TEN PROJECT ISSUE

MEMORY SCOPE DISPLAY

SOUND SAMPLER

CYMBAL SYNTH

CHORUS UNIT

RHYTHM CHIP

AND MORE

S A W DEVICES

HOW TO DESIGN SIMPLE TRANSISTOR STAGES

WIN A 600 MHz FREQUENCY COUNTER FROM thandar

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DIGEST ...................................... 7
We write the news and the readers' digest.

READ/WRITE .................................. 13
Is this what they mean by mail order?

SAW DEVICES .................................. 14
Andy Armstrong was backward until he saw these devices.

AUTOMATIC TEST EQUIPMENT ................... 19
W.P. Bond gives us the benefit of his simulating thoughts.

DESIGNING TRANSISTOR STAGES .............. 22
All the world's a stage, says Les Sage, but some stages are transistorised.

WOMEN AND IT ................................ 25
Where are the women in the information technology industry? Anna Paczuska has the information.

PROJECTS

MEMORY SCOPE DISPLAY ...................... 28
Ian Marshall kicks over the traces in this project with go-slower stripes.

RHYTHM CHIP .................................. 33
Geoff Phillips takes his metronome to the PROMs.

MODULAR TEST GEAR ......................... 36
Mike Meakin's latest board generates waves, even away from the sea.

SWITCH-MODE REGULATOR .................... 40
How to get 1A at 5V steadily and without any ICs.

MILLIFARADOMETER ......................... 44
Time to measure those large capacitors with Ray Bold's meter.

CHORUS UNIT .................................. 48
All together now — Ian Coughlan's chorus unit matches his earlier noise gate.

ENLARGER EXPOSURE METER .................. 54
Get bigger and better pictures with Doug Bollen's darkroom assistant.

CYMBAL SYNTH ................................ 58
D. Stone brings you the clash, electronically.

THE SECOND LINE OF DEFENCE ................ 60
Vivian Capel alarms his house with this project but keeps his own peace of mind.

DIGITAL SOUND SAMPLER ..................... 63
Try out the first part of this sample project.

ETCETERA

TRAINS OF THOUGHT .......................... 70
Get on the right track with Roger Amos.

OPEN CHANNEL .............................. 70
Keith Brindley's new definition of TV.

ALF'S PUZZLE ................................. 71

INFORMATION

PCB SERVICE ................................ 7
see page 7

COMPETITION .............................. 12

NEXT MONTH'S ETI ....................... 65

FOIL PATTERNS ............................. 66

CLASSIFIED ADS ......................... 72

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Industry Hampered By Lack Of Qualified Staff

A shortage of qualified staff is the main constraint on the development and use of new technologies, according to a recent survey of top businessmen.

And while this problem affects most industrial countries, including Japan, British businessmen seem less ready to invest in research and development than do their foreign counterparts and see the benefits of new technology in narrower terms.

The poll was carried out by the management consultants PA Technology, and involved 176 chief executives and company directors from Britain, Japan, the United States, Germany and the Benelux countries. Twenty-two percent of those surveyed cited shortage of qualified staff as a major constrain while none of them felt that high interest rates or trade union activity were barriers to innovation.

The Japanese, it appears, are worried about both the quantity and the quality of the skilled staff they produce. Britain was not thought to have any problem with the quality of staff, particularly graduates, but was not felt to be making the most effective use of them.

On research, a third of all business interviewees interviewed outside the UK said that the emergence of new technologies had led them to invest more, whereas in the UK the figure was only 6%. More than half the British businessmen admitted to knowing little or nothing about the services offered by universities and other research institutions while their Japanese counterparts were well informed and highly satisfied with the effectiveness of their research bodies. The Japanese also placed more importance on pure research than any of their competitors.

None of the Japanese respondents mentioned technology as a means of reducing production costs or increasing production rates. By contrast, the Western countries all placed more importance on this factor.

Scanspeak Loudspeakers In Kit Form

Wilmslow Audio can now supply kits of parts to enable constructors to build any of the loudspeakers in the Scanspeak DIY range.

The kits are based on designs in The Scanspeak Handbook, which describes a number of loudspeaker systems which can be constructed using Scanspeak drive units. Among the features of these designs are the use of phase correction rings to space each drive unit the correct distance away from the front of the cabinet and the incorporation of a vented loading system.

The drive units are said to employ a symmetrical drive system which ensures that cone excursions in each direction are identical, removing the rise in impedance at high frequencies and thus simplifying crossover design, and they also use hexagonal cross-section wire in the voice coil. This is said to remove the tiny airgap that normally exists between windings, improving heat transfer and hence the power handling.

The cabinet panels are machined from Medium Density Fibre (MDF) board and Wilmslow can also supply a variety of iron-on wood veneer finishes. They will supply The Scanspeak Handbook free-of-charge.

Wilmslow Audio Ltd, 35-39 Church Street, Wilmslow, Cheshire SK9 1AS, tel 0625 - 52959.

Micro Chips Off The Old Block

Renault have cut in half the time taken to produce a life-size model of a new car by using a computer-controlled robot. An articulated arm cuts the required shape out of a giant block of plastic and the model can be ready for finishing and painting in just two days.

Traditionally, it took twelve to sixteen weeks to progress from the initial rough sketches of a new car body to the stage where a life-size, three dimensional model was ready. Part of the time was taken up in the production of 1/5th scale models to enable the three-dimensional effect to be appreciated before moving on to the full size model.

With modern computer aided design techniques, this intermediate stage is no longer required. The stylist can create an image on a touch sensitive screen and then rotate and manipulate it to assess the design and make any changes necessary. When everyone is satisfied, the program can be used to control the robot and life-size model quickly produced.

The process is being developed at Renault's Technical Centre at Rueil, near Paris. They claim that it has reduced the time required for the initial stages of the design process to between six and eight weeks, with consequent reductions in the cost. It also frees designers to work on more valuable tasks.

Renault UK Ltd, Western Avenue, London W3 1RZ, tel 01-992 3481.

ETI November 1985

ETI Printed Circuit Board Service

We are pleased to be able to announce that we have at last appointed a new supplier and will shortly be able to resume this service.

We expect to be able to start clearing the backlog of orders in the next week or two and would hope to have everything back to normal by the time the next issue comes out.

However, we would be grateful if readers who are thinking of ordering boards would hold on for a little longer and let us get things straightened out first. For this reason we have not printed the PCB Service page again this month but hope to include it, with any price changes occasioned by the change of supplier, in the December issue.

We would like to thank our readers for their forebearance during this difficult period.

7
**COLOUR AND MONOCHROME MONITOR SPECIALS**

**SYSTEM ALPHA '14' COLOUR MULTI INPUT MONITOR**
Made in the UK by the famous REDIRECTION Co. for their own professional computer system this monitor has all the features to suit your immediate and future monitor requirements. Two video inputs RGB and PAL Composite Video. Allow direct connection to the BBC and most other makes of micro computers and VCR's. An internal speaker and audio amplifier may be connected to your systems output or direct to a VHF monitor giving superior sound quality. Many other features included PAL tube, Matching BBC case colour, Major controls on front panel, Guarantee for 1 year against any defect of audio and video controls for Composite Video input, BNC plug for Composite video, 15 way D sub for RGB input, modulates construction etc.

This must be ONE of THE YEAR'S BEST BUYS
Supplied BRAND NEW and BOXED, complete with DATA and 90 day guarantee.
SUPPLIED BELOW ACTUAL COST—ONLY £149.00

**DECCA 60, 16 COLOUR monitor RGB input**
Little or hardly used devices. To offer this special converted DECCA RGB Colour TV Monitor at a super low price of only £99.00 not including VAT. This monitor has a DECCA 16" specification and special 16" high definition PAL tube, coupled with the tried and tested DECCA 60 series this gives 60 column definition and picture quality found only on monitors costing 3 TIMES OUR PRICE. In fact WE have sold one to a well known modeller and has not been returned since the day it was received. A new Bargain at £99.00. To be sold as is no returns. Guaranteed. Although used units are supplied in EXCELLENT condition. ONLY £99.00

**DECCA 60, 16 COLOUR monitor Composite video input**
Same as above. A new Bargain at £99.00. Ideal as a second amp for AMP VR or AUDI0 VISUAL use. USED ONLY £99.00. + £9.00 Dovetail MOUNTING FOR THIS MONITOR, £10.00 MORE IF YOU WANT A MAC70 AMP AND A MAC70 INTERFACE

**BUDGET RANGE EX EQUIPMENT MONOCHROME VIDEOMONITOR**
All are fully cased and set for 240V standard working with composite video inputs. Units are pre-tested and set up for 80 column use on BBC micro etc. Even DIAL TONE screen connections are included. Only £199.00. ($249.00 approx). If none of these are suitable the DECCA 60, 16 COLOUR monitor RGB input is the one to suit your needs. Always available in SHOWROOM and WAREHOUSE.

**BUDGET RANGE EX EQUIPMENT MONOCHROME VIDEOMONITOR**
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DIARY
Offshore Computers Conference & Exhibition — October 8-10th
Aberdeen Exhibition & Conference Centre. Conference devoted to the use of computers in petroleum exploration. Thirty-seven papers from seven countries will be presented and the costs range from £160.00 + VAT for one day to £190.00 + VAT for the full event. Entry to the association exhibition is free. Contact Offshore Exhibitions and Conferences Ltd, Rowe House, 55-59 Fife Road, Kingston upon Thames, Surrey KT1 1TA, tel 01-549 5831.

Technology Engineering Fair — October 8-11th
NEC, Birmingham. For details see August '85 ETI or contact Cahners at the address below.

Interpecon UK — October 10/11th
Metropole Hotel and Brighton Centre, Brighton. For details see August '85 ETI or contact Cahners at the address below.

Computer Graphics '85 — October 15-18th
Wembley Conference Centre, London. For details see August '85 ETI or contact Cahners at the address below.

Electronic Displays '85 — October 29-31st
Kensington Exhibition Centre, London. For details see September '85 ETI or contact Network Events at the address below.

Cellular Communications International — November 5-7th
Wembley Conference Centre, London. For details see September '85 ETI or contact Cahners at the address below.

London: venue to be announced. Course designed to enable programmers, engineers and technical managers to write, debug, test and execute Ada programs and interface between Ada and programs written in other codes. Participants ideally should be familiar with either Pascal, C, Fortran or PL/I. The cost is £675.00 + VAT and details are available from ICS at the address below.

Compec '85 — November 12-15th
Olympia, London. Exhibition of professional computer hardware and software. For details contact Reed Exhibitions, Surrey House, 1 Throwes Way, Sutton Surrey SM1 4QQ, tel 01-643 8040.

International Test And Measurement Exhibition (ITAME) — November 27-29th
Olympia 2, London. Exhibition and conference covering all areas of electronic test and measurement. For details see February '85 ETI or contact Network at the address below.

Satellite Communications — December 3/4th
Tara Hotel, London. Conference on which aims to cover all aspects of the subject including DBS, SMATV, interactive services, navigation, broadcasting, technology transfer, etc. The cost is £465.00. Contact Online at the address below.

