trial run



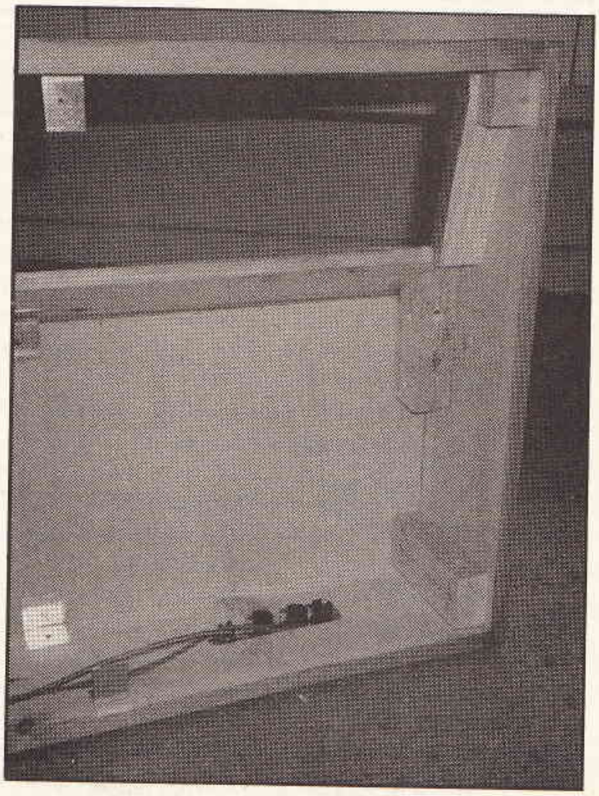
before tightening the screws. With the base/keyboard assembly in position, insert parts G on each side of the keyboard and glue in place.This concludes the carpentry. Stain and varnish is a matter of individual taste; beeswax polish produces a nice finish after they have been applied.

The rear panels can now be screwed into place. Two metalbrackets are required for the synthesizer module, and they are screwed to part C, just below the cutout. Two of the case screws from the bottom of the module are removed before sliding the module into position through the cutout. These screws are then replaced, but passed through holes in the bracket before re–entering the case, thus securing the module to the brackets.

Final Wiring

Figure 9 shows the internal wiring . The scanner PCB can be mounted on the base of the box with sticky pads; there's plenty of room! When wiring between the body of the box and the base, be sure to use a sufficient length to allow easy removal of the base if necessary. The only screened cable required is for the audio connections. When mounting the 12V power adaptor, it was necessary to drill four small holes in the plastic case for screws – this seemed to have no detrimental effect.

Now is the time to cross your fingers and try out the whole system together. Connect the audio outputs, which are at line level, to a suitable amplifier. The mains on – off switch applies power to the scanner electronics, but the



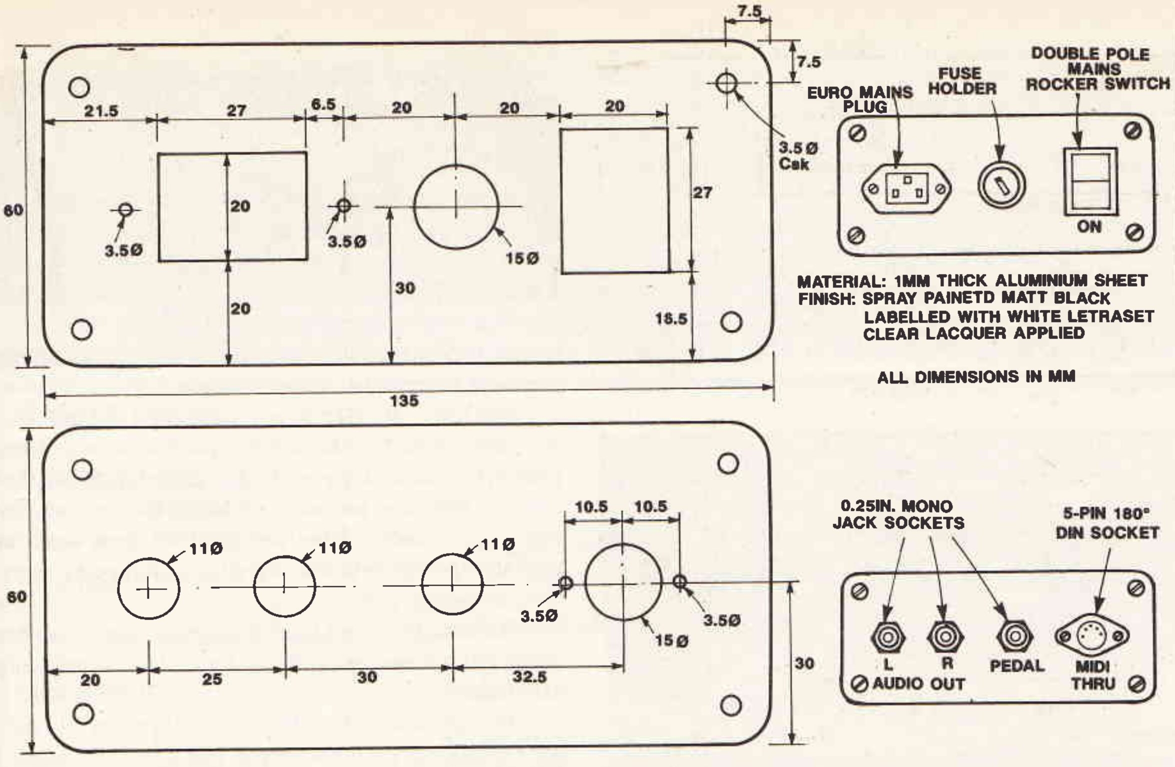


Fig.6 Rear panel layout

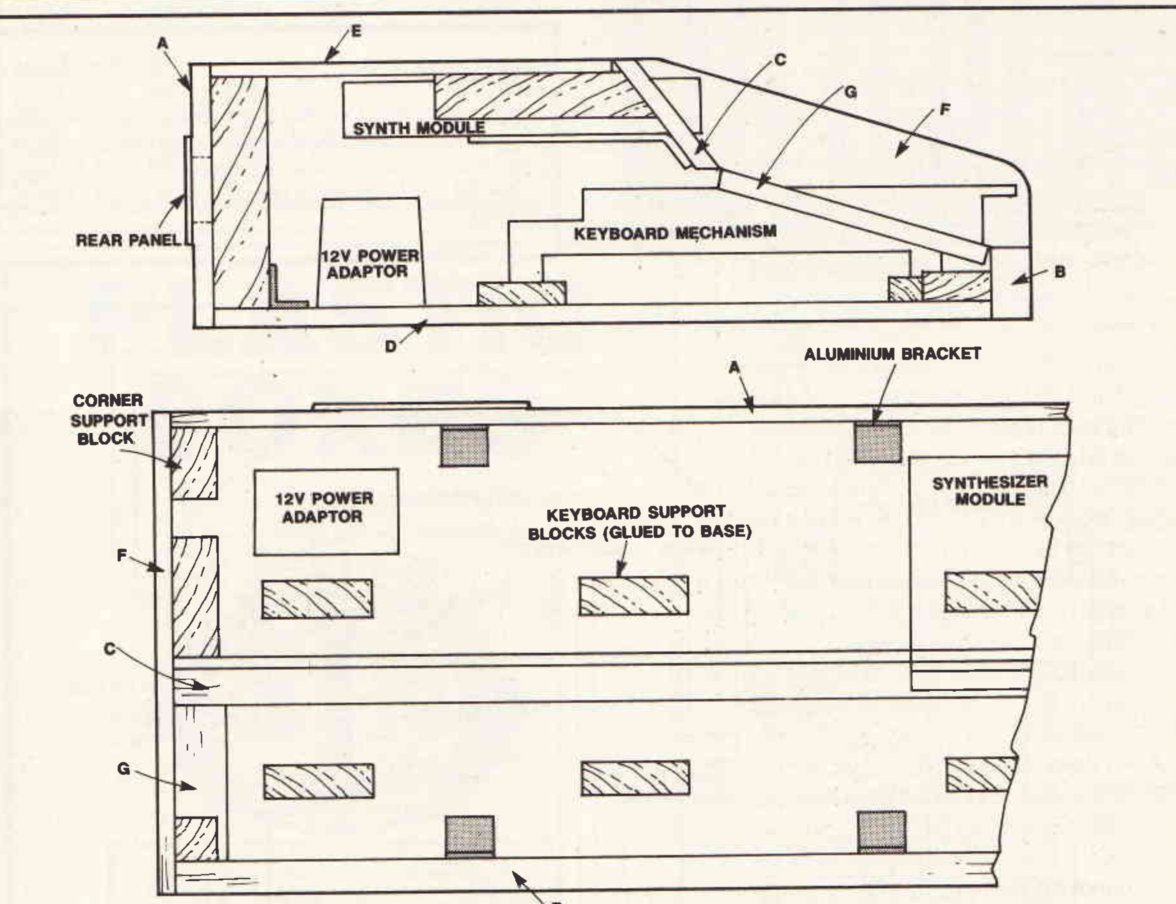


Fig.7 Box construction

synthesizer has an additional power button which must be on.

Play a key – you should hear the result. If not, double check everything. If troubleshooting is necessary, the following test points are suggested:

– follow the power feed in by checking mains input, 12V unregulated supply and the 5V regulator output.

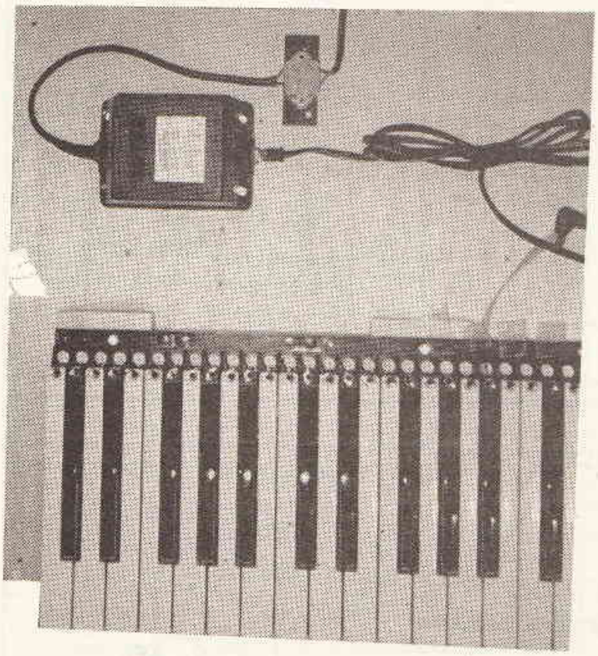
– monitor pin 9 of the E510 (MIDI output) with a logic probe. Pressing a key should result in a burst of activity, and releasing it a second burst. If this is present, then it is most unlikely that the scanner PCB is at fault.

– check for activity on the address lines, pins 1 – 7, of the E510. Inactivity on all implies absence of clock signal, or a faulty or damaged E510. Inactivity on one or two suggests a board short, or a faulty decoder IC.

With luck, all this will be unnecessary. The most likely faults are merely the connection of busbars in the wrong order, resulting in notes out of sequence, or with the NC and NO bars transposed, which will cause the particular block of eight keys to trigger a note quietly when released.

Finally, remember if using an alternative synthesizer module, that the E51O transmits only on MIDI channel 1. Most instruments, including the EMT–1O, default to this on power–up.

May I take this opportunity to wish you hours of enjoyment from your instrument, and acknowledge the creative input of Jo and Simon Brand, and the staff of Riverside Organ Studios.



The EMT–10

I chose the Yamaha EMT–10 as the synthesizer element of the prototype, mainly because of its realistic performance of a piano and reasonable second–hand price. It belongs to a family of MIDI equipment known as sound expanders, intended by the manufacturer to enhance the versatility of larger synthesizers and electronic pianos but which can just as well be used in their own right.

The EMT–10 offers 8–note polyphony and full velocity sensing over 12 basic voices; two pianos, two electric pianos, harpsichord, guitar, brass, strings, choir and three basses. The voices are not synthesized, they are sampled, and to CD quality. Nine variations on each voice are available, plus several other versatile features such as bass split, transpose, etc.

All this comes in a small black box measuring about 220 x 220 x 40mm, requiring just a 12V DC supply and a MIDI input.

Having said all that, the keyboard scanner will drive any piece of equipment accepting MIDI standard codes. If another device appeals, try it!

BUYLINES

The only components that may cause any difficulty are the E510 scanner IC and possibly the 74HC4501 analogue switches. These are both available from Maplin Electronics, PO Box 3, Rayleigh, Essex 556 8LR. Hopefully the remaining parts should cause no difficulty.

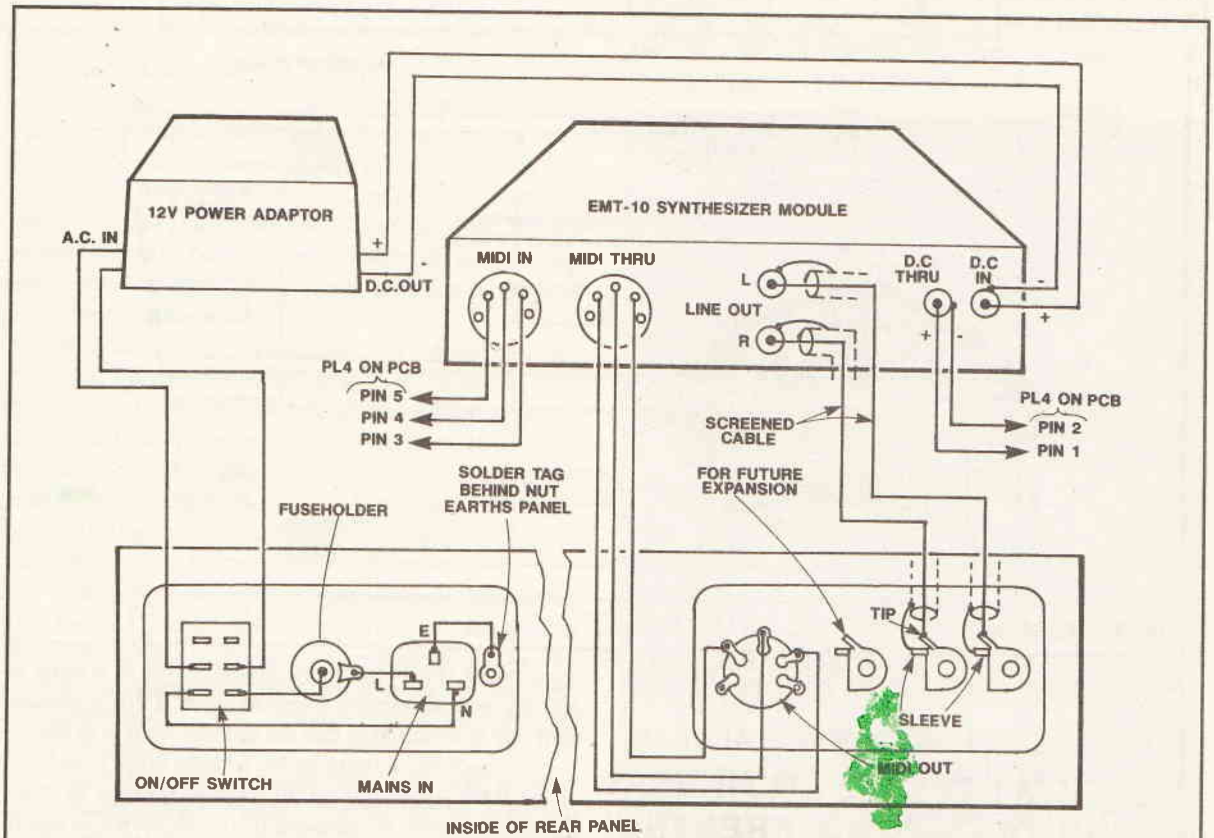
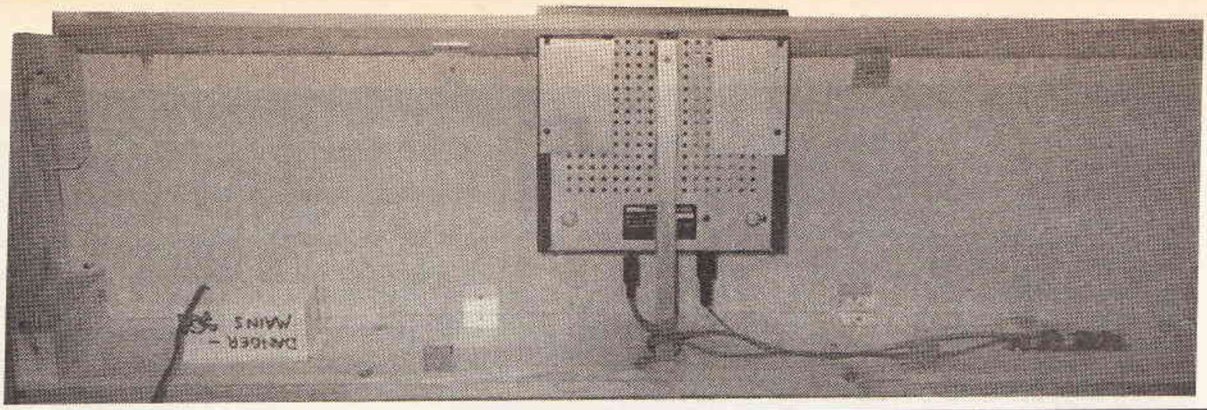


Fig.8 Internal Wiring



PARTS LIST

RESISTORS: (all 0.25W, 5%)

R1,2	200R (2 off)
R3,4	1M0 (2 off)

CAPACITORS:

C1,2	22p min. ceramic	(2 off)
C3–8	100n monolithic ceramic	(6 off)

SEMICONDUCTORS:

IC1	1510
IC2	74145
IC3–6	74HC4051
IC7	74HC00
REG1	7005

MISCELLANEOUS

XTAL1	4MHz crystal
PL1	8 way PCB plug
PL2,3	10 way PCB plug (2 off)
PL4	5 way PCB plug

8 way latch housing
10 way latch housing
Terminals for above housings
16–pin IC socket (6 off)
14–pin IC socket
Heatsink bracket for REGI
Wire, solder, etc

KEYBOARD ASSEMBLY AND BOX

RESISTORS

2k7	0.25W,5%	(18 off)