The Which Computer? Show — January 14-17th 1986
NEC, Birmingham. Exhibition of business computers and word processors. For details contact Cahners at the address below.

Electronics in Oil and Gas — February 4-6th 1986
Bath, London. Exhibition and conference devoted to electronic equipment for surveying, drilling, pumping, processing and testing in the oil and gas industries. Contact Cahners at the address below.

Electronic Production Efficiency Exposition (EPEE) — March 11-13th 1986
Olympia, London. Exhibition and conference on computer aided design, manufacture and test (CAD/MAT) in the electronics industry. Contact Network at the address below.

Addresses:
Cahners Exhibition Ltd, Chatworth House, 59 London Road, Twickenham, Middlesex TW1 3SZ, tel 01-891 5051.
ILS Publishing Co (UK) Ltd, 3 Swan Court, Leatherhead, Surrey KT22 8AD, tel 0372-379211.
Network Events Ltd, Printers Mews, Market Hill, Buckingham MK18 1JX, tel 0280-815226.
Online Conferences Ltd, Pinner Green House, Asht Hill Drive, Pinner, Middlesex HA5 2AE, tel 01-868 4466.

Short Circuit Proof Transformers
A vel Lindberg have introduced a range of miniature printed circuit mounting transformers which are internally protected against short circuits. The transformers incorporate a positive temperature co-efficient thermal cut-out which breaks the primary circuit if an overload causes them to heat up. The advantage of this system then compared with conventional fuses is that the circuit will be restored as soon as the transformer has cooled sufficiently.
All the transformers in the range have single 240V, 50/60Hz primary windings and either one or two secondary windings. Those with single windings are rated at 1VA and are available in 6, 8, 9, 12, 15, 18 and 24V versions. The dual winding types are rated at 2VA and are available in 6, 7, 9, 12, 25 and 28V versions. The manufacturers claim that the windings can be connected in parallel as well as in series.
The overall dimensions of the transformers are 44 x 37mm x 33mm high and they are designed for direct soldering to a printed circuit board. Extra rigidity can be achieved by inserting self-tapping screws through the board and into holes moulded in the underside of the plastic case.
The transformers conform to BS415 class 2 and to the relevant IEC and VDE standards. For further details contact AveL-Lindberg Ltd, South Ockendon, Essex RM15 5TD, tel 0708-53444.

New Maplin Catalogue
Due out soon is the 1986 Buyer's Guide to Electronic Components from Maplin. It will contain details of a wide range of electronic components and, as our picture shows, carries the sort of cover illustration we have come to expect from Maplin. It will be available from high street newsagents for £1.45 or by post for £1.85 direct from Maplin Electronic Supplies Ltd, PO Box 3, Rayleigh, Essex SS6 8LR, tel 0702-552911.

O After 29 years, Quad have announced that they are ceasing production of their famous Electrostatic loudspeaker, the ESL. Some 54,000 have been built since they were first demonstrated in 1956 and Quad claim that just about every loudspeaker manufacturer, reviewer and serious recording studio has used them at one time or another. They were succeeded a few years ago by the ESL 63, but Quad say that parts for the original ESL will continue to be available until the year 2000.

O Cosser have published a 12-page full colour brochure which describes their capabilities in fibre optics and optical signal processing. Copies are available from the Publicity Department, Cosser Electronics Ltd, The Pinnacles, Harlow, Essex CM19 5BB, tel 0279-26862.
ACORN COMPUTER SYSTEMS
BBC Model B + Escrow ... £299 (a)
BBC Model B + DFS ... £335 (a)
BBC Model B + DFS + Escrow ... £349 (a)
BBC B Plus ... £300 (a)
UPGRADE KITS
DFS £30 (a)
Escrow Kit ... £25 (a)
1770 DFS Kit ... £48 (a)
ACORN ADD-ON PRODUCTS
Z80 2nd Processor ... £349 (a)
6801 2nd Processor ... £175 (a)
Teletext Adaptor ... £190 (b)
IEEE Interface ... £282 (a)
Microphone ... £133 (b)
PH Light pen ... £23 (c)
Torch UCN500 products incuding the IBM Compatible GRADUATION in stock
For detailed specifications on any of the BBC Firmware, Equipment listed here or information on our complete range please write to us...

ACORN COMPUTER SYSTEMS

BBC Firmware
Basic II ROM ... £22.50 (d)
View 12.1 Word Processor ROM ... £48.00 (c)
Worduser ... £40.00 (b)
BCP ROM/Disc ... £52.00 (b)
Disc Doctor/Generic Debug ROM £26 (d)
EMXTeach, ROM ... £39.00 (c)
Printermap (FX68/Graphic ... £23 ea (a)
ULTRACALC Spreadsheet ROM 150 cm ... £60.00 (b)
Double Vpp/SHOW VIEW ROM £50 (b)
ASCII/PASS/LOGO ... £50 (d)
Oxtec LOGICAL/FORTH/8888 ROM ... £43 ea (c)
ACCELERATOR (Basic Compiler) ... £57 (b)
COMMUNICATION ROM
Term II ... £26 (d)
Communicator ... £57 (b)

ACCESSORIES

Printers
EPSON
RX80T + £210 (a)
LX80 new NLQ printer £219 (a)
FX80T £315 (a)
FX80 £345 (a)
FX100 £430 (a)
JX80 4 Colour Printer £429 (a)
HiB Colour Printer £375 (a)
KAGA TAXAN
KP810 £235 (a)
KP910 £339 (a)

DAVIS WHEELS:
JUKI 6100 £285 (a)
BROTHER HR15 £315 (a)

PAPER:
200 Sheets Fanfold: 9.5" x 11" £12 (b)
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ACCESSORIES

Printers
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Serial Interface: 8140 £29 (c)
8145 with 2K £9 59 (b)
Paper Roll Holder £17 (d)
FX80 Tractor Attachment £25 (b)
Ribbons: RX/FX80 £85 (d)
FX/RX/MX100 £10 (d)
RX/FX80 Dust Cover £4.50 (d)

KAGA TAXAN
RS232 with 2K Buffer £85 (c)
KP810/910 Ribbon £60.00 (b)
JUKI 6100
RS232 with 2K Buffer £65 (b)
Ribbon £2.50 (b)
Tractor Attachment £29 (b)
Shiel Feeder £18 (c, d)
BBC Parallel Lead £7 (d)
Serial Lead £7 (d)

ACCESSORIES

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— All modems listed below are BT approved

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The ultimate world standard modem covers all BBC computer Standards (BELL standards outside the UK) up to 1200 Baud. Allows communication with virtually any computer anywhere in the world. The optional AUTO DIAL and AUTO ANSWER (with ample consideration for facsimile devices already provided on the market. Many powered £180. Auto Dial Board/Auto Answer Board £30 each. Contact us for details. Software £4.50.

NEW WESBURY SERIES
WS1000 V2120 (27 pin V22) £296 (a)
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Ask for details on these

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This pocket sized modem capable of V21 terminal is light weight and provides an excellent solution for communications between users, with mainframe computers and bulletin boards at a very economic cost. Battery or mains operated. Microprocessor based computer developed by BBC to modem data lead £1.

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FLOPPICLONE Box Head Cleaning Kit with 20 disposable cleaning discs ensures continued optimum performance of the drives £14.50 (c)
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Dual Disc Cable £8.50 (d)
10 Disc Cable Case £1.00 (b)
40 Disc Lockable Case £14 (c)
100 Disc Lockable Box £19 (c)

MONITORS

MICROVITEC 14" RGB:
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1451 Medium Resolution ... £240 (a)
1441 Hi Resolution ... £395 (a)
1431 AP Std Res PAL/AUDIO ... £205 (a)
1451 AP Med Res PAL/AUDIO ... £280 (a)
1451 DQ3 Med Res for QL ... £239 (a)
14514N ... IBM Personal RGB Monitor £425 (d)
MITSUBISHI 14" Med Res. IBM/BBC Compatible RGB ... £299 (b)
Above monitors are now available in plastic or metal cases.

KAGA Super Hi Res Vision III RGB ... £325 (a)
Hi Res Vision I ... £225 (a)
MONOCROME MONITORS 12":
Kaga Green KY1201 G Hi Res ... £99 (a)
Kaga Amber KX1201 A Hi Res ... £105 (a)
SOFTY G 128X128 512X512 RX HI Res ... £135 (a)
Swivel Stand for Kaga Monochrome ... £21 (c)

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2712-25 .......... £2.80
6264L-15 .......... £4.50

DISC DRIVES

These are fully cases and wired drives with slim line high quality mechanisms. Drives supplied with cables manuals and formatting disk suitable for the BBC computer. All 80 track drives are supplied with 40/80 track switching as standard. All drives can operate in single or dual density format.

Single Drives:
1 x 40KT OTS/5, T5000 ... £70 (b)
1 x 40K/40KT T5000 ... £805 (b)
Dual Drives:
With (integral printer)
Stacked Versions;
PD200 2 x 104K 40 KT SS ... £179 (a)
PD400 50K/40KT DS ... £235 (a)

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TS31 1 x 40K 80KT DS ... £98 (b)
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High quality discs that offer a reliable error free performance for life. Each disc is individually tested and guaranteed for life. Ten discs are supplied in a sturdy cardboard box.

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40 SS DD £22 (c)
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80 DS DD £28 (d)

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Single Disc Cable £8 (b)
Dual Disc Cable £8.50 (d)
10 Disc Cable Case £1.00 (c)
40 Disc Lockable Case £14 (c)
100 Disc Lockable Box £19 (c)

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RIBBON CABLE

(grammage)

50 60 70 80 90 100 110 120
Small $.18 $.20 $.21 $.22 $.23 $.24$.25 $.26
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DIL HEADERS

14 pin 40 pin 100 pin 110 pin 120 pin 135 pin 170 pin 210 pin
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4264 150ns Low Power 5.00 4.45 4.00
2716 450ns 5 volt 3.85 3.45 3.30
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Available now — The ROAM BOARD for the BBC Micro. Reads Rom's via a Low Insertion Force Socket and saves their contents as files, then reloads a file into its sideways Ram as required.

Full details on request.

74LS series TTL, wide stocks at low prices with DIY discounts starting at a mix of just 25 pieces. Write or 'phone for list.

Please add 50p post & packing to orders under £15 and VAT to total. Access orders by 'phone or mail welcome. Non-Military Government & Educational orders welcome. £15 minimum.

HAPPY MEMORIES (ETI),
Newchurch, Kington, Herefordshire, HR5 3QR.
Tel: (054 422) 618

WIN A THANDAR TF600 600 MHZ FREQUENCY COUNTER (PRE-SCALEABLE UP TO 1 GHz).
Runners-up win two PFM 200A 200 MHz frequency counters from Thandar.

1. He might have said ‘On yer bike’ with periodic passion. His name’s frequently used, though the cycle’s out of fashion. 2. If you need rectifying or a switching function, it’ll need to be biased when you get to this junction. 3. Harmonically speaking, this wave’s not all there — there aren’t any evens and the odds are all square. 4. Pure reactance exists only in theories. Real components add this is parallel or series. 5. Classes A, B and C are defined with some rigour. And you shouldn’t complain if they don’t make things bigger. 6. ‘Here’s the pitch,’ said the prof, feeling alright. ‘The fact is, there ain’t one when this sound is white.’ 7. ‘Lie back and relax,’ said the girl to her sisters. ‘You’ve been behaving too much like ICs and transistors.’

POST TO: ETI (TC), 1 GOLDEN SQUARE, LONDON W1R 2AB.

1. 2. 3. 4. 5. 6. 7.

The initials of the answers form the word: I would like a Thandar frequency meter because

Name:
Address:
Signed:

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Colleges, Universities, Individuals: Build your own modular Z80A-based metal 19" rack and card Interak computer. Uses commonly available chips — not a single ULA in sight (and proud of it). If you can get your own parts (but we can supply if you can’t!) all you need from us are the bare p.c.b.s and the manuals.

Interak 1’s greatest asset — space for expansion.

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No fears about this one going obsolete — now in its fifth successful year! Send us your name and address with a 21p stamp and we’ll send you 40 pages of details (forget the stamp if you can’t afford it). You’ve already got a plastic computer for playing games, now build a metal one to do some real work: Interak, Interak, Interak!

Greenbank
Greenbank Electronics (Dept T11E), 92 New Chester Road, New Ferry, Wirral, Merseyside L62 5AG
Telephone: 051-645 3391

WIN A THANDAR TF600 600 MHZ FREQUENCY COUNTER (PRE-SCALEABLE UP TO 1 GHz).
Runners-up win two PFM 200A 200 MHz frequency counters from Thandar.

SUPER COMPETITION

In conjunction with Thandar Electronics, ETI is offering three superb frequency meters as prizes in this simple to enter competition.

All you have to do is correctly answer the questions below, which are given in the form of doggerel rhyme. The answers are all electronics terms whose initial letters can be combined to form a seven letter word in the spaces provided. Then complete the phrase underneath in no more than 12 words. This will be used as a tie-breaker. The closing date for entries is the last post, November 7th, 1985.

The first prize will be awarded to the person who gets all the answers correct and, in the event of a tie, provides the wittiest phrase. The two second prizes will be awarded to the runners-up. If no completely correct answers are received, the prizes will be awarded on a points basis — 2 points for each correctly answered quiz question and one point for providing the correct seven-letter word. Employees of Argus Specialist Publications and Thandar Electronics or their relatives are not eligible for the competition. The competition judges will be Gary Herman (ETI) and Tony Starling (Thandar) and the results will be announced in the January 1986 issue of ETI, on sale on December 6th, 1985.

ETI NOVEMBER 1985

www.americanradiohistory.com
Silly Old Buffer

Dear ETI,

I feel that I must object to your inclusion of a particular addendum in the service sheet section of the September issue. It concerns the use of the TMS4416 dynamic RAMS in my printer buffer project which was published in the July and August issues.

While it is true that the Z80 will only provide seven bits of refresh, it is possible to use software to simulate the use of all eight bits. Two methods are commonly used:
1) Regularly toggle bit 7 of the refresh register.
2) Execute continuous op-codes located between XX80 and XXF6 in the memory map, at least once every 4mS.

Either of these methods will provide satisfactory refresh for the TMS4416 and the second is, in fact, incorporated in the design of the printer buffer. The use of the Hitachi IC mentioned is definitely not required.

I did not mention this aspect of the refresh mechanism in the original article, but I do feel I should have been consulted before publication of this erroneous addendum. My thanks for this opportunity to set the record straight.

You faithfully,
Nick Sawyer,

Literary Sources

Dear Sir,

I was not surprised to learn from a letter in your magazine that Motorola are not very helpful (Read/Write, ETI September 1985). I also have had no reply from them when I asked for some literature. But I wrote to STC Electronics Services, Edinburgh Way, Harlow, Essex CM20 2DF who were very helpful and sent about twelve data sheets, all Motorola. I can also recommend Gothic Crellon, 380 Bath Road, Slough, Berkshire SL1 6JE who have been very helpful.

Yours faithfully
P.D. Yates
Chelmsford.

Goth, it's easy To C that Motorola aren't terribly interested in the home constructor. — Ed.

Fine Print

Dear Editor,

I thought I'd drop you this line after reading my August copy of ETI.

I use a BBC-B computer, Wordwise and an EX42 typewriter as a printer. The interface allowing me the printer mode came from ETI October 1983. I have had no trouble with it since setting it up.

As a writer my unit is in constant use and not to print listings! Over the last couple of days I have down loaded around twenty thousand words to the EX42 . . . no problem.

The only complaint I have about the system is speed. I find 10cps too slow for my needs so I will soon be buying a printer. As the system stands at the moment it is completely trouble free, so I have no complaints with it or the interface. Thus it did surprise me a little to see in the August issue an 'EX42 Interface for the BBCB'.

When using Wordwise all one need do to get 'CT' is to direct command 'FX6' when in the WW edit mode.

This letter is produced using Wordwise, the EX42 and your interface. . . .OK!

Yours sincerely
J.M. Stevenson
Brentwood

First someone calls us 'reputable', now a reader says he's actually persuaded one of our projects to work! Where will it all end?

The original interface was designed for use with the Microtan 65, but many of our readers wanted to use it with a BBC or other microcomputer. It is quite possible to re-write the software to provide the necessary control signals for other machines, but apparently some people aren't quite sure how to go about it. Hence the dedicated BBC B version. However, it's good to hear that someone has got the original version to work with a Beeb. — Ed. ETI

ETI NOVEMBER 1985

AUNTIE STATIC'S PROBLEM CORNER

Dear Auntie,

I have often wondered why 2's complement arithmetic works in binary but not with ordinary numbers. I don't expect you will know the answer.

S. Lister
Salisbury.

Since you have the bloomin' cheek to question Auntie's omniscience, I don't think you deserve an answer at all. However, being a kind and understanding Auntie and since I don't get paid unless I answer a question or two, I'll overlook it just this once.

Two's complement arithmetic obviously doesn't work for decimals because you can't form the 2's complement of a decimal (ordinary?) number. However, you can form the 10's complement by subtracting each digit of the number from 9 and then adding 1 to the final result, so —1111 would become 8889, —2634 would become 7366, and so on.

Suppose you are faced with the hopeless prospect of doing a subtraction having just discovered that your calculator batteries are flat.

1756 — 1186 — ???

You don't want to get involved with all the nursery school business of borrowing and such like, so what do you do? Faint? Break out in a cold sweat? Go and change your library book? No! With dignity and poise you scribble all over the second number and replace it with its 10's complement equivalent so you are left with:

7342 + 5814 = ???

This number is near as complicated as having to do a subtraction, and with your family and friends looking on in admiration you skillfully add the numbers together to give 13156.

At this point you realise that your answer can't possibly be right because its bigger than the number you started with. Don't panic! Just keep your thumb over the initial 1 and show your answer with a flourish: 3156. When you get used to this method you can avoid getting an inky thumb by starting your addition as close as possible to the left hand side of the paper. The extra 1 will then overflow onto the tens column and you can quickly slide your paper over to cover it up.

The only thing to remember is that if you are faced with something like 3733737 — 66, where the numbers are not both the same length, you must put a load of 9s in front of your 10's complement number until they are the same length. In this case you will end up with: 27272727 — 99999934 = THUMB3737671 (by the beginners' thumb method). Oh yes, and if the number you are subtracting is bigger than the one you're taking it away from, the answer will be in 10's complement form. For more advanced students there are also Chinese subtractions, or takeaways, but these are outside the scope of this reply. — Auntie.
SURFACE ACOUSTIC WAVES

Does your knowledge of SAW devices need sharpening up? Andy Armstrong looks at the cutting edge of the new technology.

Filters, resonators and delay lines are essential elements in a wide range of electronic equipment and a great deal of time and effort has been devoted to improving them, either by perfecting existing techniques or developing new ones. Filters and resonators have traditionally relied upon the frequency-selective properties of reactive networks and much has been written about the best ways of combining reactive components (capacitors, inductors and crystals) to obtain the desired results. Delays produced by passing the signal from a transmitting transducer to a receiving transducer and placing between them some substance which slows down the passage of the wave. Examples range from the glass block and piezo-electric transducers used in television receivers to the coil-spring and magnetic transducers used in some audio effects units.

These techniques remained largely unchanged until the development of the field effect transistor (FET) with its excellent switching characteristics, and the introduction of integrated circuitry which made it possible to build circuits of previously unimaginable complexity. The devices which resulted were the switched capacitor filter and the bucket brigade delay line, or charge coupled device (CCD), both of which use linear circuit elements controlled by electronic switching. The advantages in each case include greater flexibility; the break frequency of the filter and the delay period of the bucket brigade device can both be altered simply by changing the frequency of the external clock.

The development of complex digital integrated circuits has taken this flexibility several stages further. It is now possible to convert an analogue signal into digital form and then carry out a wide range of filtering operations, if necessary under software control so that the parameters can readily be changed to meet specific requirements. Delay can be obtained simply by using readily-available memory chips, allowing longer periods to be achieved without attendant degradation of the signal. Unlike linear systems, in which phase and frequency are heavily interdependent and it is not easy to design for one without affecting the other, digital filtering allows many different parameters to be specified to a high level of accuracy.

The last few years have seen the emergence of another new signal processing technology which combines many of the advantages of linear and digital techniques. Surface Acoustic Wave (SAW) devices use a tiny transducer to launch a wave across the surface of a piezo-electric crystal and another transducer to pick the wave up again. In this respect they are not unlike conventional glass delay lines, but the crucial difference is that the behaviour of the wave can be modified by the use of complex transducers which work, in effect, as multiple taps on the delay line. The result is a filter whose characteristics can be controlled almost as well as can those of a digital system but which can be produced at a fraction of the cost.

The Nature Of The Wave

Before going any further, we shall look at what a surface acoustic wave actually is and how they are generated.

The easiest way to understand what is involved is to consider a very simple crystal structure — a cubic lattice with identical atoms at each vertex. An ordinary wave set up in the structure simply involves compression and rarefaction of the atoms forming the structure. To put it crudely, one could imagine the little model atoms bouncing back and forth on their springs. The atoms are fairly stiﬄy supported, being anchored by six bonds. This type of wave is called a bulk wave, and is of the type generated in conventional glass delay lines.

A surface wave, on the other hand, involves atoms on the surface swinging about in an arc rather than a straight line. Atoms on a face of the crystal are less firmly anchored, with only five bonds attached, so the

A SAW convolver manufactured by Siemens. The lithium niobate SAW chip is the thin, dark strip in the lower half of the case and the reminder of the area is occupied with matching elements.
propagation speed of a surface wave is lower than that of a bulk wave. This means that a surface wave is confined to the surface, and does not tend to excite bulk waves. It is possible to launch bulk waves when trying to launch surface waves, and vice-versa, but they are different entities and there is limited interchange between the two. The rate of propagation of these acoustic waves is very much less than that of electromagnetic waves, so a relatively long delay line may be made in a small space.