MISCELLANEOUS

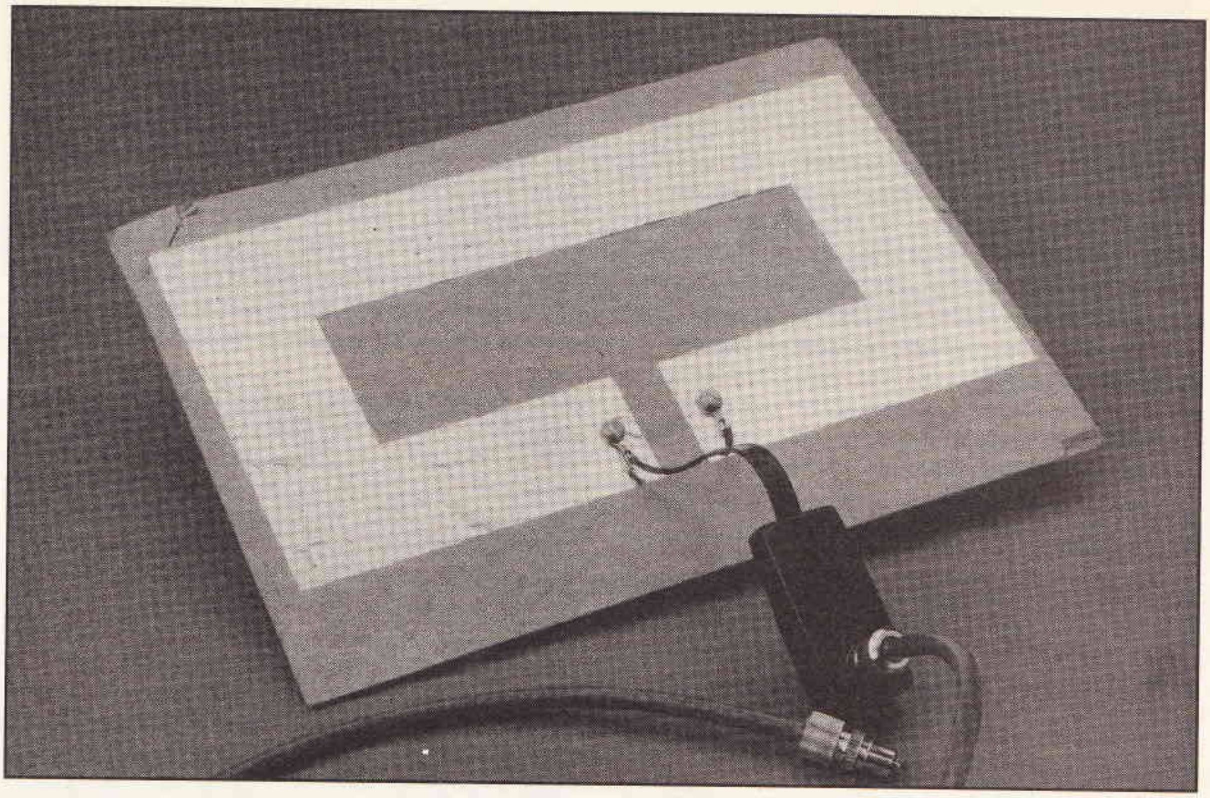
5 pin 1800 DIN plug (2 off)
5 pin 1800 DIN socket
0.25 inch jack socket (3 off)
Double pole mains rocker switch
Panel mounting mains input plug
Panel mounting fuseholder + 1A fuse
Phono plugs (2 off)
6–octave keyboard mechanism
Aluminium sheet
Paxolin block
2.5mm diameter brass rod
Veroboard
Cold–plated phosphor–bronze contact wire
Epoxy resin
Interconnecting wire
Small nuts, bolts, screws etc
Aluminium offcuts for brackets
Spray paint and lacquer
Letraset transfers

WOODWORK

Wood glue
Wood screws (assorted) Stain and varnish

14mm pine:	1070 x 143	
	465 x 143	(2 off)
10mm pine:	1070x 90	
	140 x 40	(2 off)
18mm pine:	1070 x 45	
9mm plywood:	1070 x 433	
	1070 x 255	

Offcuts for support blocks



Do It Yourself Television Antenna

A cheap way to produce a TV aerial by Stephen Waddington

A couple who live in a densely populated area of South London recenlly sought my opinion on television reception. Living in a second floor flat and unable to use a roof mounted antenna, they had resorted to the use of a five element indoor device which proved totally ineffective; reception was poor and ghosting continually occurred. This state of affairs prompted investigation.

The art of antenna design is steeped in mystery; an area avoided by all but the suitably adept academic. I offer this design as a neat solution realised after hours spent surrounded by textbooks, lengths of coaxial cable and strips of aluminium foil. The design described is two dimensional and requires mounting in a horizontal position facing the nearest television broadcast mast. Materials are inexpensive and the design allows opportunity for experimentation. The original prototype now sits on my living room wall attached to the rear of a Lowry print. Although results are not as superior as those for a high gain Yagi antenna, the design discussed provides adequate results and picture quality comparable with that of an outdoor antenna.

Before launching into the practical, a brief excursion must be made into the depths of theory, only then can any basic explanation of how the antenna works be realised. That said the theory presented is minimal and is provided solely as justification to the design. Literally hundreds of antenna blueprints exist – I do not propose to add these and have selected an old design, employing it in a new and slightly innovative fashion. For those interested in pursuing the subject further, numerous books deal with antenna theory and practical design in greater depth than is possible here.

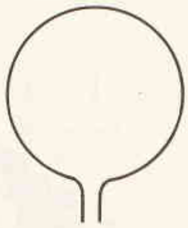


Fig.1a Circular Antenna Design

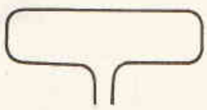


Fig.1b Half wave folded dipole design

Wave Formation

Radio and television waves form a small part of the electromagnetic spectrum and as such consist of two field components, one magnetic and the other electric. In general terms an electromagnetic wave is a form of radiation that transmits energy through space. Further examples

PROJECT

include light, heat, infra red and x–rays. The formation of an electromagnetic wave must be traced to its source. Any changing magnetic field has associated with it a changing electric field; perpendicular to each other, the two fields lie at right angles to the motion of the wave.

Any conducting body exposed to an electromagnetic wave will have an electro–motive force (EMF induced across it as a result of the varying magnetic field. Conversely, any conductor carrying alternating current will radiate electromagnetic waves. Anything from a coat–hanger to metal bookshelf might be used – rather boringly I decided to maintain convention. That said, a fellow colleague is convinced that a bedstead fixed in his loft maintains perfect radio reception.

The actual power received from a television or radio signal is very small. Equally; modern television receivers are so sensitive that fractions of a millivolt will suffice. Calculation dictates that we must induce a signal in the order of a hundredth of a microwatt. Miracles are not expected.

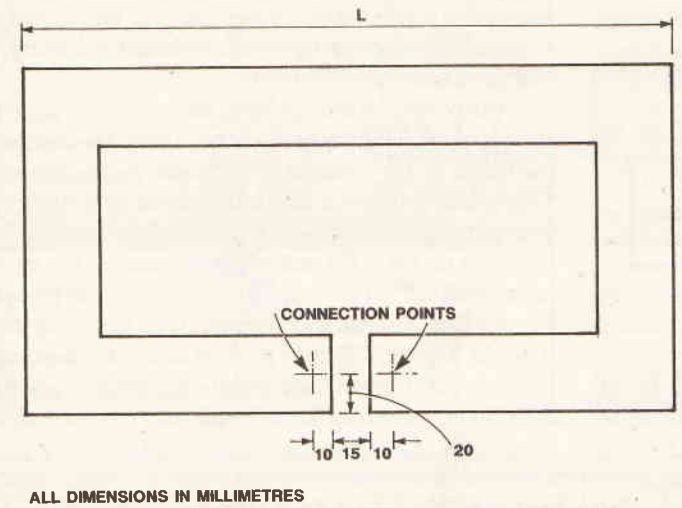


Fig.2 Half wave folded dipole design

The Basic Design

Initial investigation centred around existing television antennas. A design often used on portable television sets is that illustrated in Figure 1a. Predominantly circular in form, it is an obscure version of the half wave folded dipole. The more usual form is that illustrated in Figure 1b. Here, two conductors each of half a wave length are joined, with the lower conductor split at the centre to allow connection to a radio or television system. A resonant device, the antenna is designed so that the two main conductors match the frequency of the signal to be received. Although the design is spot tuned to a single channel, a tolerance of ± 10% allows for reception of neighbouring frequencies. Rather unusually for a television antenna, its impedance is 300Ω, a huge mismatch from the 75Ω impedance of a television antenna input.

Following experimentation I realised that the circular device could be improved to provide far superior results. Immediately the losses caused by the impedance mismatches could be eradicated, simply by introducing a matching transformer – a simple device which almost every manufacturer neglects. The impedance of the typical portable television antenna, such as that in Figure 1a, differs slightly from its companion in Figure 1b being 150Ω. Even so, the introduction of a matching transformer, doubles the efficiency of the antenna. Consideration of basic physics and electrical properties results in the modified design of a far superior nature, that of Figure 1b.

Mathematics

All electromagnetic waves are propagated through space at the velocity of light denoted by the symbol c. Though quoted for a vacuum, the value 2.99×10^8 ms^{-1}, may be considered the same for air. The wavelength of an electromagnetic wave travelling in free space is the distance between two points in the direction of propagation, where both the electric and magnetic fields repeat their values of intensity. Wavelength, conventionally denoted by the greek symbol lambda, frequency f and the velocity of light c are related to each other by the formula:

$$c=f\lambda$$

Consider for example Radio 3 which occupies amongst other frequencies, 1251kHz on the Medium Wave. The wavelength of the electromagnetic wave is calculated as:

$$\text{...Eqn 1} \quad \lambda = \frac{c}{f} = \frac{2.99 \times 10^8}{1251 \times 10^3} = 239\text{m}$$

A direct relationship results and by mathematical observation we soon realise that the higher the frequency the shorter the wavelength. Thus frequencies in the television broadcast band around 650MHz have shorter wave lengths than those of the Medium Wave radio band around 1000kHz (1MHz).

Domestic UHF television broadcasts are allocated a space within the electromagnetic spectrum referred to as Band VI and V. Carrier frequencies range between 450 and 850 MHz dependant on physical location within the United Kingdom. In this respect it would be necessary to quote several pages of data for every UHF Television broadcast station. Fortunately such information is easily obtained from either television broadcasters or any one of the numerous radio authorities or that exist within the United Kingdom. Relevant addresses are provided at the end of the article. Perhaps more convenient is the Maplin Catalogue which details broadcast frequencies throughout the country within the antenna section.

In general a domestic television antenna is required to receive eight different wavelengths to provide sound and vision for four channels. For example the transmitter for the original design is situated in Croydon and uses the following:

Station (MHz)	Vision Frequency (MHz)	Sound Frequency
BBCI	511.25	517.25
BBC2	567.25	573.25
ITV	487.25	493.25
Channel 4	543.25	549.25

TABLE 1: Domestic Television Frequency allocation.(Crystal Palace Transmitter)

Examination of Table 1 provides the precise limits of reception. In this instance the antenna is required to operate between 487.25 and 573.25MHz. The average frequency is selected and so the design is tailored to 530.25 MHz. Employing equation (1), the associated wavelength and therefore half wavelength is determined.

$$\lambda = \frac{c}{f} = \frac{2.99\mathrm{x}10^8}{530.25\mathrm{x}10^6} = 564\mathrm{m}$$

For a Half Wave $\frac{\lambda}{2} = 0.287\mathrm{m}$

The length of the antenna is thus realised. Readers must of course use data relevant to their area. Although not that different to the figures presented here, the final result may prove far from acceptable.

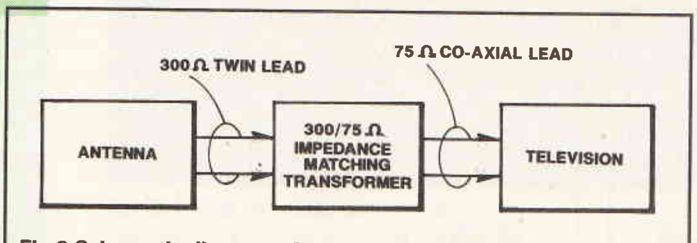


Fig.3 Schematic diagram of antenna system

The width of the conductors is not so critical, though a sense of proportion must be maintained and the width should be small compared with a half wave length. Equally a large cross sectional conduction area is important and a compromise must be attained. Figure 2 shows a plan of the antenna with actual measurements quoted in millimetres. Although the half wavelength *l*, is specific to each antenna,other measurements are not subject to regional variation and should be adhered too.

Impedance Matching

As explained earlier, a matching transformer is employed to overcome the impedance mismatch. A schematic diagram is shown in Figure 3. The antenna is connected by a twin 300Ω cable, to the 300Ω side of the matching transformer. The output from the transformer is then connected by a 75Ω coaxial cable to the antenna input of a television receiver.

Skin Effect

Domestic television antennas are often constructed from metal tubing as opposed to solid rods or aluminium foil as proposed in this instance. Such design seems inappropriate, since basic physics tells that solid conductors are the most efficient. The reasoning is easily explained; it is only direct current that flows uniformly through the entire cross section of a conductor. At high frequencies, current flow is restricted to a narrow skin around the edge of a conductor. The higher the frequency of the current, the thinner the skin becomes and so rather appropriately the phenomenon is referred to as the 'skin effect'. Figure 4 provides illustration. The current loss introduced by the use of tubing as opposed to solid conductors is minimal; additionally cost and physical weight are reduced.

The use of aluminium foil is not as ridiculous as might of first appeared. For UHF purposes, conduction will take place solely within a few micrometers of the surface. Whilst aluminium sheet would form a more permanent device, foil is fine for experimental and prototype purposes.

Construction

Construction is simple but as with all projects of this nature, the Blue Peter Code of Practise should be adhered to; 'cut with care!' The antenna is mounted on a board – thick card, hardboard or plywood will suffice. The ideal method of construction is then to trace the antenna design onto a piece of foil and then cut and secure to the board. A light adhesive such as PrittStick will provide adequate cohesion. Attempts to cut and paste each element separately should be avoided. Complication would arise when attempting to join the pieces together – the glue would act as an insulator preventing electrical contact, rendering the antenna useless.

Connection to the matching transformer is made by the use of two 6mm mushroom–head bolts fixed through the edges of the antenna in the positions illustrated in Figure 2. Provision for the bolts is made by drilling two holes through the antenna and supporting board. Spade terminals on the 300R end of the transfonner slot neatly underneath each of the bolts, which once secured between two washers, provide good electrical connection as illustrated in Figure 5. Should you opt to use a transformer which does not have a fixed cable link, a suitable length of twin 300R cable should be employed. Crimped spade connectors secured to each end would ensure a good electrical connection though the constructer may equally opt to secure the bare wires around each terminal.

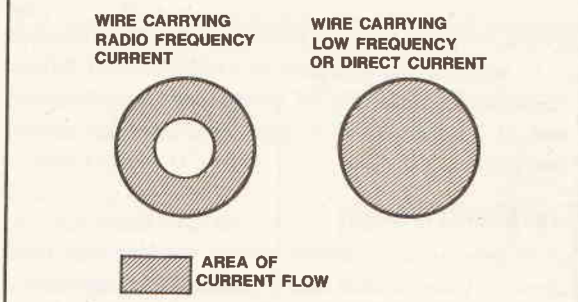


Fig.4 Skin effect

A co–axial lead is required to link the transformer to the antenna input of a television set. Two metres of coaxial cable should be sufficient – coaxial plugs should then be secured to each end allowing for connection between the matching transformer and a television set. Once completed the antenna may be plugged into a television set. The device is vertically polarized and is directional; to obtain the best results it should be mounted vertically and pointed in the direction of the broadcasting transmitter. Using the two channels at the extremities of the original calculations, orientate the antenna to give the best sound and picture quality. No formal proceedure can be provided for set up; a trial and error method is suggested.

Further Work

The half wave folded dipole design is easily adapted for VHF reception and provides excellent results for stereo FM. The sole change required is to the antenna's length – this must be changed to suit the frequencies concerned. Feel free to experiment; the beauty of this type of construction is that it is cheap and lends itself to trial and modification.

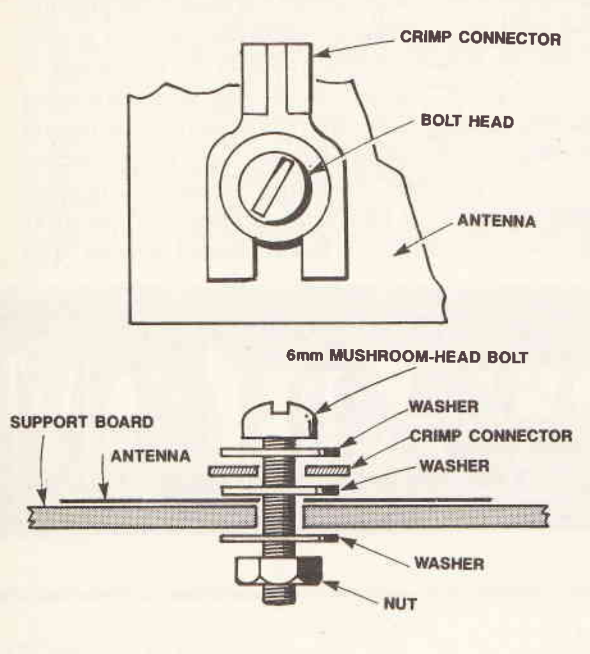


Fig.5 Connection to Aluminium foil

PARTS LIST

1 Piece 350 x 250 mm hardboard or plywood
1 Roll of aluminum foil
1 2m 75R coaxial cable
2 co–axial plug connectors
1 75R/300R Impedance matching transformer
2 6mm Mushroom bolts
6 6mm washers I 200 mm length of 300R twin cable (see text)
4 Spade connectors (see text)

BUYLINES

All material and components should be eaily obtained; the matching transformer is available from either Maplin Electronics or Tandy. Approximate Cost £3.50

Addresses

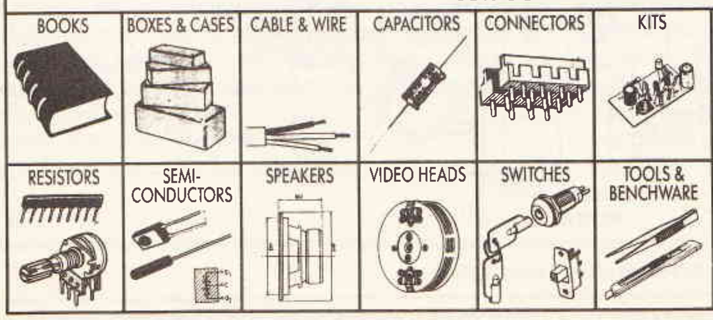
Information on frequency and channel allocation in specific areas may be obtained from:

BBC Engineering Information
White City, 201 Wood Lane, London. W12 7TS

The Radio Authority
Engineering Information, Waterloo Bridge House, London.

Independent Television Commission
Engineering Information, Crawley Court, Winchester, Hants.

The Radiocommunication Agency
Waterloo Bridge House, London. SE1 8UA



Anniversary AutoMa

Mike Meechan looks at four ways of squeezing the best performance per pound from the mic amp.