Surface acoustic wave devices are made on crystals of particular types, the main ones being quartz, lithium niobate, and lithium tantalate, all of which are piezoelectric crystals. This is essential, because the transducers which launch the wave consist of interleaved strips or fingers of metallisation which are energised with an electrical signal to excite a piezoelectric response and hence launch a wave.

Figure 1 shows the layout of a typical SAW filter. The transducers are symmetrical, so acoustic waves are radiated in both directions. The acoustic absorber on the far side is there to prevent the reverse wave from either propagating around the crystal or being reflected from the edge. Either event would interfere mightily with the intended function of the device. Similar reasoning applies to the presence of the absorber behind the receiving transducer.

In this simple example, the waves launched by each pair of fingers in the transducer may either reinforce or cancel the waves from other pairs, depending on the acoustic wavelength. The even spacing of the fingers shown would give rise to a flat-topped filter response.

LC filters can provide a good approximation to any reasonable amplitude response characteristic, but the resulting phase response is likely to be less than ideal. An enormous complexity is required to achieve a particular phase response as well, and it is normally quite impractical. A SAW filter, on the other hand, can meet this kind of requirement because phase and amplitude response may be determined separately.

The example in Figure 1 shows a transducer which has evenly spaced, fully overlapping fingers. The spacing may be varied to adjust the phase response, and the degree of overlap may be varied to control the intensity of the wave. Figure 2 shows a Sin(x)/x transducer which will give a bandpass response showing some resemblance to the square response beloved of textbook writers. Real transducer patterns have many more fingers, of course.

The method used to design SAW filters is clearly quite unlike that used for LC filters. SAW filters are normally of a type described as transversal filters, which is a fancy way to describe the process of producing a desired output by adding many signals tapped from a delay line. CCD filters also work on this principle, and another technology which lends itself to this method is digital filtering.

It is interesting to compare digital and SAW devices when used for signal processing rather than filtering. In digital terms, the accuracy of a SAW device would be perhaps five bits, but the processing speed is two orders of magnitude faster than that of any digital IC currently in use or under development. There are applications which use digital and SAW technologies together to give the best of both worlds, but such applications are a specialised topic on their own.

Real Devices

So far, all that I have said about SAW devices sounds almost too good to be true, but there are some drawbacks. A look at some device configurations will help us understand the problems and their effects on the functioning of SAW devices.

One annoying phenomenon is signal coupling via bulk waves. While the method of excitation is designed to produce surface waves, bulk waves can be launched at frequencies outside the passband. If a bulk wave is launched at a particular angle to the surface of the crystal, the wavelength at which the signals from the transducer fingers are in phase and will produce a strong wave is different from that for surface waves. If this wave reaches the receiving transducer, a response out of the normal passband will be generated. The angle between the direction of propagation of this wave and the plane of the transducers affects things in much the same way as coloured light is split by a diffraction grating. This effect is illustrated in Fig. 3.

To cause problems, the bulk wave launched by the transmitting transducer must be reflected back from the bottom face of the crystal to the receiving transducer. This effect can be minimised by roughening the bottom face of the crystal, making it as
thin as possible, and coating the bottom with a sound absorbent glue.

A far more effective measure is to offset the transducers and couple energy between them by means of a multistrip coupler, as shown in the photograph. This idea came from the RSRE (Royal Signals Research Establishment) and is now widely used the world over. In simple terms, the coupler acts both as a receiving transducer at one end and as a transmitting transducer at the other.

Another important advantage of this coupling technique is that both transmitting and receiving transducers may be weighted to give a particular response — a process known as apodising — without an unmanageable design problem. If two apodised transducers are not offset but used directly in line, to work out the response one has to consider the effect of the wave launched by each pair of fingers in the transmitting transducer on the receiving transducer as a whole, and the effect of the overall wave pattern launched by the transmitting transducer on each pair of receiving transducer fingers. This is equivalent to the mathematical process of convolution, which is rather complicated. Offsetting the transducers simplifies things considerably.

**Driving Techniques**

To use SAW filters it is necessary to couple signals to and from them, which is not made easier by the complex impedances of the transducers. To take a practical example, the Signal Technology BP1102 70MHz bandpass filter (2MHz bandwidth) is quoted as having input and output impedances of 6R in series with 10p. At 70 MHz, 10p has an impedance of 227R, which makes a series inductor to tune out the capacitance very desirable.

This is only the start of the problems. The BP1102 is quoted as having a 24dB insertion loss with a matching inductor. Without one it has a midband attenuation of 44dB. Unfortunately, even at 2MHz bandwidth, the use of a matching inductor can cause response ripple due to an effect called triple transit, in which the acoustic wave bounces from the receiving transducer back to the transmitting transducer and then back to the receiving transducer again.

Acoustic absorbers are used to minimise reflections from the end of the crystal, but the reflections in question here are from the actual transducers. They occur whenever there is a significant electrical load coupled to the transducers because power is drawn from any incident wave and dissipated. This inevitably disturbs the wave and causes reflections.

The transducer is in fact a three port network with a matched load on one of the ports. If an extra load is added on the electrical port, the mechanical impedance of the acoustic wave no longer matches.

One solution to this dilemma is to match the input source to the transmitting transducer, and use a high impedance input amplifier on the receiving transducer (the output of the SAW device). This means that there is little reflection from the receiving end and the effects of triple transit are small. Another solution is to drive the input of the device from a high impedance and with a high voltage signal, either by means of a step up transformer or a cascade amplifier with a high voltage power supply. This delivers plenty of power to the input while presenting an electrical mismatch.

![A Siemens OFW 361 SAW filter shown without its case. Note the offset of the transducers and the use of a multistrip coupler between them. The graph in the background illustrates the performance of the filter: the centre frequency is 36.5 MHz and the scale divisions represent 10dB on the vertical axis and 2 MHz on the horizontal axis.](image)

**Material Characteristics**

The BP1102 is made on a quartz substrate, ST cut (which is similar to the more familiar AT cut used for oscillator crystals), and has the advantage of almost zero temperature coefficient. In general, a given material can only offer a bandwidth which is a certain percentage of the operating frequency before the insertion loss becomes too high: for quartz this bandwidth is only a few percent.

One of the wider bandwidth filters from the same family has a 9MHz bandwidth at 72MHz, and is made on lithium niobate. It boats an insertion loss of 24dB without any electrical matching at all, but it has a temperature coefficient of about 90 PPM/degree centigrade which is a characteristic of the material. The percentage bandwidth available with lithium niobate and lithium tantalate is about 40%, which might suggest that they are suitable for use as an input "roofing filter" in, for example, television receivers. Such a filter could provide a flat response over the entire television broadcast band, and about 60dB rejection outside it.

The drawback is that the loss is much too great, and the resultant noise figure would render reception
in any but the strongest signal areas impossible. Development work aimed at low loss devices is being carried out and there are several angles from which the problem can be attacked. One of the obvious ones is to develop a transmitting transducer which launches all its energy in one direction, and a receiving one which picks up almost all the energy flowing past it.

The normal types of transducer launches the same power in each direction, so one half of the signal actually transferred to the substrate is wasted in the acoustic absorber. By the same token (because of reciprocity) the receiving transducer can pick up only one half of the incident signal power. This immediately gives a 12dB loss, even if every other aspect of the device is free of losses.

A unidirectional characteristic can be achieved by clever phasing which cancels the wave in one direction. It is difficult to prevent unwanted reflections or other response ripples when designing this type of transducer, which is why it is only used where low loss is important.

Other SAW Applications

So far we have concentrated mainly on transversal filters, but this is not the only application of SAW devices. It is also possible to produce resonators which are similar in principle to microwave resonator cavities, and if two of these are coupled a narrow bandpass filter with low insertion loss may be made. The structure of a single resonator is illustrated in Fig. 4.

The diagram shows a transducer between two reflectors. The reflectors consist of many strips, suitably phased for the frequency of operation, rather than one thick strip. If too heavy a strip is placed in the path of the wave, the mechanical impedance mismatch causes bulk waves to be launched. As it stands, this type of resonator may be used as the frequency determining element in an oscillator.

It may also be coupled electrically to another similar resonator, in which case a bandpass response can be generated similar to the response given by coupling two ordinary tuned circuits. Alternatively, two resonators may be fabricated on the same substrate and the reflector between them may be made shorter to allow coupling. The response is similar to that obtained with electrical coupling but only one substrate is used, making it preferable for large volume applications. The device layout and energy distribution is illustrated in Fig. 5. Little of the energy coupled into the input resonator is wasted, so a narrow bandwidth low insertion loss device may be made.

Another interesting SAW application not so far mentioned is the convolver, a device which carries out the mathematical process of convolution in real time on two input waveforms. Effectively it multiplies the two waveforms and integrates the result. In order for the multiplication to take place, the signal level must be high enough for non-linear to occur. To achieve this, wide transducers are used to launch the waves, and they are then focussed to a narrow beam by a metalisation structure which looks like a lens. This is logical when you consider that a glass lens bends light because the light is slowed down by the glass, and that metalisation on the SAW substrate has the same effect.

To maintain a focussed beam, a metalised track is deposited. This works in a way closely analogous to that of an optical fibre in guiding the wave. The piezoelectric effect generates a voltage in this strip, which is integrated by its capacitance.

The Future

It is likely that developments in low insertion loss devices will continue, and will result in their extensive use in receiver input stages. We may eventually achieve a low enough loss to permit the use of SAW devices in input roofing filters for television reception, covering the range 40 to 800 MHz. The technology is already at the stage where an input filter having 2MHz bandwidth at a centre frequency of 900MHz and an insertion loss of 2dB is becoming practical. This would be ideal, for example, for cordless telephone applications.

If satellite TV ever turns out to be as major a development as it is cracked up to be, there will be a strong incentive for developments such as the use of a SAW resonator to make a quadrature detector working at 1GHz. Along with a specially developed IC for the job, it is easy to imagine that a TV receiver head about the size of a matchbox could be developed, giving out digital signals for processing in the digital television sets now becoming available.

It is quite likely that some of the signal processing applications of SAW devices, for example in chirp radar, will gradually be taken over by higher and higher speed digital processors, but there is likely to be more than enough expansion in other applications to offset this for the foreseeable future.
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AUTOMATIC TEST EQUIPMENT

Things will never be the same again, now that simulation testing is here. W.P. Bond elucidATES.

The traditional approach to functional testing of digital circuits has been to use simulation techniques. The principle is simply to compare the real board or module (the unit under test or UUT) to a computer model. Input test patterns — in the form of logic highs and lows — are applied to the model and to the UUT. The computer monitors the response of the UUT and verifies it against results obtained from the model.

There are several simulator packages available. The one described here is used by Hewlett-Packard on their DTS (Digital Test Station) and is based on the Eichelberger simulator. It is table-driven, which simply means that the circuit description is held in tables. Other features of this simulator are that it is capable of handling three states (high, low and x, or unknown, states); it is fault-inserting (that is, it can simulate faults itself); and it assumes zero propagation delay. This latter may be desirable because physical circuits display different propagation delays across the board and from board to board. Zero delay is a simplifying assumption. To ensure adequate testing with the zero delay assumption, the H-P simulator introduces a variable delay before the ATE comparators strobe the UUT outputs after a test pattern has been applied. The default value of this ‘safeguard’ delay is 4 microseconds.