5

Yet another month has passed and we are again ready, poised to boldly go, split some more infinitives and make further inroads into the workings of the largest mixing console yet offered for publication within the electronics periodicals. In the past few months, we have progressed as far as the microphone/line input amplification stage, which, in the global scheme of things and after so much text, seems somewhat distant from the Master module. Readers intrigued by this apparently less than prodigious pace are referred to the outline strategy listed in part 4. To very, very briefly recap, I said that to publish further overlays or constructional details at this point in the proceedings would be tantamount to some irresponsibility on my part. It would be akin to me – in my alter ego guise of used car/timeshare salesman – trying to sell you a used car on the basis of a visual inspection by you of a shiny wheel from the notional and unseen vehicle! It is much better policy to wait until more information about the console is in your possession before tempting you with the sweetmeats of overlays and PCB layouts. In this way, all of the features and shortcomings of the mixer can be laid bare and open to scrutiny over this period and, we hope, all prospective customers will then be satisfied before contemplating spending any largish sums of the old green stuff. We at ETI have no wish for three or four completed modules to be

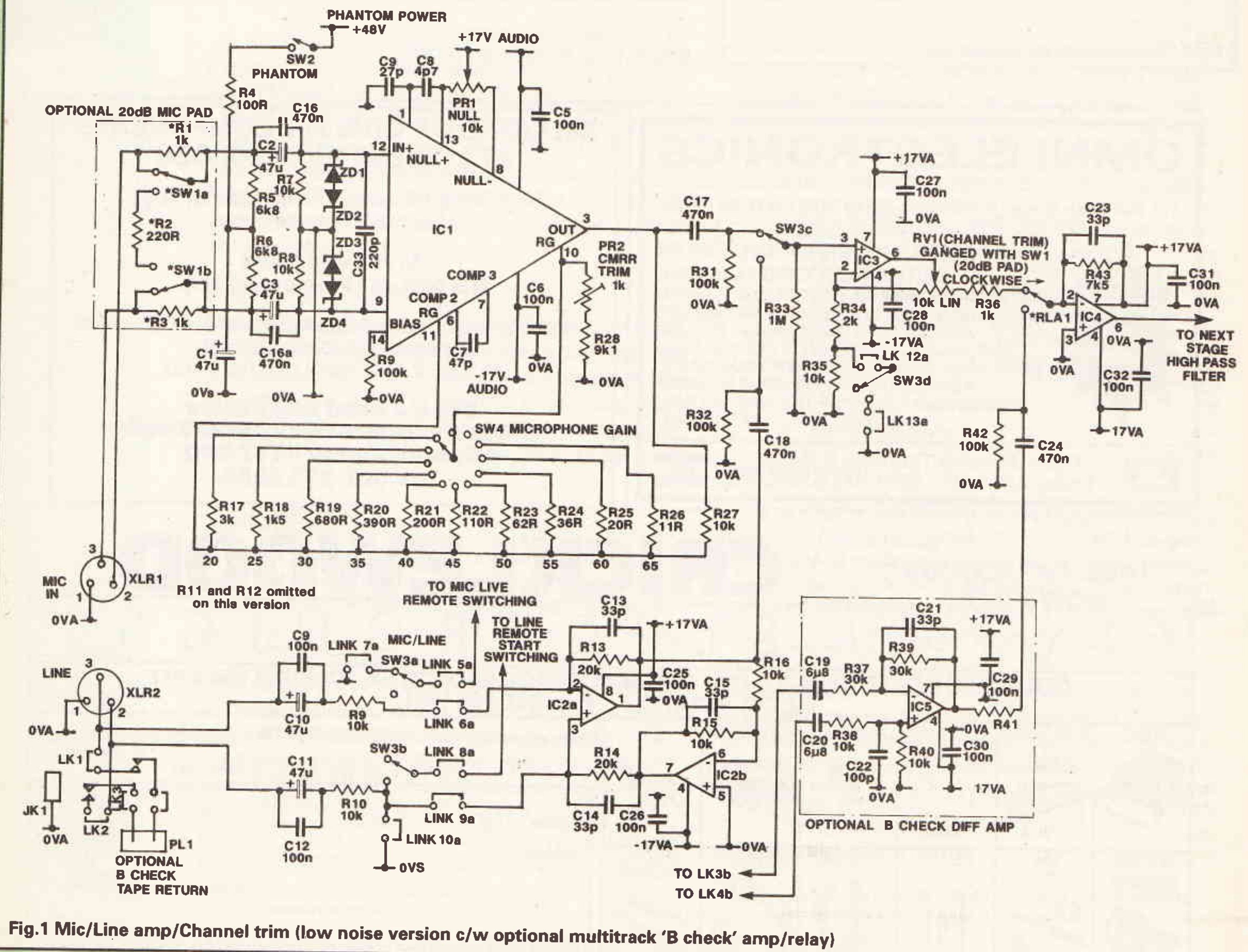


Fig.1 Mic/Line amp/Channel trim (low noise version c/w optional multitrack 'B check' amp/relay)

confined to an under–stairs cupboard because of improprieties on our part as regards costing analysis or incomplete guide prices – in simple terms, the damn thing was too dear to take to completion!

Enough of the hype! In this and following issues we move into a very important area of the console, that of the Level Architecture. I had hoped to delve into the mysteries of equalisation section – EQ – where creative or corrective tonal spectrum changes can be made to the input signal. Regrettably, text limits mean that this will be held over until Part 6. On a positive note, this means that we can explore more thoroughly and from first principles both passive and active filter sections, the relationship which these have to high quality audio in general, and to mixing consoles specifically, the various ways of implementing the EQ functions with active electronics, and the AutoMate way of doing the business.

Staying on the Right Side of the Law

Staying with the more immediate concern of this month's text, we must first complete some unfinished business in the shape of the Mic/Line input stage and gain control.

Referring to the modified input stage schematic of Figure 1 –that published last month was the general form – and using already published information and circuit diagrams, we should remember that differential amplifiers, unlike most others encountered, operate at high gain when the value of gain setting resistor between the two transistors is lowest. For a good control law, we therefore require an anti–log pot, as linear, or worse still, log law, would cramp all of the gain variation into a very small part of the control movement. This type of gain control is used on many mixer input stages and provides good operational and variable control of gain. However, I was unable to track down (no pun intended) any readily available supplies of antilog pots and so opted for a precision switched gain control cascaded to an op–amp stage with gain on this second stage fully variable between zero and plus 30dB. The switched gain control uses a 12–way, adjustable–stop PCB–mount-

ate Mixer

HOW IT WORKS

Low Noise Mic/Line/Channel Trim section

Microphone inputs (which ideally should be balanced) are inputted to the console via XLRl, a 3–pin style socket. SW1a and b are part of the RV1 'channel trim' pot and are opened and closed by pushing or pulling the pot shaft. This switch introduces 20dB of attenuation via resistors R1, R2 and R3. R5 and R6 balance the 48V phantom supply across the two legs. The supply is switched on or off via the panel–mounted SW2 with C1 and C2 (and optional electrolytic bypasses C16 and C17) providing the necessary DC blocking of the input whilst ZD1 to 4 provide transient over–voltage protection of the input amplifier should the mic be plugged in with the phantom supply on. C25 compensates the input stage current regulator while C7 and C8 provide compensation for the overall amplifier, these latter two chosen in relation to the value of Rbias needed. Rbias, R9 programmes the input stage for a given impedance and so optimises noise performance for the given input source resistance of 200R. This feature of the input amplifier means that the 2015 can provide close–to–optimum performance from source impedances up to 4k, noise being within 1dB of the theoretical limit between 500R and 2.5k. The bias resistor value has been chosen to optimise performance with a balanced transducer - noise worsens with unbalanced inputs because of the uncorrelated nature of the noise currrents of the 2015 and the bias resistor value may be altered to compensate for this.

There are two trim controls to be adjusted with respect to this IC. The 'null' control is an offset trim used to null out control feedthrough. Since the output offset at low gains is determined by the matching of the feedback resistors and at high gain by the matching of the input transistors, a click or thump may be audible when gain settings are rapidly changed. This is due to a shift in output offset and is eliminated by adjusting the null control for equal offsets at high and low gainsettings. The other is a CMRR trim and is adjusted for minimum output with a high level common mode input. As the untrimmed performance of the 2015 in both of these aspects is excellent, these trims may be omitted if so desired. Gain is adjusted with the rotary switch of SW4, this placing one of the gain setting resistors, Rl7 to R26 into circuit. The attached table shows the dB gain error for each of the settings. Capacitor C17 blocks any offset DC from the next stage so that there are no audible splats when the Mic/Line switch position is changed. Its value is chosen to give a pole at around 3.5Hz, low enough in value that succesive RC networks of similar value do not cause excessive, cumulative phase shift and colouration of in–band audio. IC3a and b are the 'Channel Trim' shared gain stages which allow a further 30dB of gain for Line signals and 20dB for microphones. SW3d includes resistor R35 in part of the gain–determining network when 'Mic' is selected. This reduces gain by about 10dB since only 15 or 20dB of further gain will be required after initial amplification by the mic pre–amp. Links 5 and 6 select this function. 'Line' gain can be altered by omitting or changing resistor values in this area. This "trim" stage is arranged to "lose" around 4dB of gain so that the 0dBu operating level is maintained after the High Pass Filter section (which gives around 4dB of gain). Headroom is not compromised by this action. Resistor R33 which grounds the non–inverting input is another 'splat eliminating' component.

Line Inputs enter via XLR2, another 3 pin socket or JK1, a switched stereo 1/4" jack socket. The socket switching contacts can be arranged via Lk1 to 4 to appear in parallel with or normalled through the Line Input XLR. This is of most use for tape returns in a multitrack situation. It is also the reason behind the inclusion of PL1, the optional 'B Check' differential amplifier and relay RLA1. I propose not to explain these components further at this point. From the input sockets, line inputs travel to IC2a and b Superbal differential input amplifier. Another optional 20dB pad can be fitted at the input to this amplifier if connection to high level, transient sources is envisaged. The optional pa1d provides the necessary attenuation. Gain in the trim section should then be adjusted accordingly. The Superbal operation was described in some detail last month and reference is made to this. The output of this section is DC decoupled and tied to ground via high–value R32 before appearing on the other contact of SW3c. Switching clicks are thus minimised. It should be noted that signal switching through SW3a and SW3b is disabled by fitting Links 5 to 10 in position 'a'. This then frees the switch contacts – which were redundant in this application anyway – to control 'Mic Live' and 'Remote Start' functions in the 'Mic' and 'Line' positions respectively. In all configurations except the budget one, this function is available and all links are fitted in position 'a'.

ing rotary switch adjusted for ten stops. Coupled with close–tolerance gain setting resistors, this facilitates precision adjustment of gain from 65 to 25dB in 5dB steps. With the cascaded op–amp stage, this affords us more than 80dB of microphone gain should we so require it.

Padding out the Mic Input

The IC employed in the flagship mic amp design of the console – Precision Monolithic's SSM 2015 mic pre–amp – in common with similar transistor differential amp/op–amp microphone amplifier configurations, suffers from an increase in distortion at low gain settings. To counteract this, there is an optional 20dB pad – attenuator – which can be fitted to the input and switched in or out as required from pot–operated push on–push off switch. Many modern–day designs of mic–input stage place great emphasis, at least in the advertising blurb, that they are "transformerless inputs, using a padless approach and special circuitry to allow single operational control of between 0 and 70dB of gain". Commendable and clever though this approach might be – and I was unable to deduce how this is done – it doesn't explain how, on the occasions – rare as these might be – that a very–high level, peaking to +10dBu source is to be close mic'd, the input channel amplifiers remain able to

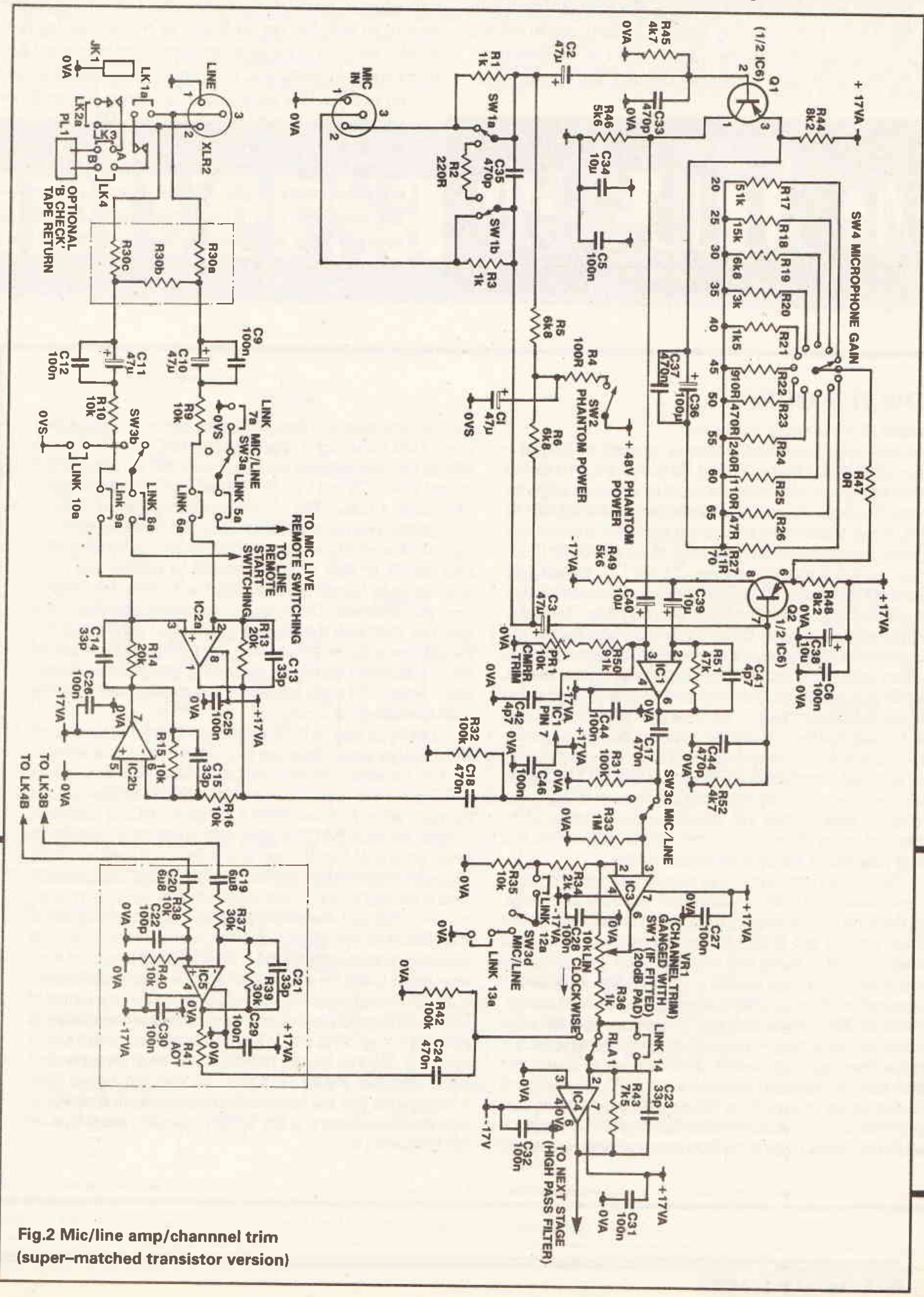


Fig.2 Mic/line amp/channel trim (super–matched transistor version)

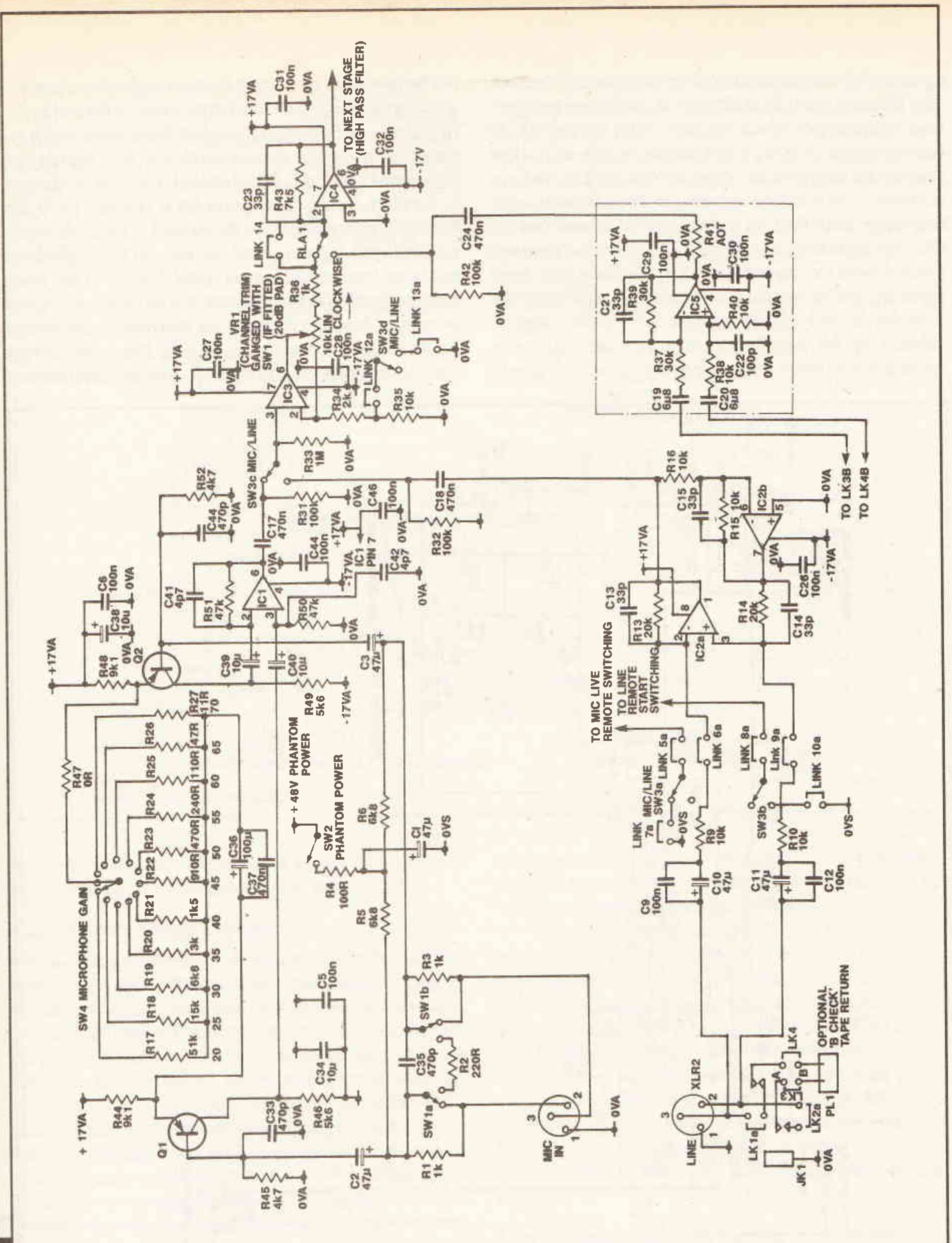


Fig.3 Mic/line amp/channel trim (Budget discrete version)

HOW IT WORKS

Mid–Price and Budget Low Noise Mic Pre–amp

Super–matched transistors Q1 and Q2 in IC6 (or their discrete counterparts, Q1 and Q2 in the budget low noise version) are operated open–loop at the input stage to create a fully floating, differential input. R44 and R49 set the collector current of the input stage, with values optimised to give the amplifier a low noise figure as per the main text. Resistors R45 and R52 provide the necessary bias for the two transistors. The gain determining elements of SW4 and R17 to R27 work as in the low–noise mic amp, providing a resistive degeneration which gives the amplifier stable gain over a wide range. R45 and R52 and 45 associated capacitors C33 and C45 also provide HF roll–off and protection from RF.