This Year’s Model

All chips are simulated by reference to a set of primitive model functions: AND, NAND (or NOR), XOR, AMP, INV, ROM, RAM and DELAY. For convenience, a disc-based library is kept of most standard TTL and CMOS devices, which can be called upon for use by the model and can also be updated. To construct a model circuit, a source ‘Topology’ file is developed. This is a description of the board to be tested, including all devices on the board, signals (or vectors) and input and output connections. The circuit in Fig. 15 was used to illustrate test points last month (Fig. 3, p.43, ETI October 1985). The following listing shows how a model is made up for this circuit:

- **Headers** — board name, code, etc.
- **Libraries** — TTL1, TTL2 (which call up the required tables).
- **Networks** — used to build models of specialised chips not already included in the standard library.
- **Main board** (5) $G1\ INP1.1 \ INP3.5 \ G2-2.(2,4) \ G1-3.3 \ G1-6.6 \ G2 \ INP2.1 \ G2-2.2 \ PL1 \ INP1.1 \ INP2.2 \ INP3.3 \ G1-3.4 \ G1-6.5 \**
- **Outputs** $G1(7408) \ G2(7404) \** — types of gate used, for library reference.
- **Inputs** $G1 \ INP1.1 \ INP3.5 \ G2-2.(2,4) \ G1-3.3 \ G1-6.6 \ G2 \ INP2.1 \ G2-2.2 \ PL1 \ INP1.1 \ INP2.2 \ INP3.3 \ G1-3.4 \ G1-6.5 \**
- **Gates** $G1 \ INP1.1 \ INP3.5 \ G2-2.(2,4) \ G1-3.3 \ G1-6.6 \ G2 \ INP2.1 \ G2-2.2 \ PL1 \ INP1.1 \ INP2.2 \ INP3.3 \ G1-3.4 \ G1-6.5 \**
- **Signals** $G1 \ INP1 \ PL1-1 \ G1-1 \ INP2 \ PL1-2 \ G2-1 \ INP3 \ PL1-3 \ G1-5 \ G2-2 \ G2-2 \ G1-2 \ PL1-4 \ G1-6 \ G1-6 \ PL1-5 \**
- **END**

The sections of this outline program labelled Headers, Libraries and Networks are self-explanatory. Main board indicates the total number of input/output pins used — in this case, five. The dollar sign, $, is simply a delimiter used to indicate the end of a data-block.

Inputs, Outputs and Packages give further details of the hardware — input pins, output pins and device types.

The section labelled Gates gives a signal name for each pin used in each gate — INP2, for example. Where a conventional name has not been used, the name is derived from the gate and its output pins — G1-3, for example. Notice that the connector, PL1, is included as a gate, and can be given the label ‘CONN’, which is a primitive.

The section labelled Signals is optional and can be used to cross-check the Gates section for extra reliability. In the Gates section, the top line shows that pins 1, 5, 2 and 4 together, 3 and 6 of gate G1 are connected to signals INP1, INP3, G2-2, G1-3 and G1-6, respectively. In the Signals section, the top line

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Fig. 15 Deriving labels for a model of a simple circuit.

ETI NOVEMBER 1985
shows that signal INP1 runs between pin 1 of PL1 and pin 1 of G1. Note that the ★Signals section includes information on the outputs from PL1 (PL1-4 and PL1-5).

★END indicates the end of the model-program.

Under certain conditions, complex devices can behave curiously. Usually, such conditions are rare, but models should take account of all possible operating conditions. Those that don’t are described as ‘optimistic’, while models which take even the most unusual and unexpected conditions into account are known as ‘pessimistic’.

The Program Guide

When the model is finished and the source file created, several utility programs are run on it. The first — called SGLST — checks syntax and cross-checks the ★Signal and ★Gates section, where both are used. It also translates the user Topology file into an actual software model.

A second program — SMSET — expands the software model generated by SGLST by breaking it down into primitive elements and interconnections. It also generates a listing of numbers assigned to the I/O signals.

At this stage, the board’s response to applied patterns can begin to be simulated. The patterns can be applied either manually or by means of ‘Automatic Pattern Generation’ (APG). (APG is sometimes expanded as ‘Automatic Program Generation’, when it refers to the whole process described here, up to and including pattern generation.)

With ‘manual patterns’, the desired inputs are set up by the user and the board’s behaviour is simulated by the model. Another program then monitors the response of the model. This enables the user to look at any particular device or signal on the model and assess its condition.

APG is a fairly straightforward technique to understand, and the results are comprehensive. But applying it to find or investigate faults can be very tricky.

Take the simple NAND gate in Fig. 16 as an example. The tri-state logic truth table (showing combinations of 0,1 and x-states) demonstrates that only in some cases does an x-state on an input produce an x-state at the output. Whenever an x-state propagates through a gate, the gate is said to be ‘sensitive’. The two input NAND gate is sensitive when one input is at a 1. Similarly, a two-input NOR gate is sensitive when one input is at a 0.

Deriving a suitable sequence of test patterns for APG is partly a matter of developing a simple algorithm to cover all possibilities and partly a matter of selecting courses of action when certain conditions apply in order to maximize the test procedure. The selection is done on a rule-of-thumb (or heuristic) basis — which is to say, by following courses of action which ‘seem to be working best’. APG can be costly in terms of the number of patterns used to track down all possible faults especially if the wrong heuristics are applied. APG is an effective and useful technique if it is only required to test for the presence or absence of faults. If you want to track down particular faults then the use of manual patterns (relying, as they do, on the tester’s experience and intuition) will probably be more fruitful.

A Catastrophic Failure

In a digital circuit, a fault is considered to be any physical defect which causes the circuit to respond abnormally. The sort of simulation-based ATE dealt with so far responds to any such faults, but the tests are static only and the faults they can uncover are generally described as catastrophic.
To say that the testing procedures are static is actually to say that the rate at which patterns are applied to UUT is much less than the propagation times of the devices connected in the UUT. Between the application of any two patterns the UUT is allowed to settle in to a stable condition. Dynamic and time related faults are not uncovered by static signal tests and, in general, the fault coverage is restricted to failures due to missing parts, incorrect installation of parts and the partial or complete death of parts. Such catastrophic faults are the commonest kind, especially in production runs of fully de-bugged circuits, but they are by no means the only or the most frustrating kind.

In essence, dynamic testing follows the same course as static testing with automatic pattern generation, but at far greater speeds, usually at data rates greater than 1 MHz up to the full system speed of the UUT (at which point component dynamic testing becomes dynamic real-time testing). Dynamic faults vary with the level of integration of the UUT. For gates, the faults are usually associated with propagation delays. Latches display dynamic faults connected with set-up and hold time. Memories are associated with data addressing faults and microprocessors with cycle-time stability faults. We'll be dealing with such dynamic faults at greater length in a future article, but for the moment it's worth noting that dynamic faults can be tested by simulation-based equipment by means of some fairly elaborate modelling techniques.

A typical simulator has been developed by Brunel University for Smiths Industries and goes by the name 'HILO'. The 'LO' of 'HILO' refers to the low-level modelling already described, while the 'HI' indicates that the simulator includes a high-level language so that programmable propagation delays can be set up for individual devices and allowing functional modelling of complex VLSI devices which do not, therefore, need to be broken down into primitive gates.
DESIGNING TRANSISTOR STAGES

Les Sage gives a general introduction to transistor amplification stages and some rules of thumb for those who want to design their own.

One of the main problems facing an engineer in designing an amplifier is the choice of circuit to suit the particular application. Many engineers, unaware of the advantages and disadvantages of different circuits, play safe and go straight for an IC op-amp, widely accepted as a near perfect black box amplifier with high input impedance, low output impedance and an infinite open loop gain which, if biased correctly, is sure to work better than a hastily designed transistor stage.

The op-amp has many uses, some of which are beyond the scope of simple discrete transistor circuitry. But there are many applications for which a discrete one or two transistor amplifier will do the job better and cheaper. An IC is only a box of transistors which a designer has put together as a circuit block, usually with a view to as many applications as possible. Many areas of design are often compromised in the process of IC design.

IC op-amps are often deficient in one or more of these areas:

1) Very low noise amplification (even today the lowest noise amplifiers are still built with discretes);
2) High frequency amplification (op-amps are so stuffed full of transistors their transition frequencies are best measured in weeks rather than nanoseconds!);
3) Distortion (the majority of op-amps have class B output stages and hence generate crossover distortion which is undesirable, especially in low level pre-amplifiers);
4) High power output (amplifiers above 20W are generally discrete circuits).

The aim of this series is to demonstrate some useful amplifier circuit building blocks using transistors. Performance is discussed — and ways of improving it — and various two transistor circuits are shown for situations where one transistor is just not good enough.

The Common Common Emitter

In order to design a single stage amplifier (for example, Fig. 1), its purpose must first be established. Let's assume that we're designing a circuit to boost the output level of a tape-recorder phono socket to drive a power amplifier.

The tape-recorder gives an output level of 100mV RMS from a 1k0 source impedance. The power amplifier, however, requires 1V RMS and has an input impedance of 50k ohms.

Having opted for the familiar common emitter configuration of Fig. 1, the first choice to be made is R3 — the collector load with which the transistor will develop the required signal output. The calculated value will affect all other calculations and values, so it must be carefully worked out from the first. A number of things must be considered.

If the circuit is to be battery powered, a high value for R3 will be preferred to save battery power. If it is to be mains powered, lower values will mean improved performance.

In theory, maximum power transfer requires our circuit source and load impedances to match the preceding tape-recorder load and succeeding power amp source impedances. This means the circuit should have a 1k0 input impedance and a 50k output impedance.

Perhaps the dominant argument against the theory, particularly in this case, is that distortion enters the practical picture. Transistors are essentially nonlinear current-operated devices. To maintain minimum distortion the maximum amount of voltage negative feedback must be applied. Since we require an overall voltage gain of 10, we must ensure there are no significant signal voltage losses at input or output.

Fig. 1 Common emitter amplifier circuit discussed in text.
or the amplifier would need extra gain to compensate.

The two practical starting criteria are that, as a rule of thumb, the input impedance of the amplifier should be around 10 times greater than the signal output impedance (in this case, greater than 10k) and that the amplifier output impedance should be around 10 times lower than the load impedance (or less than 5k). Thus signal losses are kept low.

Now back to the original problem — the choice of R3. To be 10 times less than the power amplifier’s input impedance, it should have a value of 4.7k.

The next practical assumption to make is that the potential at the transistor’s emitter should be about 10% of the supply voltage — in this case, about 1.2V. This is to ensure DC stability over a wide range of operating conditions.

Many engineers, unaware of the advantages and disadvantages of different circuits, play safe and go straight for an IC Op-amp . . .