In operation, the input transistors convert a differential input voltage to a differential output current at their collectors. This is fed to a standard differential operational amplifier to obtain a single ended output voltage. Using a differential rather than single–ended output from the transistor stage means that CMRR is improved by an order of magnitude and good overall rejection can be achieved using just 1% tolerance resistors. Gain is set by the ratio of the selected gain determining resistor to the sum of R50 and R51. The line input operates exactly as in the Low Noise version.

operate with normal headroom considerations. This is a very practical example of the type of conflicting performance compromises which we must suffer throughout the console design. Sticking with a normal mic pre-amp at low gains means increased distortion; introducing a pad reduces common mode rejection and using an alternative means of first-stage amplification ie the Superbal means that the OSI–optimum source impedance–condition isn't met and residual noise increases as a result. Under most high-level operating conditions, the first-mentioned is probably the least objectionable of the three since only CMRR is slightly affected by the introduction of a pad, although extra resistance does make for an increase in noise. This should not be too much of a worry because high level signals – which were the cause of the introduction of the pad in the first place – will tend to mask low level noise which is, naturally, of greatest nuisance with low level signals ie a high signal noise ratio is maintained. Personal preference is therefore for the first option but it is really up to the individual constructor as to the method or methods which are employed. If not required, the pad can be switched out and is no detriment to normal, good, low-level and low-noise performance. In any event, it is an option which can be omitted from the design at the individual's discretion. As an aside, it is interesting to note that some mixing console designers use a straight op-amp micamplification

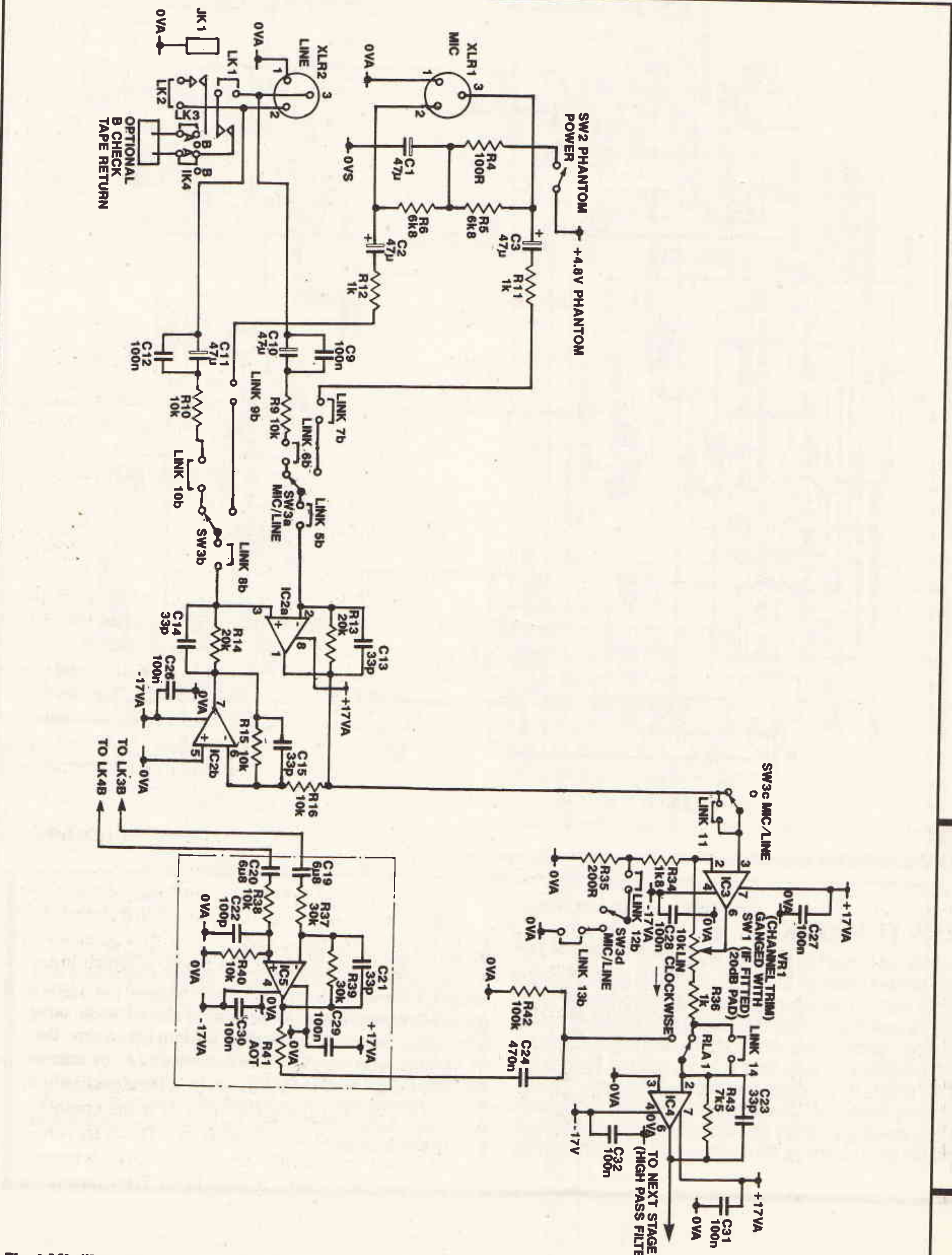


Fig.4 Mic/line amp/channel trim (Budget version)

PROJECT

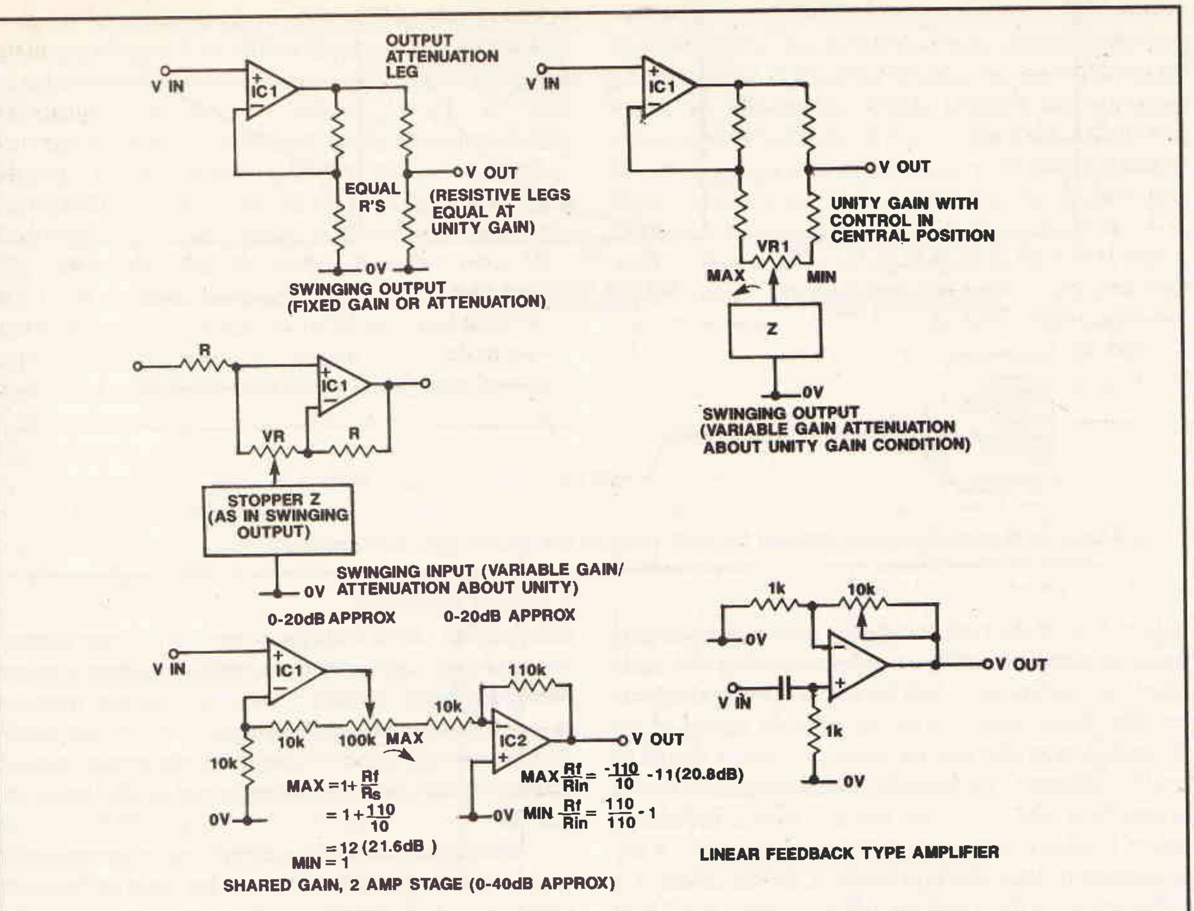


Fig.5 Swinging Inputs and outputs.

technique at all levels of mic input signal, thus seriously compromising overall console noise performance.

Line Input Stage Design

The Mic/Line pushbutton switches the input to the Superbal unbalancing and buffering stage – the mic amp stage is bypassed when Line is selected and the line input is subject to less amplification. Gain for a Line level input is restricted. About 0 to 30dB seemed an appropriate range if domestic line levels are to be readily accomodated. These values are easily altered and I shall discuss how later in this part of the article. Again, its the case of the greatest good. If high level transient line inputs are likely to be encountered – which is slightly improbable since the connecting circuitry will be powered from supply rails similar in magnitude to those within the mixer and headroom should not be a problem – an input attenuator can be fitted. The gain range of the "Trim" should be adjusted accordingly.

The manner adopted to provide this supplementary, second–stage gain is worthy of note in that it uses no attenuators in series with the input signal path. Many commercial designs use a variable 30 or 40dB pad – attenuator – coupled to a fixed gain amplifier. For 15 dB of gain, for example, the input would be attenuated to –30 and then amplified by 45dB, with all of the dire aural consequences which this brings. In the circuit offered here, input headroom, signal–to–noise and distortion are optimised for any given input signal magnitude within the range of the amplifier and there is no unnecessary attenuation of signal.

HOW IT WORKS

Budget Mic/Line Input stage.

Again, as in the previous example, many components within the network are similar to those of the low noise option and where this is so, I shall not repeat the circuit description. The SSM 2015 and associated gain setting switch are omitted in this version and the mic applied, via its own 1k impedance network, and through SW3a and SW3b and associated links in the 'b' position, to the Superbal. This imparts 20dB of gain to the signal which is then passed directly to IC3, 4. This section now acts as the main amplifier for the mic and so conditions on SW3d are reversed – links in 'b' position – with the switch introducing another 30dB of gain into the network when 'Mic' is selected, total gain range for the mic within this stage being 0 to 60dB. With this particular version of the budget amp, there is again the option of a gain pot–operated 20dB Mic Pad. The line input is as for the low–noise version.

Love, Peace and Swinging Inputs and Outputs

Whenever gain is made up, it is important that any noise is due, in the main, to the gain stage (which, hopefully, has been optimised to reduce noise as much as possible). In this way, all subsequent noise contributions, which, in any case should be small, will thus be masked. At no place in the gain swing should it be necessary to attenuate unwanted gain. This is of particular importance where the gain control is at its minimum setting, since any attenuation, (and hence headroom lost) early in the signal paths effectively lost from that point onwards in the console audio pathway and can never be recreated. This is the concern of so–called 'level architecture'. It is in just this respect that the circuitry of the Channel Input Trim control scores

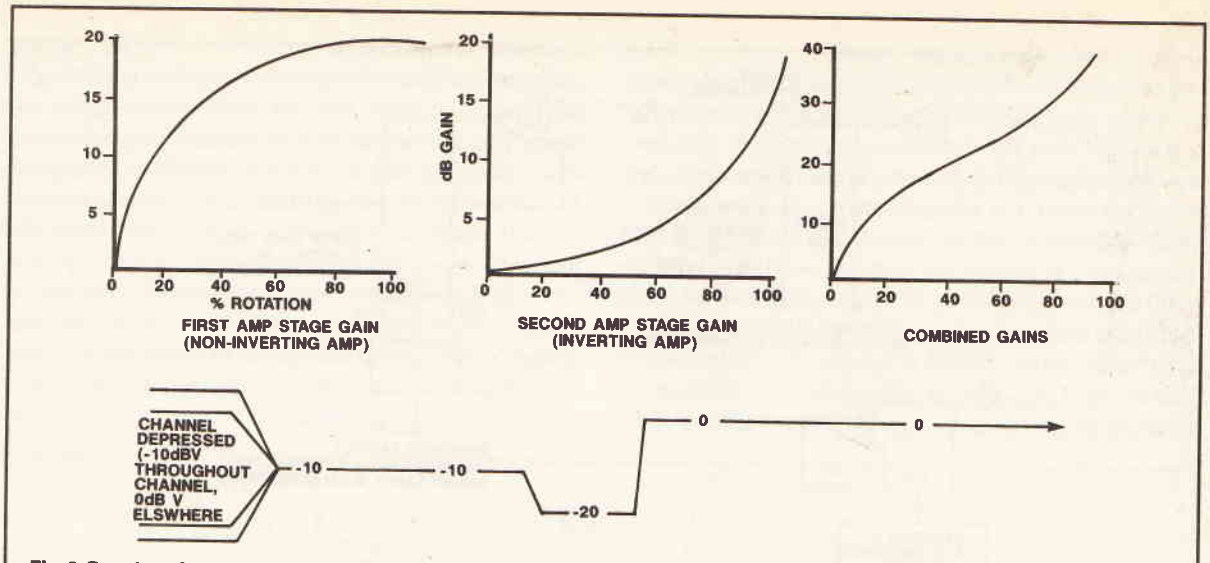


Fig.6 Graphs of gain Vs control rotation for each stage of the shared–gain configuration

highly. It is of the type commonly known as 'swinging input', a quaint–sounding name for something which has absolutely nothing to do with the sixties or hippy chicks etc etc. For those familiar with the network layout of the Baxandall–type tone control, some similarities should be readily apparent. Technically, the Baxandall achieves symmetrical boost and cut using a source impedance versus feedback impedance ratiometric approach. A development of this which achieves a similar effect is a circuit known as the 'swinging, output' control. It will help in our explanation of the swinging input if we first discuss the swinging output since the former is an enhanced–performance derivative of the latter!

Refer to Figure 5. The diagrams show typical swinging input and output controls. In the first, the gain control is enclosed within the feedback leg of a non–inverting amplifier, this configuration meaning that a buffer amp or low impedance source – needed for an inverting configuration – is obviated here. Unfortunately, the output impedance of this network varies with the setting of the gain control, necessitating a high subsequent load impedance or output buffer amplifier. Serious drawbacks arise and manifest themselves as problems with phase margin and consequent instability, loss of headroom and departure from and serious modification of a workable and useable gain control law.

Unity gain results when attenuation in the output leg is equivalent to that in the feedback leg since the feedback ratio exactly nulls the effect of the output attenuation. The maximum gain condition results when the feedback leg is shortened with respect to ground and the output attenuator is lengthened. This means that loop gain of the amplifier is at a maximum while an almost completely unattenuated signal is available at the network output terminal. At minimum, the feedback resistor is shortened (reducing loop gain) while the attenuator leg is lengthened, reducing output signal amplitude. A small resistor connected between the pot wiper and ground controls the gain range. Using

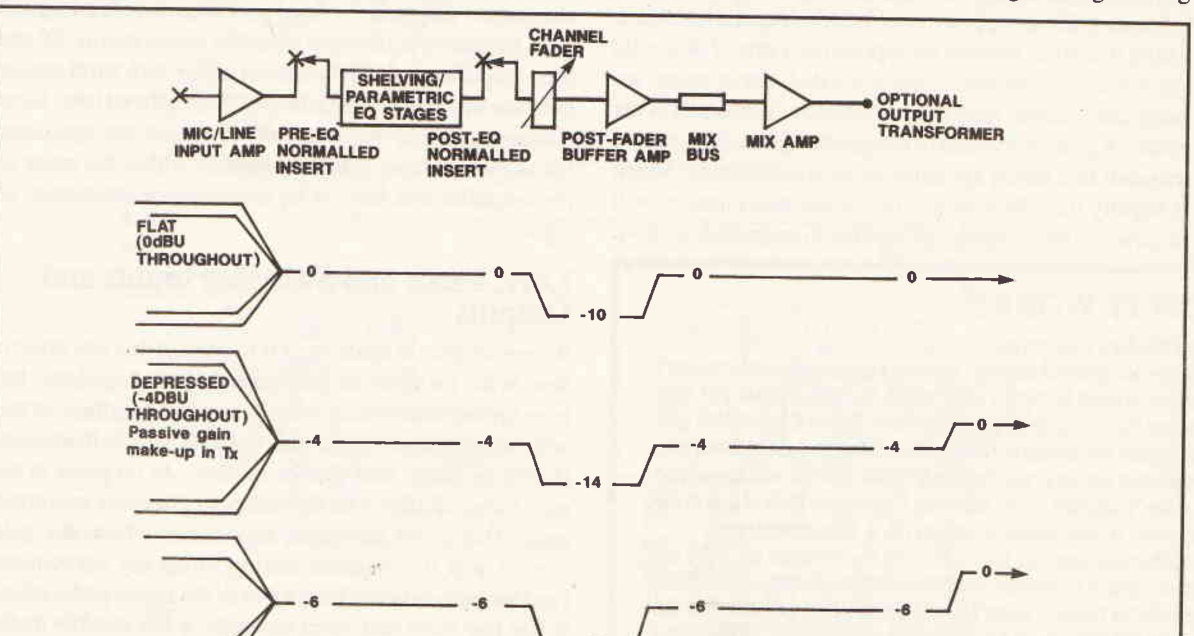


Fig.7 Level architecture in mixing desks

PROJECT

complex impedances such as capacitors and inductors achieves frequency–conscious boost and cut and we shall explore this avenue of development in greater depth in the EQ section.

We can improve on this simple and elegant circuit by making it into a swinging input. With this configuration – shown also in Figure 5, the output attenuation network is transferred to the input while the feedback network remains as it was in the swinging output type. At the minimum gain setting, the input signal is subjected to maximum attenuation while the amplifier –with its long feedback leg – delivers only a small amount of gain. At maximum boost, these conditions are, of course reversed, with minimum input attenuation and a short feedback component providing high gain.

Further improvements are made in the way that the circuit performs with respect to noise. At unity gain, when input attenuation and amplification are exactly nulled, the loop has around 20dB of gain. At first sight this may seem to be somewhat noisier than the corresponding Baxandall type. However, all of the network resistance values can be made lower by an order of magnitude and so Johnson or thermal noise is significantly reduced with respect to the Baxandall layout. Moreover, any noise injected into the system (at unity gain) appears in equal quantities at both the inverting and non–inverting inputs ie common–mode noise at a differential input. Ideally, common mode signals should not appear at the output. One notable drawback is that at unity gain, there is at first attenuation of the signal but ignored at any reasonable settings of the gain control. Headroom is preserved because at no setting of the gain control is any attenuation introduced and, as an added bonus, the distribution of gain between the two stages means that the gain control law is as ideal as we could wish for. From the graphs of gain versus pot rotation shown in Figure 6, we can see that around the centre of the control –where most adjustments will probably be made – we have a very happy state of affairs in that dB change in gain is almost linear with respect to control rotation. This dB change becomes rather more cramped at extreme settings of the control, but it is a more than acceptable compromise.