For the amplifier to provide maximum output voltage swing, the collector potential should lie halfway between the supply rail and the emitter potential. For a 12V supply it must be:

\[ V_{CC} = (V_{EB} + V_{CC})/2 = 6.6 \text{V} \]

From this we can calculate the nominal collector current (which, together with Vcc, helps us determine a suitable transistor):

\[ I_C = (V_{CC} - V_{CE})/R_e = 1.14\text{mA} \]

The emitter potential has already been set at 1.2V, so R4 can be found from

\[ R_4 = 1.2/(1.14 \times 10^{-3}) = 1k0 \text{ (approx).} \]

R1 and R2 are there to ensure that the transistor base is always supplied with enough current to maintain the required collector potential, and the third rule of thumb is that there should always be more than 10 times the required base current, \( I_B \), flowing down the bias chain R1-R2.

To determine the base current we have to look in the manufacturer’s data sheet for probably the most important of the h-parameters — hFE, or the transistor’s common emitter DC current gain. For a BC237B, hFE is between 250 to 500. The variation is due to manufacturing tolerances and we should take the worst case figure of 250 to ensure that our circuit will function correctly with any BC237B transistor. The rules of thumb used earlier make sure that a higher-than-minimum gain transistor will not significantly alter the bias condition we are about to set up.

Since base current equals collector current divided by current gain

\[ I_B = h_{FE} = (1.14 \times 10^{-3})/250 = 4.5 \mu A \text{ (approx).} \]

The current flowing down the bias chain, R1-R2, should be around 10 times \( I_B \), say, 50 \mu A.

Since the transistor is passing collector current, the base potential must be about 0.6V above the emitter potential, \( 1.2 + 0.6 = 1.8 \text{V} \). It is clear that the potential across R1 equals \( V_{CE} = 1.8 \text{V} \), or 10.2 V. Resistor R1 should pass around 50 \mu A and should therefore be about 10.2/(50 \times 10^{-6}) ohms. This comes to about 204k and, bearing in mind that it is better to round down than up, since more current is better than not enough, we set R1 at the preferred value, 180k.

To avoid a build up of errors due to rounding, it is best to recalculate \( I_B \) the current through R1. \( I_B \) equals 10.2/(180 \times 10^3), or about 56.6 \mu A.

Since the current through R2 equals the current through R1 minus the transistor’s base current, we can calculate R2 simply:

\[ R_2 = 1.8/((56.6-4.5)\times 10^{-6}) = 34.5 \times 10^3 \Omega \]

The nearest preferred value is 33K.

The DC conditions are set up so that the transistor is operating in its linear region and can provide amplification. The amount of amplification, or voltage gain, is determined by the total collector impedance divided by the total emitter impedance.

Collector impedance equals R3 plus the load plus the transistor’s own output impedance all in parallel.

In most cases, the transistor’s output impedance can be ignored as it is high compared with the collector load resistor. This makes collector impedance equal to 4K7/50K, or around 4K3. The transistor’s emitter impedance is derived using complex mathematics from the diode equation and can be approximated to 25R/1f (mA).

Note that we’ve assumed that collector current and emitter current are equal. This is acceptable for modern day transistors with hFE figures of 200 or more. Some transistors have an hFE of well over 1000, these days, and in these cases the transistor’s base-emitter current is so small, compared with collector-emitter current, that it can be safely ignored.

The emitter resistance can now be calculated by dividing 25 by the emitter current in mA, bearing in mind that emitter current equals collector current:

\[ R_e = 25/(1.14 = 22R \text{ (approx).} \]

This internal resistance is in series with the emitter load resistor R, so the total emitter impedance is 1022R. The voltage gain is simply

\[ A = 4300/1022 = 4.2 \text{ (approx).} \]

This is not the value of 10 we were aiming for. To get that, the total emitter resistance should equal the collector resistance divided by the required gain, or around 430R.

**Current Complications**

To alter R4 now would upset our carefully calculated bias conditions. So we introduce R5 and C2 to lower the effective emitter resistance to AC signals and therefore to increase AC gain. The DC bias conditions are left unaffected, thanks to C2, but to AC signals, R4 and R5 are in parallel.

R5 should be chosen so that \( R_5 + (R_4/R_5) = 430R \). Thus:

\[ (1/10000)+(1/R_5) = 1/(430 - 22) \]

\[ R_5 = (408 \times 10^3) = 680 \text{ (approx).} \]

R5 determines the overall AC gain and can, of course, be altered for different requirements or even replaced with a small preset potentiometer. C2 should be large enough to pass the lowest audio frequency expected without too much attenuation. Very roughly, the AC gain will drop by 3dB when the
reactance of $C_2$ plus $R_5$ equals $R_4$, assuming $R_e$ is low. If the roll-off frequency is to be 20Hz, then,

$$C_2 = \frac{1}{(2\pi \times 320 \times 20)} = 24.8\mu F$$

$$= 22\mu F$$

is the nearest preferred value.

A quick look at the input impedance will complete the calculations around this single stage amplifier. Then we will be able to look at its limitations and some ways to overcome them.

The transistor's input impedance can be calculated from $h_{ie}$ and the total emitter resistance:

$$R_{in} = 250 \times 430 \times 107 k\Omega$$

But the input signal also has to flow through the DC bias resistor, $R_2$, which can be considered to be in parallel with the input, reducing input impedance.

Total input impedance is approximately $107k\Omega/33k$ or about $22k$. Happily, this still falls within the original specification that input impedance should be greater than 10k.

**Maximum Gain**

An interesting problem faced by many enthusiasts is to estimate the stage gain when the emitter resistor, $R_4$, is totally decoupled — that is, when $R_5$ is a short circuit. Some people guess quite high figures. Others like to impress and bring out a string of complex formulae straight from a text book. The formulae are enough to put most people of electronics for life, while the guesses are usually wrong.

There is a very simple rule of thumb which works very well. The gain of a common emitter stage without emitter degeneration is simply 20 times the supply voltage.

**The formulae are enough to put most people of electronics for life, while the guesses are usually wrong . . .**

Since this often surprises people, here are the sums:

Assuming high frequency operation, so that $C_2$ is very low, and making all the other simplifying assumptions we've used so far,

$$A = \frac{R_3/(R_5/R_4)+R_e}{R_e}.$$  

Since $R_5=0$, this reduces to

$$A = \frac{R_3}{R_e}.$$  

Now, $R_e = 25 \times 10^{-3}/I_e$ and $I_e = I_C$.

In turn, $I_C = (V_{cc}-V_{ce})/R_3$, so:

$$R_e = \frac{(25\times10^{-3}/R_3)/(V_{cc}-V_{ce})}{X40}.$$  

$$A = \frac{(V_{cc}-V_{ce})}{(25\times10^{-3})}$$

$$A = (V_{cc}-V_{ce}) \times 1000/25 \text{ or } (V_{cc}-V_{ce})x40.$$  

If we now assume that component values are chosen so that $V_{ce}$ lies more or less midway between $V_{cc}$ and 0V, then $V_{cc}-V_{ce}$ will be $0.5V_{cc}$. Thus the gain equation reduces to $A = 20V_{cc}$.

The last assumption we made follows from setting the transistor's operating point. The '20V_{cc}' rule of thumb then gives the maximum gain under ideal conditions and without taking input and output loading into account. In practice, maximum gain is usually between $1/2$ and $1/4$ of this figure.

**In order to design a single stage amplifier, its purpose must first be established . . .**

Next month, we'll discuss refinements to the common emitter circuit and other configurations. There will be considerably less mathematics, since the principles we've dealt with above should be enough to stand you in good stead should you want to make your own practical calculations.
WOMEN AND IT

Information technology is the industry of the future. But where are the women in the electronics and computer industries, asks Anna Paczuska?

Discussion of the chronic skills shortage which threatens the British IT industry centre on whether the government or industry should finance the training programmes needed to make good the shortfall. Both parties, however, effectively ignore one area of recruitment — women.

Girls and women represent a sizeable resource of untapped skill at technician and technologist levels as well as at professional level. This is particularly important in view of the population trends which predict a significant downturn in the overall number of school leavers over the next ten years.

Token statements about the need to encourage women appear in all the reports. The Department of Trade and Industry (DTI), for instance, states in one report about skills shortages at technician level:

'Everyone involved with the provision or application of IT skills should encourage initiatives at school level and onwards to increase the number of people studying relevant disciplines and in particular to interest girls in IT-related subjects.'

But 'everyone' clearly does not include the DTI. There is no further mention of women in that report beyond the predictable generality that 'initiatives are under way to attract more women into the industry.'

Invisible Signs

But where? The DTI certainly isn’t doing it. It is constrained by the government’s lack of funding for training schemes and consequently lays stress on the need for training initiatives at industrial level, pointing to such solutions as in-house and in-company training schemes, private provision as offered by the electricians’ union (EEPTU) and courses only part funded by the government, such as the Microelectronics Applications Programme (MAP). The DTI is also keen on training courses provided for third parties by, for example, BT at Stone, Cable and Wireless at Portcurno and systems houses such as STC/IAL at Bath.

In its latest report the DTI argues optimistically that:

'there are signs that a more positive message is being transmitted by IT companies both to girls at school and to women already in employment.'

A gross inflation of the facts.

Not only are there pitifully few training schemes at all, but girls hardly get a sniff at those. Engineering Industry Training Board (EITB) figures for 1984 show that only 5.3% of scientists and technologists in the electronics industry were female, and only 9.9% of the trainees. Similarly, 4.4% of the technicians, including draughtsmen, (sic) were women, and they were only 8% of the trainees.

The problem is that not only do employers do very little to encourage and recruit girls, but circumstances in the schools positively discourage them.

Far fewer girls than boys present themselves for training. Figures show that Computer Studies is, at present, a minority subject in all schools and that everywhere it is studied very largely by boys. The attitudes are formed early on and harden as the pupils get older. A study by the Equal Opportunities Commission (EOC) in Croydon concluded that the ratio was less than 2½ to 1.

Girls’ learned helplessness with machinery and technology is initiated at an early level and reinforced from all sides. According to a representative of Acorn Computers we spoke to, a recent ‘Audit of Great Britain’ showed that in all households owning micro-

Women only earn 83% of the average male rate for the same work . . .

...then place the finger on the button and depress...

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computers, boys are 13 times more likely than girls to be using them. Only 4% of home micros are used by mothers. The idea that microcomputing is a boys' game has prompted Angela Bowie, professor of Business Administration at the University of Strathclyde to describe home computers as 'the Meccano sets of the 80s'. Bowie says that by putting up with this state of affairs, women are seriously handicapping their daughters. She argues that more women should buy and use home computers.

Girls And Boys Together?

Lack of access to home computers is not the only barrier between girls and new technology. Research shows that even in school, where computers are available, girls get a raw deal. A recent study funded by the EOC and conducted by Lorraine Culley in Midlands secondary schools revealed that whenever boys and girls worked together in mixed groups, the boys hogged the machines and the girls rarely got a turn. Small groups of boys in a mixed class argued so much about turns and were so noisy that they effectively demanded all the teacher's time. The girls in their groups tended to work more co-operatively but got less teacher time (because of the boys) and so were often stuck for long periods and achieved less. Small wonder girls opt out of computer studies.