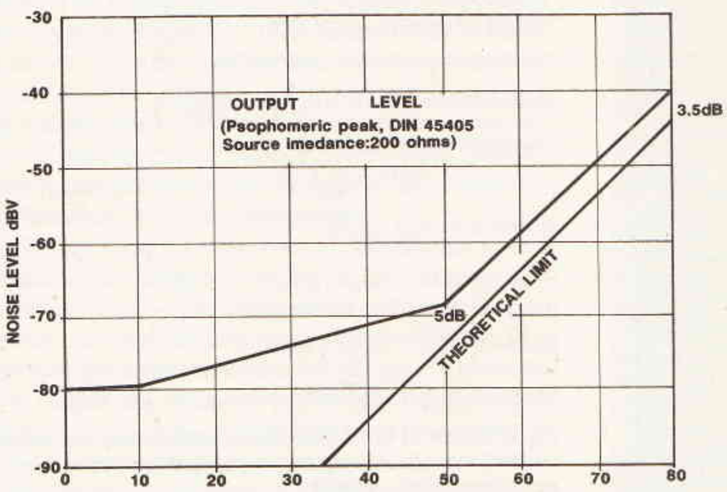


Fig.8 Graph of noise level limits for mic amplifiers

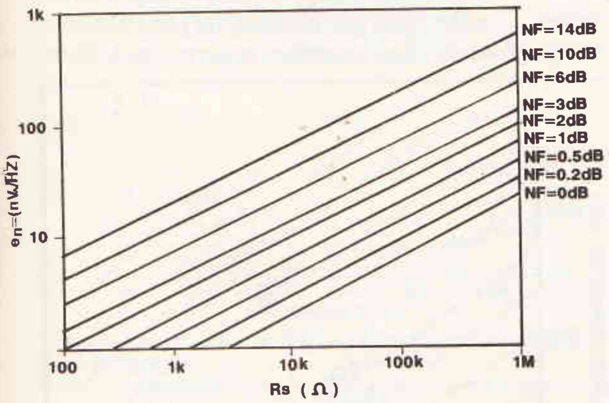


Fig.9 Graph of effective noise voltage Vs noise figure and source resistance

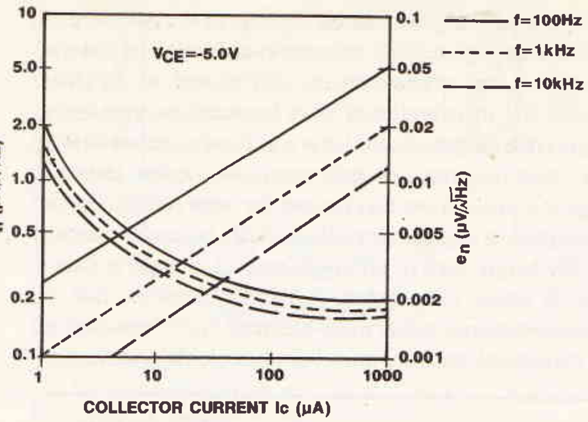
and then reamplification to the original level. This is slightly detrimental to ultimate noise performance and it is in this respect that this next circuit scores very highly.

A further evolution of the swinging input type – which can achieve quite excellent performance at very high gains – is shown at the bottom of Figure 5. With this configuration, noise criteria are always met since the first gain block always provides more than ample amplification, thus allowing its noise to swamp the second stage. Most importantly, this feature holds true at the minimum gain setting. This is immaterial in respect of noise – but not headroom – since any front–end noise added will be at the same level as the console noise floor which has been determined already. Resistor values around the amplifier are the primary determinant of the noise performance of the amplifier, so this performance should be able to be all

It is this type which features here.

Although the SSM 2015 does everything that we could wish for, and at a very reasonable price, I thought that for some applications, the cost of the IC, relatively inexpensive though it is, might be off–putting to prospective constructors and so set about devising another from first principles. From a designer's point of view, it is sometimes quite satisfying – though only sometimes – to forsake the all–singin', all–dancin' IC method of curing a problem in a circuit and go back to basics to design from scratch using only the manufacturers' data sheets and application notes for inspiration and guidance. It also avoids any accusations of plagiarism of established commercial designs although all commercial mic amps do, in the main, seem to be a variation on a theme. We've already mentioned instrumentation amplifiers and the methods of implementation. The mic amp can therefore be considered as a special type of one of these. Even before we consider source impedance and optimising of the input device for low noise, the use of a multi op–amp style instrumentation amplifier means very closely matched resistances for good CMRR, an attribute which is, of course, of great–importance in any front end stage.

The high precision and ultimately costly resistors required mean that we can therefore abandon this type. What we can do is to make use of a transistor/op–amp hybrid which, with accurately–matched transistors, can offer CMRR's in the region of 100dB – dependent upon transistor matching – with only 1% tolerance resistors and a trim control. This type, however, is not without its drawbacks, notably in the way that gain linearity and



Equation 1; $e^2_{R(}source) = 4kTRs\ V^2/Hz$

Equation 2; $e^2\ (amplifier) = e^2_n + (inRs)^2\ V^2/Hz$

Equation 3; $\text{Noise Figure} = 10Log_{10}\left(1 + \frac{e^2n+(inRs)^2}{4kTRs}\right)$

Equation 4; $e^2_n = 4KTr_{bb} + 2qIcre^2$

$= 4KTr_{bb} + \frac{2(KT)^2}{qIc}\ V^2/Hz$

Equation 5; $in = \sqrt{\frac{2q\ Ic}{h_{FE}}}\ A/\sqrt{Hz}$

Equation 6; $e_N^2 = e_N^2 + (in^2r_s^2) + 4KTrs$
(total noise = Equation 1 + Equation 2)

Equation 7; $\frac{2K^2T^2}{qIc} - \frac{2qIcrs^2}{h_{FE}} + 4KTrs$

Equation 8; $\frac{d(Vn^2)}{d(Ic)} = \frac{-2K^2T^2}{qIc^2} + \frac{2qrs^2}{h_{FE}} = 0\quad I_{Captimum} = \frac{KT}{9} \times \frac{\sqrt{h_{FE}}}{rs}$

Fig.10 Graph of RMS input noise voltage(en) and noise current Vs collector current for typical low noise bi–polar transistor – BC560

accuracy are not as good as with the multiple op–amp type. I used the SSM2220, a super–matched PNP transistor pair manufactured by our friends at Precision Monolithics Inc. and cheaper by about £5.00 than the similar NPN–type LM394. A further and even cheaper version uses discrete low noise transistors adequately suited to the purpose of low level, low–noise amplification. We have already discussed in Part 1 an amplifier noise model and looked at the contributing sources of noise in amplifier design. These were e_n and i_n. These are just one of a number of inter–related factors which must be considered in any low–noise, low–distortion, front–end design. For a thorough understanding of the approach to be taken as we follow the yellow–brick–road discrete avenue of design, some detailed explanation of the terms which will appear at regular intervals throughout the text should first be given.

Noise Figure

In any low level amplifier, noise is the single most important design parameter and performance–limiting factor, and since resistors – both biasing and load – are an integral part of amplifier design and application, these will contribute noise because of thermal activity. When the input signal is amplified, so, unfortunately, is the noise and if the ratio of signal power to noise power is the same at the output of the amplifier as at the input, it is said to be 'noiseless'. This is, of course, a practical impossibility and real amplifiers degrade this ratio by adding noise of their own. The degree of this noise impairment is known as the noise figure of the amplifier and is expressed as a ratio in the form;

$$NF = \frac{Sig_{IN} \times N_{IN}}{Sig_{OUT} \times N_{OUT}}$$

and noise figure in dB = 10 log (NF of the power ratio)

In our application, we can consider an amplifier with a noise figure of less than 3dB to be good and of less than 1dB NF to be excellent. Put simply, the noise figure is the amount by which the EI_N is higher than the thermal noise of a 200R resistor (or other specified value). EI_N is the noise measured at the amplifier output plus the amplifier gain when the input is terminated with a resistor of the nominal value of microphone impedance ie 200R. Although noise figure is easy to calculate and to manipulate within calculations, it is not an all–encompassing panacea and sight should not be lost of the fact that ultimately, it is the signal–to–noise ratio which is important. Choosing a high source resistance – where this is both possible and applicable – yields a very good noise figure, since e_s swamps e_n, but signal to noise is poor because of the inherent high level of noise. Quite simply, this means not fiddling with the value of the source resistance in an effort to make the noise figure better because it does so at the expense of making the amplifier noisier!

It is also important when comparing performance specification figures to compare like with like. This is not as elementary as it might first appear. Many commercial manufacturers of all types of consumer equipment – not just audio – are guilty of what might be termed 'specmanship'. Power amplifiers quoted as having "40 watts total peak music ouput power" (which equates to a useable 7W per channel RMS) are more common in the commercial marketplace than is healthy for good customer relations. The same hype is applied to the mic amp. If the

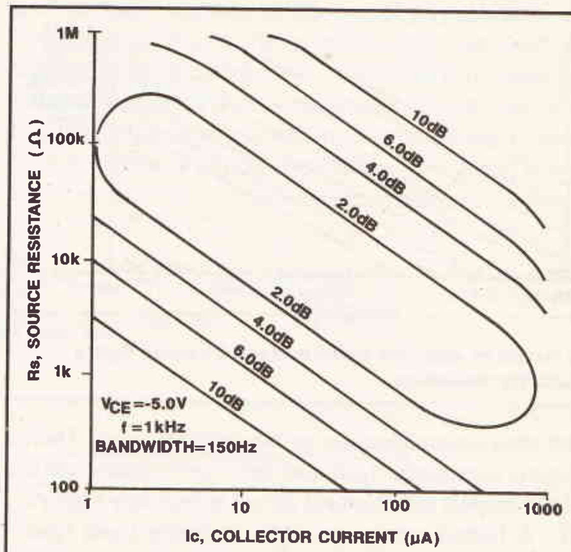


Fig.11 Graph of contours of constant narrow band noise figure for BC560C

noise specification of microphone amplifiers is to be meaningful, gain, terminating resistance, bandwidth, the noise measuring instrument, reference level and weighting networks – if any – must all be documented. The type of measuring instrument used is very important because of the way in which an RMS meter will yield a figure around 7 to 10dB better (quieter) when compared to an identical measurement taken using a peak measuring piece of apparatus such as a PPM. Quoting lower source resistances, higher reference levels or the use of weighting networks are all techniques used to cloud the spec somewhat and produce better figures. All measurements in this series are taken with a PPM referred to PPM4 (0dBu) and with 200R terminating resistance.

Your Terminated, Sucker?

Some mention should be made of the choice of 2k input impedance of the typical console mic. input. Normal textbook convention for maximum power transfer is for the source impedance to look into an equal value load impedance. Why not then arrange for the mic to look into a corresponding 200R load impedance? It is because it is the output voltage capability of the source which is of most interest to us here since an equal impedance termination would dissipate most of the hard won power and 6dB of signal level. From the point of view of noise, the source impedance now seen by the amplifier is half that of the mic alone (two equal Z's in parallel) so the Johnson noise is 3dB less. Although noise performance is not unduly impaired, it is better not to lose an immediate 6dB of signal level.

Furthermore, from the point of view of frequency response, the microphone, being predominantly inductive in character, has an impedance which rises with increasing frequency. Terminating with an equal value resistor would therefore mean that the complex impedance so created would constitute a first order low pass filter, rolling off much of the useful HF response. Having a source impedance around five times that of the capsule allows the microphone to behave as an unloaded voltage generator.

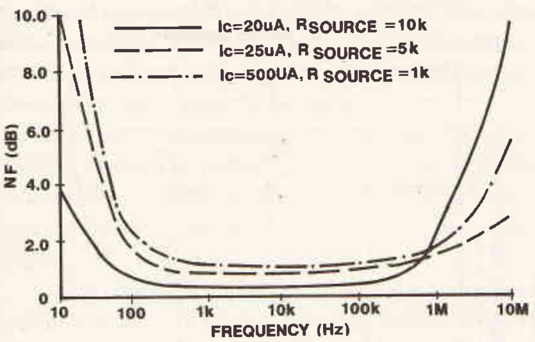


Fig.12 Graph of noise figure Vs frequency at diffent Ic for BC560C

Noise

Now that the concept of noise figure and measuring techniques are understood, we can look at E_{IN}. In addition to low distortion, the E_{IN} (which sets the lower limit on useful input signal level and maximum input signal level before overload) is probably the single most important parameter within the sphere of noise and its reduction. E_{IN}, or self–noise of –129.6dBu (with a 200R source) in a mic input is equivalent to an SPL of 26dB on a typical microphone. Room noise – typically around the 33dB mark – is likely, in many applications, to limit small signal capability. A typical pre–amp maximum signal handling capability of –20dBu before undesired increase in distortion and with an input sensitivityof 0.3mV for full output provides a total input signal range of 105dB. Input sensitivity or gain controls therefore set the best combination for the room or background noise and expected peak signal amplitudes.

This is where noise voltage and noise current enter the scene in earnest. Noise voltages and noise currents alter in magnitude and in ratio to one another, lower collector current giving rise to lower noise current – not inconceivably, since the current noise is due to minor random discontinuities in the device currents. The ratio between the two – noise impedance – can therefore be altered.

Given identical measuring conditions, resistors of different values will produce the same NOISE POWER, it is only the noise current and noise voltage ratios which change. This optimum value is the already documented optimum source impedance, the resistance value at which the device is optimally quiet for audio purposes. It is fortuitous that this OSI is also the value which coincides with that required for optimum device transfer characteristics ie there is good frequency versus phase linearity response and so the device will be stable at high frequency, high feedback–low gain configurations.

We can bring the OSI down to that of conventional dynamic microphones by reducing the ratio of the amplifier inherent voltage and current noise using transformers – oh–oh – or by paralleling identical input devices. This technique maintains the same noise voltage but alters the noise current and so proportionally changes the noise impedance. This is because the base spreading resistance of the transistor – which, as far as noise is concerned, adds itself to the source resistance of the mic – is reduced by a factor dependent upon the number of devices placed in parallel. Noise is reduced by the square root of this factor. Shot noise (or Schottky noise) contribution can be reduced by maintaining a high collector current (which reduces dynamic emitter resistance). Voltage noise is inversely proportional to the square root of stage current whereas current noise increases proportionally to the square root of stage current. Fortunately, high current noise is of less importance when dealing with low impedance sources such as microphones. Optimisation of impedances is not necessary since the outputs of the front–end pair can be assumed to be feedback–derived zero impedance. With transistors as the front end amplifiers, care must be taken that the base–emitter junctions are not allowed to demodulate RF and inject it into succeeding stages. Judicious filtering of the input is thus required.

With bipolar transistors, the theoretical value for emitter–base voltage noise is a function only of absolute temperature and collector current thus:

$$e = KT\sqrt{\frac{2}{q\,Ic}}$$

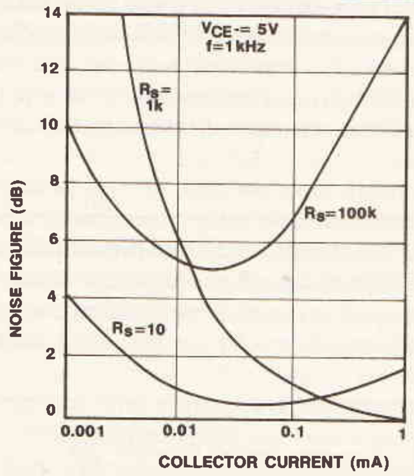
This formula indicates that a reduction in voltage noise, e , can be made low in value by increasing collector current. This is indeed bourne out in practise as Ic is increased until a level is reached where parasitic transistor noise limits any further reduction. This noise floor is usually created by and model led as an equivalent resistor (r_{bb}') – the so–called 'base–spreading resistance' – in series with the base of the transistor. This is, in fact, the real part of h_{ie} . Low parasitic transistor noise is therefore an important factor in ultra–low noise applications such as a mic pre–amp where Ic is pushed to the limit. In theory, a low noise design can therefore be done on paper with the minimum of bench–testing. In practise, I found this not to be the case!

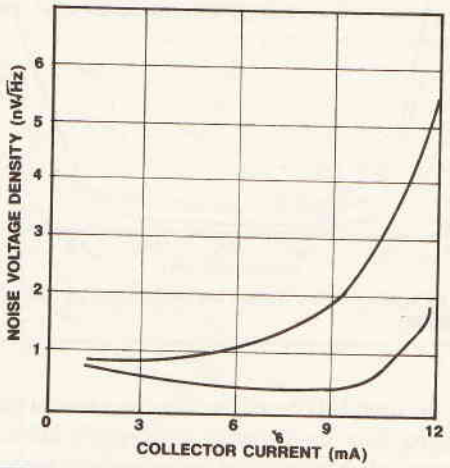
A further transistor noise component is base current noise and for finite source impedance, current noise must be considered as a quadrature addition to voltage noise. See Equation 2 in Fig 10. With a super–matched pair, base current noise is a well–defined function of collector current and can be expressed as in Equation 5. To find the collector current which yields the minimum overall equivalent input noise with a given Rs, the total noise formula can be differentiated with respect to Ic and set to zero for finding a minimum. Equation 1 and 2 are added together to give Equation 6. This is differentiated as in 7 and yields 8. For very low Rs – as in the case of a microphone – the r_{bb}'of the transistor must be added to Rs in the calculation.

Referring to the graph of collector current, I_c against e_n and i_n Graph 1 and using the adjacent equation 1, we can see e_n drops and i_n rises with increasing collector current means that we have a very simple method of optimising transistor noise for a given operating source impedance. Equation 2 shows that a noiseless source has an irreducible source of Johnson noise from the value of its source resistance. The amplifier adds its own noise as in Equation 3, consequently the amplifier's noise voltage is added to the input signal while its noise current generates a noise voltage across the source impedance. Since, as far as audio frequencies are concerned, the noise is uncorrelated, the square of the values are added. It is the low noise designer's job to pick Ic from the graph of e_n and i_n versus I_c so that the $e_n + (i_n Rs)$ term is minimised for a given source resistance in the area of the frequency of interest. From this graph, we can see that the sum of voltage terms with a 1k resistance is minimised for a collector current around 0.1mA. In the interests of lower noise figure, it might be provident to use a slightly lower Ic since current noise is reducing at a greater rate than voltage noise is increasing. However, with a low source resistance – 200R microphone – low e_n is slightly more important so in practise we in fact run the transistor at a higher collector current than at first seems optimum. Again, it's a case of compromising and conflicting operating conditions. Plugging the values into Equation 3 yields a noise figure of 0.8dB, which is pretty respectable. Graphs 2 and 3 of narrowband noise figure versus collector current and noise figure versus frequency bear out this value.

Referring to the circuit diagram, we can assume that both bases are grounded (by R45 and R52) so the emitters will be at 0.6V. This leaves Vcc – 0.6V to be dropped across the emitter resistors ie 16.4V. Since I_EQI_C, we control Ic with this emitter resistor. Collector resistor values are chosen such that the Vce max voltage isn't exceeded and also that the transistor operates as near 0V as possible so that full output swing is available. Although sometimes forgotten about, high common–mode signals are fully amplified by the transistors and then nulled by the following stage so that the maximum Vce rating of the transistor can be exceeded or operating headroom seriously eroded. The degenerating resistive element controls gain.

I have also shown, in Figure 4, the budget mic/line amplifier. This uses the much ridiculed straight op–amp amplification technique, so that the OSI condition isn't met and the noise performance of the stage worsens by around 10dB. The first type is similar in many respects to the low noise version although the mic amp IC is omitted. Gain in the second stage –after the Superbal – is altered by SW3d, the switch scuttling some of the gain when 'Line' is the source connected. This is the opposite to the situation in the low noise amp where gain was reduced when Mic was selected (since most of the gain was being made up by the SSM 2015). First stage mic amplification is provided by the Superbal which is switched to 'Mic' via SW3a and 3b. It should also be noted that there is now no provision for a front panel operated 'pad' in this second option of the budget pre–amp but the dual gang pot which replaces the pot/switch combination of the previous designs acts as a continuously variable attenuator. The choice of which





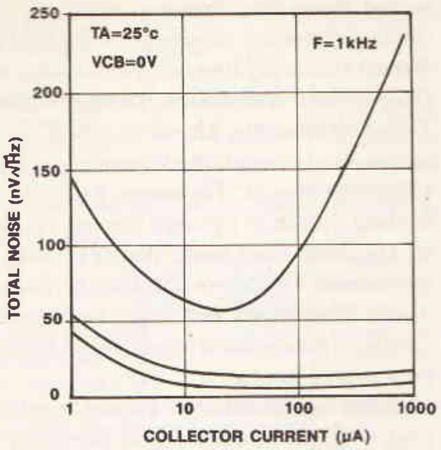
GAIN	R_G	$R_{PREFFERRED}$ (dB)	GAIN ERROR
20dB	3.076kΩ / –	3kΩ / 51k	+0.14 / +0.3
25dB	1.4k / –	1k5 / 15k	-0.48 / +0.05
30dB	711.2R / –	680R / 6k8	+0.34 / -0.2
35dB	379.2R / –	390R / 3k	-0.23 / +0.4
40dB	207.2R / –	200R / 1k5	+0.3 / +0.4
45dB	114.7R / –	110R / 910R	+0.36 / +0.8
50dB	63.95R / –	62R / 470R	+0.26 / +0.05
55dB	35.78R / –	36R / 240k	-0.06 / +0.4
60dB	20.07R / –	20R / 110R	+0.03 / +0.8
65dB	11.27R / –	11R / 47R	+0.2 / +0.6
70dB	– / –	– / 11R	– / +0.4

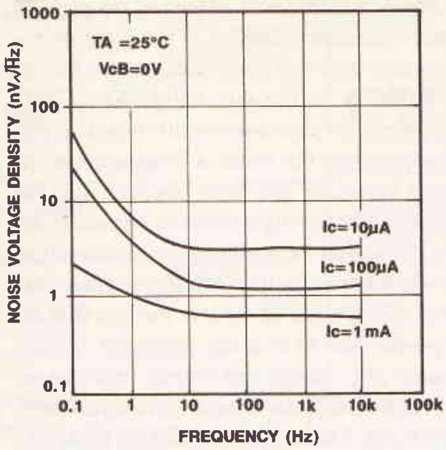
KEY

SSM2015 / SSM2220/MPSA55

SSM2015 figures by calculation, others by measurement

Fig.13 Typical performance characteristics of SSM2220





MIC Amplifer type	Equivalent input Noise *see note	Noise at20dB gain	Distortion (a1KHZ)Max gain min gain	CMRR @100Hz	CMRR @1KHz	CMRR @10KHz	Frequency response
SSM2015 integrated mic amp	-129dB (rms) -125dB4	-88dBA	0.007% 0.01%	99dB@ max gain 73dB @ min gain	97dB@ max gain 71dB @ min gain	96dB @ max 70dB @ min	20Hz-20kHz -0.3+0dB
SSM 2220 Discrete super-match	-128.5dBU(rms) -124.5dB4	-87dB4	0.008% 0.015%	91dB @ max gain 70dB @ min gain	80dB @ max gain 63dB @ min gain	79dB @ max 62dB @ min	20Hz-20KHz -0.3/40dB
MPSA55 Discrete Low noise transistor	-128dBU(rms) -12.4DB4	-85.5dB4	0.04% 0.05%	79dB @ max gain 55dB @ min gain	83dB @ max gain 70dB @ min gain	79dB @ max 55dB @ min	20Hz-20KHz -0.3/40dB
Straight Operational amplifier	-115dBU(rms) -111dB4	-84dB4	0.04% 0.08%	72dB @ max gain 70dB @ min gain	72dB @ max gain 53dB @ min gain	71dB @ max 45dB @ min	20Hz-20KHz -0.3/40dB

Note 1 EIN measured with 20KHz bandwidth, 200Ω termination, unweighted, max gain
Noise measurements (peak) done using PPM to CCIR/468-2 (dB4) = 0dBu (0.775Vrms referred to PPM4)
Noise measurements (rms) done using HEWLETT PACKARD 3561A oynamic signal analyzer
CMRR measured with +10dBu common mode tune at the input

Fig.14 Typical performance characteristics

option to fit is left entirely to the individual. The reader might be intrigued by Pl1, the section marked "Optional B Check Amplifier" and relay RLAl. All of these are to do with the multitrack and automation options and will be discussed at a later stage in the series.

For ultimate performance, the first–described circuit is the one to use. The subject of attenuation of the input is a sensitive one. The danger of overload from high level mic input transients is a very real one and not to be dismissed lightly.

Line level transients are less probable and as such, the line input pad should prove unnecessary in just about every situation. As with the mic input attenuator, if its inclusion is deemed expedient, then so be it. Enough information has thus far been presented for the constructor to make an informed and correct choice.

It has been said by more than one professional designer of mixing desks that the gain structure of the unit is one of the most important elements of the overall design because of the way in which noise, headroom and crosstalk are all related and in fact, inextricably linked to the relative signal levels at various points throughout the audio pathway. It is a reasoning with which I can heartily concur.

System Level Architecture

In order to maximise headroom, some designers have found it expedient to create what is known as 'non–unity level architecture' within the console. This means simply that all signal levels within the console are operated at a level other than 0dBu. Popular 'depressed' levels include –4dBu, –6dBu and –l0dBv. The –10dBv is a bit of a non–standard one – as the units might suggest – being the operating level found in some consoles manufactured in Japan. A unity level architecture is employed throughout the AutoMate with all levels being at a nominal 0dBu. This is for no better reason, I guess, than that coming from a corporate broadcast engineering background, I was used to all access and insert points in the audio pathway in a studio environment being both in–phase and at 0dBu. Being familiar with this operating philosophy, I adopted it as the norm. However, it is interesting to explore the reasons why non–unity levels can be attractive and useful to the engineer. In doing so, perhaps some intrepid constructors out there will then wish to alter the level architecture of the console when they come to build it. This is not as difficult or as ominous as it may at first appear, since it entails merely changing some resistor values in and around several key gain blocks within the console.

Anyway, taking the –4dBu level as a first example. As per Figure 5, everything operates at 4dB below 0.775V RMS, with 4dB of gain made up in the output stage using transformer step–up techniques. Again, the introduction of a transformer into a critical part of the audio pathway is somewhat deleterious to good performance – we have already discussed and documented at an earlier date the many shortcomings of this component. The –6dBu is a better choice because gain in the output stage can then be made up using an electronically–balanced amplifier stage. These facts still do not explain why a depressed operating level should be desirable at all. Again, it all has to do with the preservation of operating headroom. The erosion of headroom is most problematic in the input channel strip and specifically before the main gain–controlling element of the strip, the channel fader. Transient input signals and equalisation quickly gobble the margin between normal operating and overload (clipped) conditions. Lowering the operating level throughout the channel (channel–depressed architecture) is anothered favoured configuration because of the way in which headroom is extended where it is needed most (in the uncontrolled part of the input strip). Normal (0dBu) operating levels are then restored either in the post–fader buffer or in the mix–amps. Again, this outward improvement in headroom performance is not without cost. Residual mix bus noise is brought closer to the actual wanted signal by the same amount that the channel is depressed, but the high levels of signal on the bus – which was the reason for the depressed architecture in the first place –should swamp this. Another slight minus point from an operating point of view is the fact that all channel insert points now have a non–standard –l0dBu operating level and other sends and returns, aux mixes, PFL and solo mix busses will also have to have subsequent gain adjustments made. There is, I'm afraid, no complete and ideal solution.

Next month, we look at the op–amp as it relates to our application and then move on to discuss equalistion.

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PARTS LIST – LOW NOISE MIC/LINE PREAMP

RESISTORS (all resistors 1/4W, 1% tolerance unless otherwise stated)

R1, 3, 36	1k (R1, 3 optional, see text)
R2	220R (optional, see text)
R4	100R
R5, 6	6k8
R7 – 10, 15, 16, 27, 35	10k
R11, 12	_
R13, 14	20k
R17	3k
R18	1k5
R19	680R
R20	390R
R21	200R
R22	110R
R23	62R
R24	36R
R25	20R
R26	11R
R28	9k1
R, 31, 32, 42	100k
R30	(optional pad resistors, see text)
R33	1M
R34	2k
R37, 39	30k (optional, see text)
R38, 40	10k (optional, see text)
R41	AOT
R43	7k5
PR1	100k vertical multiturn preset (cermet)
PR2	1k vertical multiturn preset (cermet)
VR1	10k LIN pot c/w optional push on switch
CAPACITORS	
C1, 2, 3, 10, 11	47u 63V electrolytic
C4	27p polystyrene
C5, 6, 25-32	100n disc ceramic
C7	47p polystyrene
C8	4p7 polystyrene
C9, 12	100n polyester
C13-15, 21, 23	33p polystyrene
C16, 16a, 17, 18, 24	470n polyester (C16, 16a optional)
C33	220p polystyrene
SEMICONDUCTORS	
IC1	SSM 2015
IC2	NE5532A
IC3	NE5534A
IC4, 5	TL071 (IC5 optional, see text)
ZD1-4	6V2, 400mW zeners
MISCELLANEOUS	
XLR 1, 2	PCB-mounting 3 pin chassis socket
JK1	1/4" PCB-mounting stereo switched socket
SW2	PCB-mount DPDT latching switch
SW3	4 pole PCB-mounting latching switch
SW4	12 way PCB-mounting rotary switch
PL1	2-way Minicon plug

Veropins, PCB, IC sockets to suit etc

PARTS LIST – (SSM 2220 SUPER MATCHED TRANSISTOR VERSION)

RESISTORS (all resistors 1/4W, 1% tolerance unless otherwise stated)

R1, 3, 36	1k (R1, 3 optional, see text)
R2	220R (optional, see text)
R4	100R
R5, 6, 19	6k8
R9 – 10, 15, 16, 27, 35	10k
R11, 12	_
R13, 14	20k
R17	51k
R18	15k
R20	3k
R21	1k5
R22	910R
R23	470R
R24	240R
R25	110R
R26	47R
R28	–
R30	(optional, see text)
R31, 32, 42, 51	100k
R33	1M
R34	2k
R36	1k
R37, 39	30k (optional, see text)
R38, 40	10k (optional, see text)
R41	AOT
R43	7k5
R44, 48	8k2
R45, 46, 49, 52	4k7
R47	optional link
R50	91k
PR1	10k vertical multiturn preset (cermet)
VR1	10k LIN pot c/w optional push on switch (SW1)
CAPACITORS	
C1, 2, 3, 10, 11	47u 63V electrolytic
C4	–
C5, 6, 25, 26, 29, 30, 31, 32, 43, 45	100n disc ceramic
C7, 8	–
C9, 12	100n polyester
C13-15, 21, 23	33p polystyrene
C17, 18, 24, 37	470n polyester
C19, 20	6u8 polyester (optional, see text)
C22	100p polystyrene (optional, see text)
C33, 35, 44	470p polystyrene
C34, 38-40	10u 63V radial electrolytic
C41, 42	4p7 polystyrene
C36	100n 63V radial electrolytic
SEMICONDUCTORS	
IC1, 4, 5	TL071 (IC5 optional, see text)
IC2	NE5532A
IC6	SSM2220
MISCELLANEOUS	
XLR 1, 2	PCB-mounting 3 pin chassis socket
JK1	1/4" PCB-mounting stereo switched socket
SW2	PCB-mount DPDT latching switch
SW3	4 pole PCB-mounting latching switch
SW4	12 way rotary switch
PL1	2-way Minicon plug

Veropins, PCB, IC sockets to suit etc

PARTS LIST – BUDGET DISCRETE MIC/LINE PREAMP

RESISTORS

(all resistors 1/4W, 1% tolerance unless otherwise stated)

R1, 3, 36	1k (R1, 3 optional, see text)
R2	220R (optional, see text)
R4	100R
R5, 6, 19	6k8
R9 – 10, 15, 16, 27, 35	10k
R11, 12	_
R13, 14	20k
R17	51k
R18	15k
R20	3k
R21	1k5
R22	910R
R23	470R
R24	240R
R25	110R
R26	47R
R27	11R
R28	–
R30	(optional, see text)
R31, 32, 42	100k
R33	1M
R34	2k
R36	1k
R37, 39	30k (optional, see text)
R38, 40	10k (optional, see text)
R41	AOT
R43	7k5
R44, 48	9k1
R45, 52	4k7
R46, 49	5k6
R47	optional link
R50, 51	47k
VR1	10k LIN pot c/w optional push on switch (SW1)
CAPACITORS	
C1, 2, 3, 10, 11	47u 63V electrolytic
C4	–
C5, 6, 25, 26, 29, 30, 31, 32, 43, 45	100n disc ceramic
C7, 8	–
C9, 12	100n polyester
C13-15, 21, 23	33p polystyrene
C17, 18, 24, 37	470n polyester
C19, 20	6u8 polyester (optional, see text)
C22	100p polystyrene (optional, see text)
C33, 35, 44	470p polystyrene
C34, 38-40	10u 63V radial electrolytic
C41, 42	4p7 polystyrene
SEMICONDUCTORS	
IC1, 4, 5	TL071 (IC5 optional, see text)
IC2	NE5532A
Q1, 2	MPSA 55
MISCELLANEOUS	
XLR 1, 2	PCB-mounting 3 pin chassis socket
JK1	1/4" PCB-mounting stereo switched socket
SW2	PCB-mount DPDT latching switch
SW3	4 pole PCB-mounting latching switch
SW4	12 way rotary switch
PL1	2-way Minicon plug

Veropins, PCB, IC sockets to suit etc

PARTS LIST – (SSM 2220 SUPER MATCHED TRANSISTOR VERSION)

RESISTORS

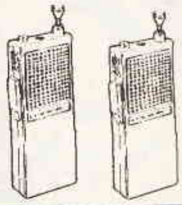
(all resistors 1/4W, 1% tolerance unless otherwise stated)

R1, 3, 7, 8, 17-13	–
R4	100R
R5, 6	6k8
R8, 10, 15, 16, 38, 40	10k (R38, 40 optional, see text)
R11, 12, 36	1k
R13, 14	20k
R34	1k8
R35	200R
R37, 39	30k (optional, see text)
R41	AOT
R43	7k5
VR1	10k LIN pot c/w optional push on switch
CAPACITORS	
C1, 2, 3	47u 63V electrolytic
C4-8	–
C9, 12	100n polyester
C13-15, 21, 23	33p polystyrene
C16-18	–
C19, 20	6u8 polyester (optional, see text)
C22	100p polystyrene (optional, see text)
C25-32	100n disc ceramic
SEMICONDUCTORS	
IC1	–
IC2	NE5532A
IC4, 5	TL071 (IC5 optional, see text)
MISCELLANEOUS	
XLR 1, 2	PCB-mounting 3 pin chassis socket
JK1	1/4" PCB-mounting stereo switched socket
SW1, 2	PCB-mount DPDT latching switch
SW3	4 pole PCB-mounting latching switch
SW4	12 way rotary switch
PL1	2-way Minicon plug

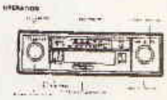
Veropins, PCB, IC sockets to suit etc

AMAZING ADAPTER BUG
Built into a standard 13A adapter, plugs into any 13A socket and transmits to a normal FM radio. Directly powered from the mains the unit will transmit conversations etc indefinitly! Price is £26.00 ref M26P1

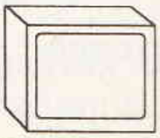
WALKIE TALKIES 1 MILE RANGE
Pair of small pocket sized walkie talkies complete with cases etc. They will operate (subject to buildings etc) up to 1 mile apart. 2 PP3 9v batteries required. £30.00 ref M30P1



CAR STEREO AND SPEAKERS
Complete system comprising of stereo cassette player, stereo FM radio plus AM band, Pair of good quality speakers all for just £19.00 ref M19P1



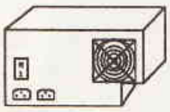
AMBER MONITORS
12" high res screen 12v 1A supply needed. Hercules\TTL input ie sep HOR and VER sync plus video required. Brand new and-housed in an off white plastic case. £22.00 ref M22P1.



12V SOLAR PANEL
Ideal for trickle charging car batteries etc. Panel is made from amorphous silicon, is waterproof and comes with fly leads. Size is 30cm x 30cm x 4mm. £15.00 ref M15P1. Other sizes stocked

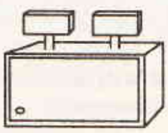
C64 TAPE STREAMER
Originally made for the Commodore 64 Computer but may be adaptable for other machines. Unit is supplied with its own operating system, and two tapes. Approx 20 times faster than normal tape systems! £25.00 ref M25P1.
Extra tapes are available at £4.00 each ref M4P1 or 10 for £25.00 ref M25P2.

PC POWER SUPPLIES
Brand new units made by Aztec either 110v or 240v input giving 5v at 15A, 12v at 5A, -5v at .3A and -12 at .5A. Fully cased with on/off switch and built in fan. £15.00 ref M15P2
Also available is a 200 watt version at £22.00 ref M22P2.
Both types have standard PC fly leads.

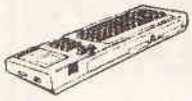


40 PAGE CATALOGUE AVAILABLE CONTAINING OVER 1,500 SIMULAR PRODUCTS FREE ON REQUEST!

EMERGENCY LIGHTING SYSTEM
Complete system give sup to 3 hours light from an integral 10AH sealed lead acid battery. The battery is kept fully charged by the mains, as soon as the mains fails the two powerful lamps are switched on and remain on until power is restored. Maintenance free. £19.00 complete with battery. ref M19P2



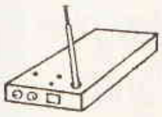
AMSTRAD 464 COMPUTERS
Customer returned units complete with a monitor and circuit diagrams. These units are generally not working and are not returnable. Price is £35.00 ref M35P1.