Against this pessimistic picture, the Acom survey found that girls in 'girls-only' schools did much better at computer studies than girls in mixed schools. It seems that girls are better off without the boys around. The EOC has suggested that computer clubs for girls only may help to encourage them. It has also made a long list of recommendations about the extended use of computers across the curriculum, the training of teachers, and the preparation of non-sexist teaching materials in order to give all pupils equal chances to acquire computer skills.

Where such measures have been implemented, says the EOC, girls do much better. But they cost money and with the present government policy of education cuts it is doubtful whether such schemes will be widely implemented.

All Work And No Pay

Sadly, girls are still worse off than boys even if, against all the odds, they eventually make it into a skilled job in the IT Industry. A Computer Staff Salary Survey recently produced by Computer Economics revealed that not only do women lose out in terms of access to the most skilled jobs (only 1.6% of data processing managers are women) but they earn only 83% of the average male rate for the same work.

Recent studies in the US have shown similar results. A group of researchers at Stanford University who monitored women working in high technology companies over a ten-year period found that even women with qualifications as good as men earned only 72% of the average male rate.

Women form a majority only at one level of the electronics industry: they make up 78% of all clerks, typists and secretaries . . .

In Britain, the disparity between male and female salaries may partly be explained by the fact that highly skilled women tend to find top jobs in smaller companies, which pay less. But it also indicates that companies think female skills are somehow worth less than men's. As the Computer Skills report says: 'Generally, the largest proportion of women was to be found at the bottom of each category of job, and in the less prestigious of the jobs.' That is borne out by the EITB figures which show that women form a majority only at one level in the electronics industry; they make up 78% of all clerks, typists and secretaries. They also provide a third of all manual workers.

A Woman's Place

Despite the massive growth of the computer industry, women's work remains limited to traditional areas. If girls receive any training at all at school, it is for jobs as typists and secretaries. Ironically keyboard skills are essential for work with computers, but as all sectors of industry become increasingly computerised the demand for workers with keyboard skills alone gets smaller and smaller. At the same time the women workers who make up the bulk of assemblers and semi-skilled operators in the electronics industry lose their jobs.

It's a problem that has forced even the frequently complacent EEPTU into action on the issue of retraining their women members. Last year, EEPTU members at Thorn-Ericsson in Darlington made a demand for retraining as part of their annual pay claim. After a long battle, the management agreed to sponsor retraining for women workers. They go in pairs for week long courses to the union college at Cudham, in Kent, where they get an introduction to analogue electronics. The scheme has only just started but research officer, Liz Allen, says there is a lot of enthusiasm for it among the women.

'Times have changed. Women want to hang onto their jobs. Though many of our
women members have long service in companies, what they know isn't valued.'
Liz also points out why training is such a difficult demand to negotiate.

'Many managements still don't believe that semi-skilled workers can be skilled.'
Management prejudice toward training girls and women as technicians is, unfortunately, all too common. The Women's Technology Centre in Leeds, which provides £6000 a year with training in electronics and computing, found work placement difficult in its first year.

'Some firms wanted to know why girls wanted to become trainees at all,' says Christine Donaldson, one of the Centre's organisers. Now, after three years many firms have become more enlightened, probably because the courses have been so successful.

The Centre takes on 30 under-25s and 30 over-25s every year, recruiting with a positive discrimination policy which gives priority to women from ethnic minorities, the disabled and women with few or no formal qualifications. The Centre also has a child-care co-ordinator who helps place children in day-nurseries or with registered childminders.

The results are excellent. Of the second year intake, 63% got jobs in related areas and a further 18% went on to higher training, including three to degree level courses. Much of the credit is undoubtedly due to the staff who have to re-apply for funding from the European Social Fund and Leeds Council each year, but stick with the course despite the insecurity.

Money, Money, Money

Which is remarkable. Money is the main problem and money is what many schemes, even the good ones, don't get. Funds have just been withdrawn from the Girl Technician Training Scheme funded by the Manpower Services Commission and run by the EITB. The scheme grew out of the Scholarship Scheme for Young Women set up in 1975. This was designed to publicize the idea that the electronics industry can and should recruit women. In 1979 it was replaced by the Girl Technician Training Scheme in which employers receive a £5000 grant (now £6000) for every girl technician they recruit above normal requirements.

Over 150 employers, mainly the large concerns who employ the most technicians, took on girl trainees during the Scheme's life. Training used to be 'off-the-job' in workshops or FE colleges during the first year, and 'on-the-job' in subsequent years. Of the 346 girls recruited in 1979-80, 61% completed their training. The drop-out rate among girls was higher than for boys in the first year because, as EITB researcher Sue Peacock says, 'Some girls found they just couldn't cope with being the only girl.' But as Sue points out, the scheme has been successful and has been responsible for the slight rise in the number of girl technicians in electronics over the past few years, from around 1% when the scheme started to over 4% of the total.

Despite the scheme's success, the MSC has withdrawn its funds because money for training 16- to 19-year-olds must now go to YTS, not to higher level schemes.

'It's a pity our scheme can't continue just because of a silly rule,' says Sue Peacock. 'The girls on our courses have been highly motivated because they're pioneers. Five percent of them went onto employer sponsored degree schemes, compared with 3% of the boys.'

Even at the highest skill levels in the computing industry, women still suffer from prejudice and discrimination. Traditional and inflexible attitudes have been responsible for driving some women out of the centre of the industry into home-based initiatives. Indeed the DTI seems to think this is a good idea. In its latest report it says:

'The very flexibility of IT, and the fact that IT systems make it possible to work from home as well as the office or factory, means that it is particularly well suited as a career for women at all levels.'

It would be even more useful if one of us could type!

Doubtless the DTI have taken a cue from firms which have set up on just such a basis.

The systems house, F International, is a case in point. Set up as far back as 1962 it started as a small group of freelance programmers working from home. They now handle consultancy, design, training and systems analysis, employing over 850 people in the UK alone. Some 96% of their workers are female, but that's not because they are biased against men. As their representative told us, it's because 'men think they have to go out to work.'

Staff at F International work on a project contract basis, and work literally pours in — from multinationals and large British concerns alike. Barclays, Babcock Construction, British Steel, Lloyds, Shell, the Department of the Environment and British Aerospace are just some of their customers. Like everyone else, they're currently advertising for staff. They take only people with a minimum of 4 years experience in data processing and believe that they make available a pool of skills which traditionally organised firms cannot do.

'Why should career people with skills be lost to the computer industry just because they have family commitments?' their representative asked me. Why indeed? But do women have to go back into their homes and turn information technology into a glorified cottage industry before manufacturers will recognise, and pay for, their skills?

ETI NOVEMBER 1985

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OSCILLOSCOPE MEMORY DISPLAY

Does your oscilloscope persistently refuse to display slowly-changing waveforms? Ian Marshall explains how a little updating can help.

Anyone who uses an oscilloscope will know how frustrating it is trying to view slowly-varying signals on a standard short-persistence screen. Waveforms such as capacitor charge/discharge curves, modulation envelopes in electronic music generators, transducer outputs and physiological signals just cannot be satisfactorily displayed and studied on a general-purpose oscilloscope.

This add-on memory display allows flicker-free viewing of such signals by storing ten seconds of signal and constantly refreshing the screen fast enough to eliminate visual flicker. The displayed waveform scrolls across the screen in real-time or can be frozen for closer examination.

As a spin-off from the explosive growth of microprocessor electronics, a wide range of complex support chips are now available at virtually throw-away prices. This opens up all sorts of possibilities for projects which would have been inconceivable even ten years ago.

A microprocessor need not be used to take advantage of Random Access Memory (RAM) and analogue/digital interface IC’s. The memory display uses simple gates, flip-flops and counters to perform a function which does not warrant the complexity of microprocessor control. Apart from an inexpensive 2K RAM, an ADC and a DAC, only a few popular CMOS logic IC’s are required to build the display and all components should be readily available from a number of mail-order suppliers. The total cost should not exceed £35.

The Circuit

The input signal is first filtered and then sampled 200 times per second by the ADC. The resulting 8-bit data is stored in the 2K of RAM. In between the storing of samples, the entire memory contents are read out to the DAC and are displayed on the scope screen, with the timebase triggering from the scroll/freeze logic. The refresh rate of 200 per second (5ms per frame) is high enough to create the illusion of a continuous, flicker-free display.

In the normal scroll mode the control logic ensures that successive
samples are stored in consecutive memory locations, so the memory contents are gradually updated at the rate of one location per display sweep of 5ms. The displayed waveform thus appears to scroll across the screen, taking ten seconds for all 2048 addresses to be updated.

In the freeze mode, successive samples are stored repeatedly at the same memory address at the extreme left-hand side of the display. The stored data corresponding to the other 2047 addresses remains unchanged from sweep to sweep, so the display appears stationary.

The circuitry conveniently breaks down into the control logic (Fig. 2),

![Fig. 2 Circuit diagram of the control logic.](image)

The ubiquitous 555 timer, IC1, is used in the astable mode to produce a clock 0 of approximately 400kHz. The clock signal drives both the scroll/freeze counter IC2 and the memory address counter IC7.

In the normal scroll mode of operation, the Q12 output of counter IC2 is taken back to its reset input via flip-flop IC3a which introduces a delay of one clock cycle before the count sequence restarts. Thus 2049 clock cycles elapse before the count sequence repeats itself. The rising edge of output Q11 drives flip-flops IC4a and IC4b to generate a Start Convert (SC) pulse for the ADC. This pulse is synchronised with the clock, and is one clock cycle in duration.

Once the ADC conversion is initiated, its Busy line goes high until conversion is complete (in nine clock cycles) when it returns to the high state. The rising edge of Busy sets flip-flop IC5a, and flip-flop IC5b generates an Output Enable (OE) pulse to request data from the ADC. OE is gated with 0 to produce a Write Enable (WE) pulse for the RAM; the gating is necessary to ensure that WE is not still low when the RAM address lines are changed as counter IC7 advances on the next clock edge. Gates IC6a and IC6b are used to provide a delay from 0 to the WE gating, matching the propagation delay through flip-flop IC5b. By means of the OE and WE pulses, the new sample of data is stored at the address specified by IC7. Samples are taken every 2049 clock periods, giving a sampling rate of approximately 200 per second. Because 2049 clock cycles elapse in between successive sample/store operations, the address counter IC7 will have advanced 2049 states - one complete count sequence of 211 plus one extra state. Thus each sample is stored progressively one memory location later, overwriting the old data. With the oscilloscope sweep triggered from

The scroll/freeze logic (IC2 and IC3a) dictates the relationship between sampling interval and display memory address and, with ICs 4, 5 and 6, produces the pulses necessary to take a sample, store it at the correct address, and trigger the oscilloscope sweep. In the scroll mode, IC3a introduces a one-cycle delay so that 2049 clock cycles elapse in between samples. During this time the memory address counter IC7 has progressed through all its states and advanced one further count. Samples are thus stored progressively one memory location later each time, creating the scrolling action. In the freeze mode, the sampling pulses are synchronised with the memory access so that only the one location is updated repeatedly while the rest of the data remains frozen.