AMSTRAD 6128 COMPUTERS
Customer returned units complete with a circuit diagram and built in 3" disc drive. These units are generally not working and are not returnable. Price is £29.00 ref M29P1.

CORDLESS MICROPHONE
Small hand held battery operated microphone that transmits to a standard FM radio, good range. Our price £15.00 ref M15P3.

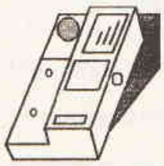
VIDEO SENDER
Transmits video pictures from a video recorder or cam corder to any television in the housel. Can also be used to transmit from cam corder to video recorder, no more trailing wires! £15.00 (ex psu) ref M15P4 £20.00 (inc psu) ref M20P1.



BUILT BUG
Built and tested superior FM bug 100m range, fits in match box all you need is a 9v battery and an ordinary FM radio! £14.00 ref M14P1.

ULTRASONIC ALARM SYSTEM
Complete alarm system that comprises a detector that simply plugs into a 13A socket in the area you wish to protect and a receiver which plugs into a 13A socket where you wish the alarm to sound. You could put one in the garage and one indoors or perhaps protect your neigbours house etc. Fully adjustable sensitivity. £25.00 for complete system ref M25P3.

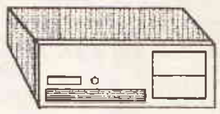
TALKING TELEPHONE COIN BOXES
Phone bill too high? fit one of these and save. Fully programmable for different call rates, chargebands, time of day etc. accepts 10p, 50p and £1.00 coins. Phone box actually speaks to you with built in voice synthesiser. Wall or desk mounting. Two types available 1 with built in lock at £29.00 ref M29P2 the other with no lock but easily adaptable is just £23.00 ref M23P1. Unit takes 4 C cells and is used in conjunction with an ordinary phone. Supplied with full instructions, BT approved.



STEAM ENGINES
Ever wanted one? brand new units made by the famous Mamod company complete with fuel, burner etc £30.00 ref M30P1.
Other models stocked, including traction engine at £59.

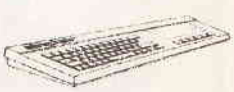
BULL ELECTRICAL
250 PORTLAND ROAD HOVE SUSSEX BN3 5QT TELEPHONE 0273 203500
MAIL ORDER TERMS: CASH PO OR CHEQUE WITH ORDER PLUS £3.00 POST PLUS VAT.
PLEASE ALLOW 7 - 10 DAYS FOR DELIVERY
NEXT DAY DELIVERY £8.00
SHOP OPEN MON-SAT
FAX 0273 23077
VISA

PC CASES
Full size off white metal cases ideal for building your own PC four drive bays, attractive plastic front panel. £24.00 ref M24P1.



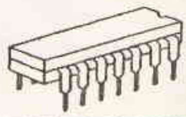
VIEWDATA SYTEMS
Made by Tandata these contain every thing you need to start dialling into databases and bullitin boards such as Prestel etc just plugs into a standard tv or monitor. Complete with modem, infrared remote controlled keyboard and console. £20.00 ref M20P2

SPECTRUM +2 COMPUTERS
Refurbished popular computer with built in cassette deck and 128K of memory £32.00 each ref M32P1 PSU £15.00 ref M15P4
We also have some requiring attention at £19.00 (non returnable) ref M19P3



SPECTRUM +3 COMPUTERS
Refurbished popular computer with built in disc drive and 128k of memory £45.00 each ref M45P1 PSU £15.00 ref M15P5. We also have have some requiring attention at £25.00 (non returnable) ref M25P4.

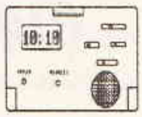
EPROMS
Clean erased eproms at bargain prices! 27C64 pack of 10 for £7 ref M7P1, 27C256 pack of 10 for £9 ref M9P1, 27C512 pack of 10 for £10 ref M10P1.



WINDUP SOLAR POWERED RADIO
Compact unit with built in hand charger and solar panel just a few turns of the handle powers the radio for some time! Our price is just £14.00 ref M14P2

BUGGING TAPE RECORDER
Contains voice activated switch so only actual conversations are recorded! takes a standard audio cassette and uses AA batteries. £20.00 ref M20P3

TALKING ALARM CLOCK
Wakes you up by telling you the time also speaks the time at the push of a button! Battery operated £14.00 ref M14P3.

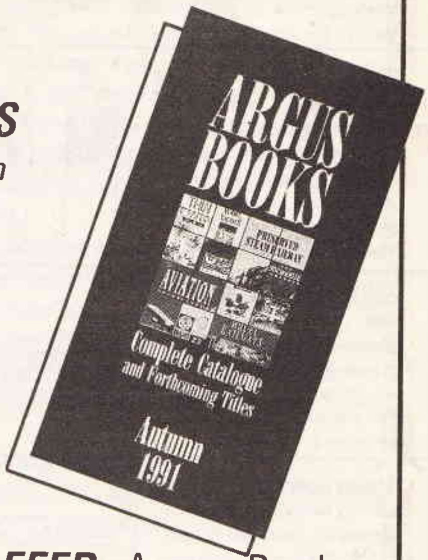


BROADBAND RADIO RECEIVER
Covers VHF 54-176 mhz (CB, air FM, TV, PB, WB etc etc) hand held unit with squelch control and carrying strap. £15.00 ref M15P6 Superb value.

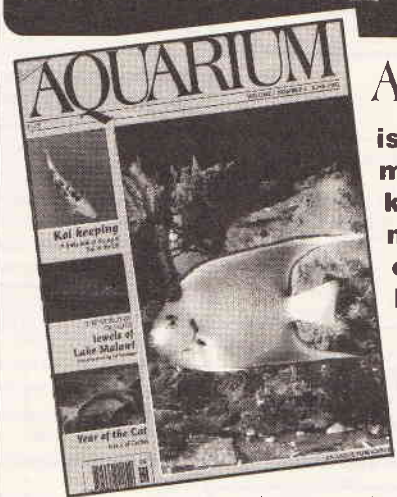
12 BAND WORLD COMMUNICATIONS RECEIVER
Mains or battery operated covers 9 short wave bands plus FM, LW and AM bands.
Exceptional value at £19. ref M19P4

SINCLAIR C5 MOTORS
12v 29A (full load) complete with 4 to 1 reduction gearbox giving 800 rpm output. Motor measures 8" x 4" with toothed pulley output. £40 ref M40P1. We also stock 13" wheels with tyres at £6 each ref M6P1, 16" at £6.00 ref M6P2 and an electronic speed controller kit at £17 ref M17P1.

NOTE: SOME OF OUR PRODUCTS MAY NOT BE LICENSABLE FOR UK



ARGUS BOOKS
AVIATION
Complete Catalogue and Forthcoming Titles
Autumn 1991



AQUARIUM
Koi keeping
Jewels of Lake Malawi
Year of the Cat

PHOTO -COPIES

ETI has been published for over 19 years and in this time a great many interesting and valuable projects and features have graced its pages. Although backnumbers are available fr only the past 12 month's issues we can supply photocopies of any individual article ever published in ETI.

Photocopies cost £1.50 per article regardless of their length. Please note that projects published over serveal issues must be ordered as a series of individual articles, each for £1.50.

PCB foil patterns (where published) and any errata are included with all photocopies as applicable.

BACK NUMBERS

- Missed part one of your favourite project?
- Lost that all-important article on circuit design?
- Desperately need that Tech Tip to get your latest materpiece going?
- Pining for that one issue to complete your valuable collection?

ETI Backnumbers Department to the rescue!

Please not that backnumbers are only held for 12 months. Articles published pervious to this must be obtained through our photocopy service.

Please supply photocopies of the following articles from ETI (complete in block capitals):

Month .... Year .... Page (if known) ....

Title ....

Month .... Year .... Page (if known) ....

Title ....

Month .... Year .... Page (if known) ....

Title ....

I enclose a cheque/postal order made out to ASP Ltd. to the value of £1.50 per photocopy ordered.

Total remittance £ .... Date ....

Name ....

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Send the completed form and your remittance to:

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Argus House
Boundary Way
Hemel Hempstead
Herts HP2 7ST

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Month .... Year .... Month .... Year

Month .... Year .... Month .... Year

I enclose a cheque/postal order made out to ASP Ltd. to the value of £2.20 per issue ordered.

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.... Postcode ....

Send the completed form and your remittance to:

ETI Backnumbers Department
Argus Subscription Services
Queensway House
2 Queensway
Redhill
Surrey RH1 1QS

BARGAINS - 10 New Ones This Month

FREE POWER! Can be yours if you use our solar cells – sturdily made modules with new system bubble magnifers to concentrate the light and so eliminate the need for actual sunshine – they work just as well in bright light. Voltage output is .45 – you join in series to get desired voltage – and on parallel for more amps.

Module A gives 100mA	Price **£1**	Order Ref. BD631
Module B gives 400mA	Price **£2**	Order Ref. 2P199
Module D gives 700mA	Price **£3**	Order Ref. 3P42
Module E gives 1A	Price **£3.50**	Order Ref. 3.5P4

LIMITED SUPPLY ITEMS
are only described in our newsletter. Over 50 appear in our current issue. If you order something this month you will recieve this and the next three issues posted to you free of charge.

ULTRA THIN DRILLS Actually 0.3mm. To buy these regular cost a fortune. However, these are packed in half dozens and the price to you is **£1** per pack, Order Ref. 797B.

BUILD YOUR OWN NIGHT LIGHT, battery charger or any other gadget that you want to enclose in a plastic case and be able to plug into a 13A socket. We have two cases, one 3¼ x 2¼ x 1¾" deep, **£1** each Order Ref. 845. The other one is 2½ x 2¼ x 1¾" deep, 2 for **£1**, Order Ref. 565.

SAFETY LEADS curly coil so that they contract but don't hang down. Could easily save a child from being scalded. 2 core, 5A, extends to 3m, **£1**, Order Ref. 846, 3 core, 13A extends to 1m, **£1** each, Order Ref. 847, 3 core, 13A, extends to 3m, **£2** each, Order Ref. 2P290.

POWER SUPPLY WITH EXTRAS mains input is fused and filtered and the 12V dc output is voltage regualted intended for high class equipment, this is mounted on a PCB and also mounted on the board but easily removed, are 2 12V relays and a Piezo sounder. **£3**, Order Ref. 3P80B.

100W MAINS TRANSFORMER normal primary 20-0-20 at 2.5A, **£4**, Order Ref. 4P24. 40V at 2.5A, **£4**, Order Ref. 4P59. 50V ar 2A, **£4**, Order Ref. 4P60.

PHILIPS 9" HIGH RESOLUTION MONITOR black & white in metal frame for easy mounting, brand new still in maker's packaging, offered at less than price of tube alone, only **£15**, Order Ref. 15P1.

16 CHARACTER 2-LINE DISPLAY screen size 85mm x 36mm, Alphanumericc LCD dot matrix module with integral; micro processor made by Epson, their Ref. 16027Ar, **£8**, Order Ref. 8P48.

INSULATION TESTER WITH MULTI METER internally generates voltages which enables you to read insulation directly on megohms. The multimeter has four ranges, AC/DC volts, 3 ranges DC milliamps, 3 ranges resistance and 5 amp range. These instruments are ex British Telecom but in very good condition, tested and guaranteed OK, probably cost at least £50 each yours for only **£7.50**, with leads, carry case **£2** extra, Order Ref. 7.5P/4.

THIS MONTH'S SLIP
A £60 unit for less than £10 — switch mode power supply with outputs +12V at VA, +5V at 16A and -12V at ½A. Enclosed in plated steel case, brand new, offered at a special price of £9.50 until July 31st, Order Ref. 9.5P1.

BRUSHLESS DC 12V FAN tiny, only 60mm square, good air mover but causes no interference, **£8**, Order Ref. 8P26.

MAINS 230V FAN best make "PAPST" 4½ square, metal blades, **£8**, Order Ref. 8P8.

2MW LASER Helium neon by Phillips, full spec. **£30**, Order Ref. 30P1. Power supply for this kit form with case is **£15**, Order Ref. 15P16, or in larger case to house tube as well **£18**, Order Ref. 18P2. The larger unit, made up, tested and ready to use, complete with laser tube **£69**, Order Ref. 69P1.

1/3 HP 12V MOTOR — THE FAMOUS SINCLAIR C5 brand new, **£15**, Order Ref. 15P8.

SOLAR CHARGER holds 4AA nicads and recharges these in 8 hours, in very neat plastic case, **£6**, Order Ref. 6P3.

SOLAR CELLS with terminals for joining in series for higher volts or parallel for extra current: 100m A. **£1**, Order Ref. 361, 400mA. **£2**, Order Ref 2P199, 700mA. **£3**, Order Ref. 3P42, 1A, **£3.50**, order Ref. 3.5P/4.

AIR SPACED TRIMMER CAPS 2-20 pf ideal for precision tuning UHF circuits, 4 for **£1**, Order ref. 818B.

FIELD TELEPHONES just right for building sites, rallies, horse shows, etc., just join two by twin wire and you have two way calling and talking and you can join into regular telephone lines if you want to. Ex British Telecom in very good condition, powered by batteries (not included) complete with shoulder slung carrying case, **£9.50**, Order Ref. 9.5P/2.

PROJECT BOX is approx 8" x 4" x 4½" metal sprayed grey, louvred ends for ventilation otherwise undrilled. Made for GPO so best quality, only **£3** each, Order Ref. 3P74.

ULTRA SONIC TRANSDUCERS 2 metal cased units, one transmits, one receives. Built to operate around 40kz. Price **£1.50** the pair, Order Ref. 1.5P/4

10W ISOLATION TRANSFORMER if you want to isolate some small piece of equipment from earth this may be what you are looking for. You can use it either with a tapped input or a tapped output. It's upright mounting and well insulated. Price **£1**, order Ref. 821.

3V SOLAR PANEL price **£3**, Order Ref. 3P99B

BT POWER SUPPLY UNIT ouput 9.5v AC at 600mA, in black plastic case with 13A plugs to go straight into socket and approximately 3 metres of twin output lead. Price **£1.50**, Order Ref. 1.5P7

JUST ARRIVED
3 core 15A flex. Price is 10m for **£2.50**, Order Ref. 2.5P/3. You can have this in multiples of 10m up to 100m. Also available: 3 core 20A flex, 10m for **£3**, Order Ref. 3P109 and 2 core 20A flex, 15m for **£3**, Order Ref. 3P110.

FERRITE AERIAL ROD 8" long x 3/8 diameter, made by Mullard. Complete with 2 coil formers. 2 for **£1**, order Ref. 832B

3 GANG.0005 MFD TUNING CONDENSER with slow motion drive. Beautifully make by Jackson Brothers and current list price is probably around £20. Yours for **£5**, Order Ref. 5P189

DIGITAL FREQUENCY METER.
This is a hand-held instrument with an LCD display allowing 8 digits of frequency to be read, has internal nicad batteries, and a power supply which will recharge the batteries. Ideal for field and service work as well as general and industrial applications. Has high and low BNC inputs and a plug-in antenna which enables remote tests. It covers a very wide range of frequencies: switch position 'A' covers 10Hz to 20MHz and switch position 'B' covers 20MHz to 1200MHz.. Price **£99**, but it compares very favourably with instruments selling at over £500 by our competitors. Order Ref. 99P2.

MAINS ISOLATION TRANSFORMER stops you getting "to earth" shocks. 230V in and 230V out. 150 watt upright mounting, **£7.50**, Order Ref. 7.5P/5 and a 250W version is **£10**, Order Ref. 10P79.

MINI MONO AMP on PCB. Size 4: x 2" with front panel holding volume control and with spare hole for switch or tone control. Output is 4 watt into 4 ohm speaker using 12V or 1 watt into 8 ohm using 9V. Brand new and perfect, only **£1** each, Order Ref. 495.

5RPM MAINS DRIVEN This is a shaded pole motor, **£5**, Order Ref. 5P54.

POWER SUPPLY UNIT mains in, dc out, cased 4.5V 100mA, **£1**, Order Ref. 104, 6V 200mA **£1**, Order Ref. 103, 6V 700mA, **£1**, Order Ref. 103A, 9V 500mA, **£2**, Order Ref. 2P134, 24V 200mA, **£2**, Order Ref. 2P4, 12V 2A, **£6**, Order Ref. 6P23.

AMSTRAD POWER UNIT 13.5V at 1.9A encased and with leads and output plug, normal mains input **£6**, Order Ref. 6P23.

AMSTRAD 3.5 FLOPPY DRIVE brand new and cased, **£35**, Order Ref. 35P4.

ATARI 65XE at 65K this is quite powerful, so suitable for home or business, unused and in perfect order but less PSU, only **£19.50**, Order Ref. 19.5P/5B.

80W MAINS TRANSFORMER two available, good quality, both with normal primaries and upright mounting, one is 20V 4A, Order Ref. 3P106 the other 409V 2A, Order Ref. 3P107, only **£3** each.

12V SOLENOID has good ½" pull or could push if modified, size approx 1½" long x 1" square, **£1**, Order Ref. 232.

WATER VALVE 230V operated with hose connections, ideal for auto plant spray or would contol air or gas into tanks etc., **£1** each, Order Ref. 370.

HANG-UP PHONE won't clutter up your desk or workbench, current model has push button dialling, last number recall, internal alarm etc. Ex B.T. in good condition and fully working ready to plug in, **£5**, Order Ref. 5P123.

ELECTRONIC BUMP & GO SPACESHIP sound and impact controlled responds to claps and shouts and reverses or diverts should it hit anything! Kit with really detailed instructions, will make ideal present for budding young electrician. Should be able to assemble but you may have to help with the soldering of the components on the PCB. Complete kit **£8.95** Order Ref. 9P9.

500V BRIDGE MEGGER developed for GPO technicians the Ohmeter 18B is the modern equivalent of the bridge megger. 9V battery operated, it incorporates a 500V generator for insulation testing and a null balance bridge for very accurate resistance measurement. Ex B.T. in quite good condition with data & tested. Yours for a fraction of original cost, **£45**, Order ref. 45P2.

EXPERIMENTING WITH VALVES don't spend a fortune on a mains transformer, we can supply one with standard mains input and secs. of 250-0-250V at 75mA and 6.3V at 3A, **£5**, Order Ref. 5P167.

15W 8 OHM 8" SPEAKER & 3" TWEETER made for a discontinued high quality music centre, gives real hi-fi, and only **£4** per pair, Order Ref. 4P57.

CLEAR THAT SMOKE according to a 'Which' report, many ionisers available from chemists and similar have such a poor output that they are next to useless. Our ioniser kit, however, uses mains transormers and is so powerful you can feel the ion output on the back of your hand and it will clear smoke in seconds. Complete, cased kit, price **£16**, Order Ref. 16P5.

ULTRASONIC TRANSMITTER RECEIVER with Piezo alarm, built into preformed case is triggered by movement disturbing reflected signal, intended for burglar alarm, car alarm etc. has many extras, time delay, auto reset, secret 'off' device, etc. A £40 instrument, yours for **£10**, Order Ref. 10P76.

STEREO HEADPHONES extra lightweight with plug, **£2** each, Order Ref. 2P261.

BT TELEPHONE LEAD 3m long and with B.T. flat plug ideal to make extension for phone, fax etc. 2 for **£1**, Order Ref. 552.

WATER PUMP very powerful with twin outlets, an ideal shower controller, mains operated, **£10**, Order Ref. 10P74.

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PCB SERVICE August

E9208-1 dynamic Noise Limiter ....F

E9208-2 Touch Controlled Intercome (2 boards) ....H

E9208-3 MIDI Keyboard...K

E9208-FC Battery charger ....F

PCBs for the remaining projects are available from the companies lists in Buylines.
Use the form or a photocopy for your order. Please fill our all parts of the form. Make sure you use the board reference numbers. This not only identifies the board but also tells you when the project was published. The first two numbers are the year, the next two are the month.
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E8806-6	Bicycle Dynamo Backup	D
E8807-1	Bar Code Lock (2 bds)	N
E8807-2	Analogue Computer Power Board	L
E8807-3	Bell Boy	F
E8807-4	Logic Probe	C
E8807-5	Updated FM Stereo Decoder	J
E8807-6	Breath Rate Display Board	F
E8808-1	Breath Rate Main Board	H
E8808-2	Breath Rate Switch Board	C
E8808-3	Telephone Recorder	D
E8808-4	Analogue Computer Main Board (2 bds)	M
E8809-1	Spectrum EPROM Emulator	M
E8809-2	Frequency Meter (2 bds)	P
E8809-3	Travellers' aerial Amp	E
R8810-1	Gerrada Marweh Bikebell	E
E8810-2	Peak Programme Meter (2 bds)	N
E8810-4	TV to RGB Converter	E
E8810-5	Electron RGB Buffer	C
E8811-1	Ni-Cd Charger	E
E8811-2	Chronoscope (3 bds)	P
E8811-3	Digital Transistor Tester	G
E8812-1	Doppler Speed Gun (2 bds)	K
E8812-2	Small Fry Mini Amp	D
E8812-3	Thermostat	E
E8812-4	Burglar buster Free PCB	D
E8812-5	Burglar Buster Power/relay Board	E
E8812-6	Burglar Buster Alarm Board	C
E8812-7	Burglar Buster Bleeper Board	C
E8901-1	EPROM Programmer mother board	M
E8901-2	Variat-Ion updated Main Board	H
E8901-3	Variat-Ion Emitter Board	E
E8901-4	In-car Power Supply	C
E8901-5	Granny's Hearing Booster	E
E8902-1	Compressor/Limiter/Gate	L
E8902-2	Ultrasonic Horn	D
E8902-3	Stepper Motor Driver Board	L
E8902-4	Quest-Ion (2 bds)	K
E8903-1	Intelligent Plotter Solenoid Board	H
E8903-2	MIDI Programmer	L
E8903-3	Balanced Disc Input Stage	F
E8903-4	Digitally Tuned Radio	G
E8904-1	Camera Trigger	E
E8904-3	Intelligent Plotter Main Board	O
E8904-4	Kinetotie Tie Board	N
E8904-5	Kinetotie Control Board	E
E8905-1	Guitar Tuner	H
E8905-2	Camera Trigger Ultrasonics (2 boards)	F
E8905-3	Bench Power Supply (2 boards)	H
E8906-1	PC edge connector	F
E8906-2	MIDI converter CPU	N
E8906-3	MIDI converter keyboard	N
E8906-4	MIDI converter control	M
E8906-5	AF signal generator	G
E8906-6	Mini bleeper	C
E8906-7	Caravan heater controller	G
E8907-1	MIDI Patch Bay	G
E8907-2	Priority Quiz Switch	E
E8907-3	Camera Trigger Infra-reds (2 boards)	G
E8907-4	Aerial Amplifier main board	E
E8907-5	Aerial Amplifier power supply	E
E8908-1	Intercom master station	L
E8908-2	Intercome slave station	F
E8908-3	Intercom power mixer	E
E8908-4	Digital joystick to mouse conversion	H
E8909-1	Twin Loop Metal Locator	H
E8909-2	Trembler movement detector	D
E8909-3	Field power supply (spec 3)	C
E8909-4	Micro monitors active filter	F
E8909-5	Chronoscope auto-reset	C
E8910-1	Multimeter	H
E8910-2	MIDI Mapper	M
E8911-1	Smoke Alarm main board	F
E8911-2	Smoke Alarm power supply	F
E8911-3	Frequency Meter (3 board)	O
E8911-4	Serial Logic Scope	L
E8912-1	Mains Failure Alarm	D
E8912-2	Surveilance PCB	D
E8912-3	Slide/tape Synch	E
E8912-4	Pedal Power	L
E8912-5	Digital Noise Generator	K
E9001-1	20 metre Receiver	J
E9001-2	Wavemaker FG	L
E9001-3	Motorcycle Intercom	F
E9001-4	Low Voltage Alarm	C
E9002-1	EPROM Emulator	N
E9002-2	Superscope Mother Board	M

E9002-3	Superscope CRT Driver Board	K
E9002-4	Superscope Timebase Board	K
E9003-1	Superscope Y1 input board	J
E9003-2	Superscope Y2 input board	J
E9003-3	Superscope switch generator	E
E9003-4	Business power amp board	L
E9003-5	Business power supply board	J
E9003-7	Water hole	G
E9003-8	Super Siren	D
E9003-9	Val's badge	F
E9004-1	Bass Amplifer DC Protection	F
E9004-2	Bass Amplifier Graphic Equaliser	L
E9004-3	Bass amplifier Micro	N
E9004-4	Quad Power Supply	O
E9005-1	Business Display	O
E9005-2	Phone Lock and Logger	F
E9006-1	Dark Room Timer	G
E9006-2	Telephone Extension Bell	C
E9006-3	Telephone External Bell	D
E9006-4	Fecko Box	G
E9006-5	Bug Spotter	E
E9007-1	Guitar Practice Amp	G
E9007-2	Digital Frequency Meter	M
E9007-3	Footstep Alarm	E
E9007-4	Transistor Tester	C
E9007-5	Decision Maker	J
E9008-1	AC Millivoltmeter	K
E9008-2	Temperature Controller	N
E9008-3	FM Generator	L
E9009-1	Slide Projector Controller	E
E9009-2	Ultimate Diode Tester	D
E9009-3	The Entertainer	G
E9010-1	Component Tester	F
E9010-2	Active Contact Pickup	E
E9010-3	R4X Longwave Receiver	C
E9011-1	The Autocue (2 boards, 1 double sided)	N
E9011-2	Infra-lock transmitter (2 boards)	K
E9011-3	Infra-lock receiver	H
E9011-4	Four-track cassette recorder (record/playback one channel)	F
E9011-5	Four-track cassette recroder (Bias/erase oscillator board	K
E9012-1	Infra Switch	F
E9101-1	Remote Control – Main Board	J
E9101-2	Remote Control - Display Board	H
E9101-3	Remote Control Timeswitch – Transmit bd	E
E9101-4	SBC Micro Controller Board	F
E9101-5	SBC Practice Interface Board	F
E9101-6	5 in 1 Remote Sensing Switch	E
E9102-1	Remote Control Timeswitch – receiver bd	F
E9102-2	anti theft Alarm (2 bds)	H
E9103-1	Ariennes Lights	L
E9103-2	64K EPROM Emulator	N
E9103-3	SSB Radio Receiver	G
E9103-4	Active Loudspeaker board	H
E9104-1	Testmeter Volts	E
E9104-2	Active Direct Injection Box	F
E9104-3	EPROM Eraser	F
E9104-4	Digital Tachometer	F
E9104-5	Radio Calibrator	F
E9105-1	Modulator Laser (2 boards)	H
E9105-2	Thyristor Tester	F
E9105-3	Frequency plotter	K
E9106-1	Laser Receiver	F
E9106-2	Temperature Controller – Power Supply	G
E9107-1	Temperature Controller – Main Board	K
E9107-2	Temperature Controller – Probe PCB	F
E9107-3	The Foot Tapper – Volume Control (2 sided)	J
E9107-4	The Consort Loadspaker	H
E9108-1	Pulsed Width Train Controller	E
E9108-2	Model Speed Controller – Main Board	F
E9108-3	Model Speed Controller – Power Supply	F
E9108-3	Model Speed Controller – Power Supply	F
E9109-1	Geiger Counter	E
E9109-2	Hemisync Waveform Generator Board	G
E9109-3	Hemisync Pulse Generator Board	F
E9109-4	Hemisync Power Supply Board	C
E9109-5	Nighfighter Main Processor Board	O
E9110-1	Freeze Alarm	E
E9110-2	Document Saver	E
E9110-3	Proto-type Designer	J
E9110-4	Nightfighter – Sound to Light (2 sided)	L
E9110-5	Nightfighter – Ramp Generator Board	F
E9110-6	Nightfighter – Cyclic Crossfade (2 sided)	M
E9110-7	Nightfighter – Strobe Board (2 sided)	J
E9110-8	Nightfighter – 8 Channel Triac Board	N
E9111-1	Digital Code Lock	L
E9111-2	Switched Mode Power Supply	E
E9111-3	Nightfighter Mode Selection (2 sided)	J
E9111-4	Nightfighter – Display Board (2 sided)	M
E9111-5	Nightfighter – Bass Beat Trigger (2 sided)	L
E9111-6	Nightfighter – Sequence Select (2 sided)	H
E9111-7	Nightfighter – Master Controller PSU	K
E9111-8	Nightfighter – Output Switch (2 sided)	M
E9112-1	Nightfighter Sensor Switch Master Control (2 sided)	L
E9112-2	Nightfighter Sensor Switch Channel Control (2 sided)	L
E9112-3	Nightfighter Sensor Switch Sound Trigger	H
E9112-4	Nightfighter Connector Board	F
E9112-5	Nightfighter Sensor Switch PSU	K
E9112-6	Nightfighter 8-Channel Input Interface (2 sided)	P
E9112-7	Power On and Overload Regulator	P
E9201-1	Laboratory Power Supply	F
E9201-2	Test Card Generator Board	M
E9201-3	LED Star (2 sided)	L
E9201-4	Enlarger Timer Main PCB (2 sided)	N
E9201-5	Enlarger Timer Selector Board (2 sided)	K
E9201-6	Enlarger Timer Switch PCB	E
E9203-1	MIDI Switcher – Main Board	L
E9203-2	MIDI Switcher – Power Suply	E
E9203-3	Sine Wave Generator (surface mount)	F
E9204-1	Auto Car Lights	F
E9205-1	Bat Detector	E
E9205-2	Pond Controller	F
E9206-FC	Stereo amplifier	G
E9206-2	Xenon flash trigger Main Board	J
E9206-3	Xenon flash trigger Flash Board	F
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E9207-2	Mini Baby Bug Monitor	C
E9207-3	Ultrasonic Audio Senmder (2 bds)	H
E9207-4	Camera Add-on unit (4 bds)	O
E9207-5	AutoMate 5V/48V Mkaer power supply	J
E9207-6	AutoMate Precision 17V power supply	J
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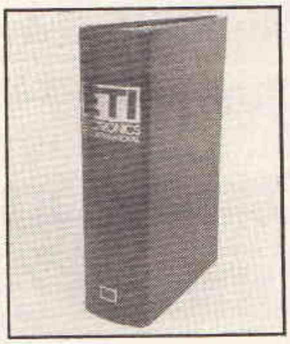

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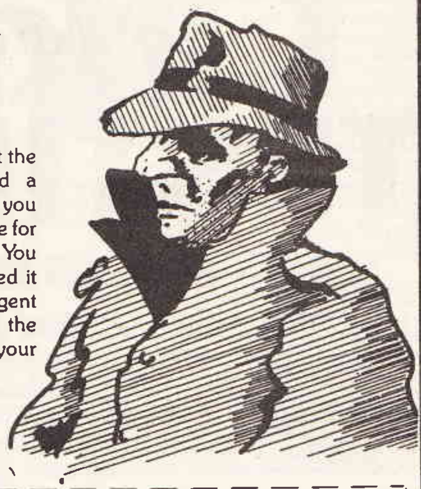

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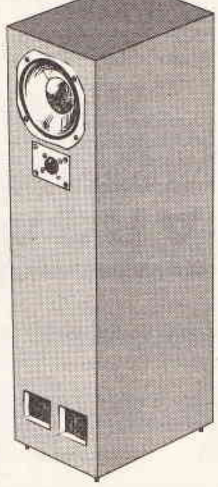

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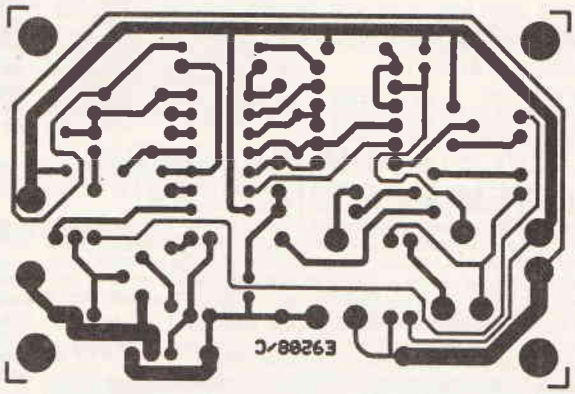
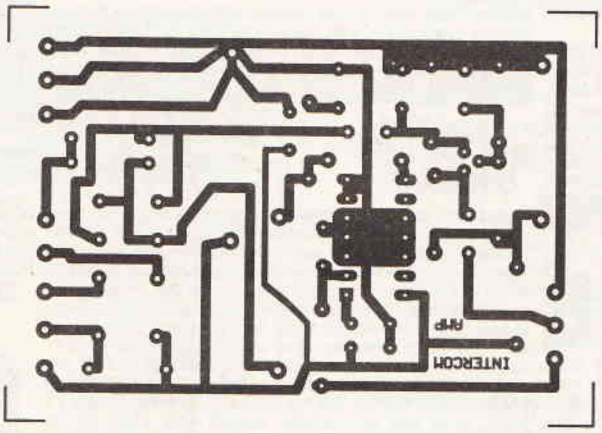
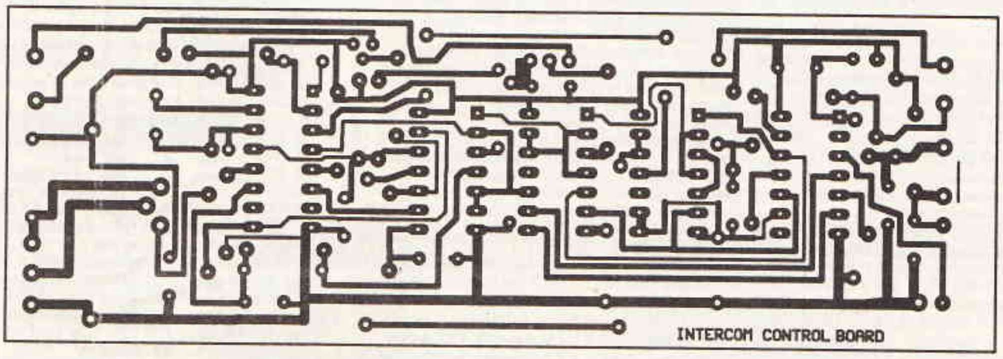


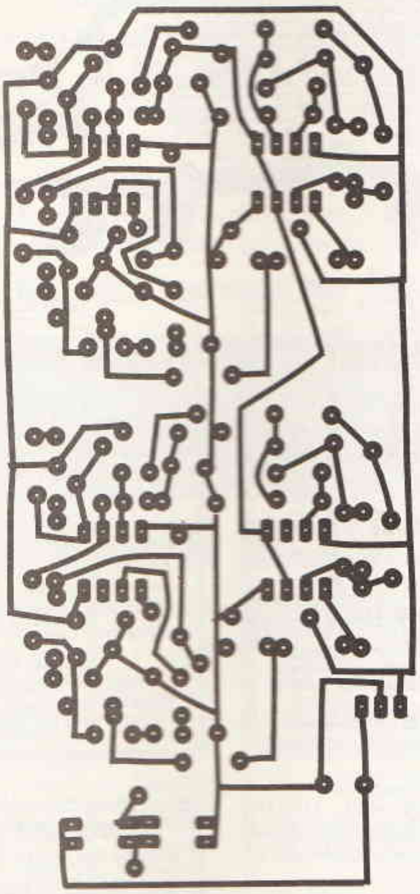
The Smart charger foil



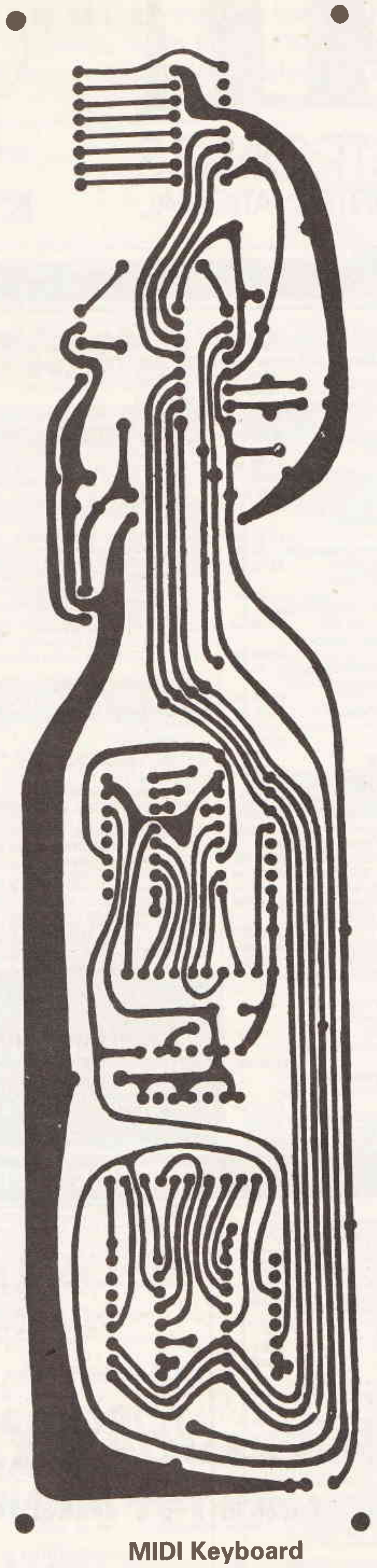
Touch Intercom (Amplifier)



Touch Intercom (Control)



Dynamic Noise Limiter



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The above articles are in preparation but circumstances may prevent publication

Last Month

Our July issue featured:

- Improved LED bike lamp
- Surround sound decoder
- Camera Add-on unit
- Mini baby monitor
- Ultra-sonic audio sender
- New concepts in optical connectivity
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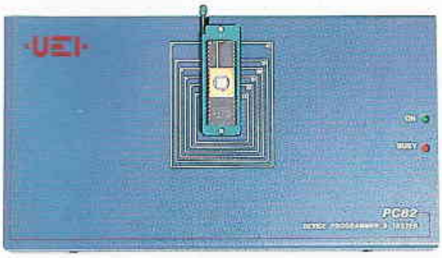

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