**Construction**

With the exception of the mains transformer, all the circuitry is contained on a single printed circuit board of standard Eurocard dimensions (160 x 100mm). For ease of etching and assembly, the board is single-sided and requires several wire links to be inserted as shown on the overlay diagram. The short links can be of bare tinned copper wire, but the longer runs carrying the ADC control signals should be of insulated wire.

Once the links are in place, proceed with the IC sockets, resistors, capacitors and terminal pins. The copper tracks are quite fine and closely spaced around counter IC7.

**HOW IT WORKS — CONTROL LOGIC**

OE the display appears to scroll, the input signal appearing at the right-hand side, moving to the left across the screen, and disappearing. At the sampling rate of 200, all 2048 addresses are updated in just over 10 seconds which is therefore the amount of signal displayed.

In the freeze mode, the reset line of IC2 is held low by R3 and the counter is allowed to cycle freely in natural binary fashion. Output Q11 has positive transitions every 2048 clock cycles and the resulting SC pulses, ADC conversion and sample storage are thus synchronised with the memory address counter. Successive samples are stored at the same location in memory whilst the contents of the other 2047 RAM addresses remain unchanged. The display therefore appears stationary with the latest 10 seconds of signal frozen.
In the memory channel, the signal to be displayed is applied to potentiometer RV1 and a portion of it is taken from the wiper to the second-order low-pass filter (LPF) formed by IC9. With a frequency cut-off of approximately 50 Hz, this filter prevents aliasing of the signal by removing frequency components above half the sampling rate. The filter has a passband gain of 5, (set by resistors R7 and R8), so that in conjunction with the gain control RV1, a wide range of signal levels can be accommodated.

The ZN1427 analogue-to-digital converter (ADC) IC10 is an 8-bit successive approximation type, taking nine clock cycles to complete a conversion. It has an on-chip reference voltage of 2.56 V which is available at pin 8, and in this design it is used as the reference input by connecting it to pin 7. A conversion is initiated by an SC pulse, whereupon the Busy line goes low. At the end of the conversion period, Busy returns high, signalling the control logic that a sample has been taken and is ready to be stored in memory. The subsequent OE pulse from IC8 enables the ADC outputs (which are normally in the high impedance state) placing the new data on the data bus. Simultaneously, WE is taken low so that the new data is written into the 2k58 RAM IC8, the address being dictated by address counter IC7. This sampling and storage operation occurs in between successive display sweeps.

During a display sweep, the RAM is in read mode and counter IC7 clocks round the address lines. The data from the RAM is read out in sequence to the digital-to-analogue converter (DAC) IC11, and the reconstructed analogue signal is available at pin 4. The output from the DAC is passed through low-pass filter IC12, which cleans up the discrete step-like nature of the waveform for subsequent display. The output signal lies in the range zero to +2.55 V, so a zero input voltage will emerge from the memory channel with an offset of 1.27 V. This is of little consequence as the output signal is displayed on an oscilloscope which will have a Y-shift control.

Mains transformer T1 together with bridge rectifier BR1 and reservoir capacitor C10 provide an unregulated supply of approximately 8 V DC. This is regulated to 5 V by IC13 to supply the logic circuitry and the positive rail of op-amps IC9 and 12. Capacitors C11 and C12 are situated close to the regulator to improve its stability and noise performance.

The op-amps and the ADC also require a negative supply, at a total current of typically 3 mA. Rather than use a centre-tapped transformer and another reservoir capacitor/regulator combination, this low current is provided by voltage converter IC14. This is configured to generate a mirror-image of its supply voltage, -5 V from +5 V. The total current drain of the memory display at 5 V is approximately 90 mA.

Fig. 3 Circuit diagram of the memory channel.

**HOW IT WORKS — PSU**

![Circuit diagram of the power supply](image)

**NOTE:**
- IC13 = 78L05
- IC14 = TLC7660
- BR1 = W01 OR SIL
- BR1 = W01 OR SIMILAR
- 100 V, 1.5 A BRIDGE RECTIFIER

Fig. 4 Circuit diagram of the power supply.

ETI NOVEMBER 1985
and the RAM IC8, so care is necessary to avoid forming solder bridges between adjacent tracks and pads around this part of the layout. Solder in the regulator IC13, but do not insert any of the other ICs at this stage.

The case can be of plastic or metal (or even French-polished walnut for that matter!), the only restrictions being that it should be large enough to house the PCB and mains transformer and that it should have a front panel, preferably of metal, on which to mount the various sockets and controls. The mains cable enters from the rear, being securely anchored in the back panel by a strain relief grommet. The PCB is fastened to the base of the case by means of short spacers. The transformer frame and all accessible metal parts of the case must be earthed, which may mean taking earth wires to the individual panels. Assembly screws do not necessarily provide an adequate low-resistance contact.

The choice of transformer is not critical, and practically any mains type capable of providing between 6 and 9V RMS at 200mA will suffice. The raw DC voltage at the input of IC13 should not fall below 6.5V or the regulating action will be lost. If this happens, the output will track the input voltage but at about 1.5V lower. On the other hand, the mean input-output differential at the load current of 90mA should not be

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**ETI NOVEMBER 1985**

**PARTS LIST**

<table>
<thead>
<tr>
<th>RESISTORS (all 1/4W,5%)</th>
<th>C5</th>
<th>C6,12</th>
<th>C7,11</th>
<th>R1</th>
<th>R2</th>
<th>R3</th>
<th>R4,R5</th>
<th>R7</th>
<th>R8,14,15</th>
<th>R9,11</th>
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| BR1                    | W01 or similar bridge rectifier |     |     |     |     |     |

<table>
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<tr>
<th>MISCELLANEOUS</th>
<th>SK1-3</th>
<th>SW1</th>
<th>T1</th>
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<tbody>
<tr>
<td>SK1-3</td>
<td>BNC chassis mounting sockets</td>
<td>Single pole toggle, slide or pushbutton switch with NC contacts</td>
<td>Mains transformer 6V, 200mA (1.2VA) secondary</td>
</tr>
</tbody>
</table>

| PCB; mounting pillars; case strain relief grommet for mains lead; knob; IC sockets, 4 off 8-pin, 5 off 14-pin, 2 off 16-pin, 1 off 18-pin and 1 off 24-pin DIL; nuts, bolts, etc. |     |     |     |

---

**Fig. 5 Component overlay for the memory display PCB.**
greater than 6V or the dissipation in the regulator will exceed the safe limit (0.6W for the TO-92 package). If you already have, say, a 12V transformer in the junk box, the excessive power can be burned off by using a TO-220 packaged 7605 regulator for IC13. This device has a maximum dissipation of 2W in free air, rising to about 8W with a 10°C/W heatsink fitted.

Mount the BNC sockets, the switch and the potentiometer on the front panel. A toggle switch is specified for the Freeze control, but if momentary action is preferred a biased toggle switch or a pushbutton could be used instead. The contacts should be normally closed. The wiring to the controls is straightforward and should present no problems provided the diagrams and the photograph are followed carefully. Construction is completed by wiring the mains lead to the transformer primary and connecting the secondary winding to the PCB.

Testing

Apply power to the Memory Display and check the DC voltages present at the input and output of the regulator, IC13. The output should be within 0.3V of 5V. If all is well, insert the negative rail generator IC14 and check that -5V is present on its output, pin 5.

Having established that the power supply is working correctly, testing of the logic circuitry can begin. It is assumed that the constructor has access to an oscilloscope, and waveform diagrams are given in Fig. 6. Proceeding systematically, using these waveforms and the circuit description given in How It Works, it is possible to insert the IC’s in easy stages and check circuit operation. In this way any problems due to solder bridges, bent IC pins and so on will be quickly brought to light.

Begin with the master clock (φ) generator IC1. A square wave at a frequency of approximately 400kHz should be seen at its output, pin 3, and should appear on the appropriate pins of other IC sockets across the board. Next insert IC2 and IC3 and monitor the Q11 output (pin 15) of IC2, which should be a squarewave of period 5ms. Now insert IC4 and check that each rising edge of Q11 produces a Start Convert (SC) pulse. At this stage the ADC (IC10) can be inserted, ensuring that SC and the clock φ reach it, and that its Busy line goes low for nine clock cycles after the SC pulse. With ICs 5 and 6 in place, the rising edge of Busy should be seen to produce the OE and WE pulses.

With the control logic tested and working properly, proceed to the memory channel. Insert the address counter IC7 and check that its outputs are clocked round by φ. If all is well, the RAM IC8 can be carefully put in its socket, and the DAC IC11 and the two op-amps IC9 and IC12 inserted. It should now be possible to apply a signal or a voltage level to the input socket and see the scroll/freeze action at the output socket, with the ‘scroll’ display sweep triggered by the OE pulse which is brought out to SK3, the Trigger output.

In Use

The waveform to be displayed is applied to the input socket and the oscilloscope trigger input is connected to the trigger socket. The scrolling display is taken from the output socket to the scope Y-input. Set the scope timebase to produce a sweep of approximately 5ms so that the memory contents are read out once per sweep. Adjust the gain control until the displayed waveform is as large as possible without exceeding the range of the ADC. This will be apparent as clipping of the display in much the same way that analogue circuits clip if over-driven.

The frequency response is from DC up to a theoretical maximum of 50Hz (set by the low pass filter, IC9), but if sine waves are being viewed the practical upper limit is set by the number of cycles that can be resolved across the screen. This is a function of the trace width, screen size and viewing distance.

Potentiometer RV1 sets the input impedance at 10k, so some form of buffer amplifier will be required if you wish to look at high-impedance sources such as capacitor charge/discharge curves.

Refinements

Experimentally-minded readers have plenty of scope (sorry!) for modifying the memory display from the form presented here. The most obvious possibilities include expanding the memory and/or having more than one memory channel. The primary design constraint is that however much waveform is stored, it must all be read out to the scope screen at least 50 times per second to avoid annoying visible flicker. Happy viewing!
**THE RHYTHM CHIP**

Geoff Phillips calls it the RHYTH.ROM because it uses EPROM data to generate and measure beats-per-minute. Hit me, hit me!

The simple method of designing a beats per minute meter is to use the standard analogue frequency-to-voltage approach. Each beat is made to trigger a monostable of set pulse width and the pulses are then averaged in an integrator. The output voltage of the integrator is proportional to the beat rate and an analogue meter can be scaled accordingly. A design employing this technique would have a slow response time and always has an annoying ripple voltage present which causes the meter needle to quiver at the beat rate being measured. As always, a digital method is best. A binary counter can be used to count up pulses from an oscillator during the time period between two beats of the music. The number of pulses in the counter is a measure of the beat rate.

The problem is converting this count into a binary coded decimal number equivalent to the beats per minute. Real time calculation of the beat rate digitally would involve the use of microprocessors. If all the answers are calculated by hand, however, they may be loaded into an EPROM which would then act as a look-up table giving a correlation between the binary number in the counter and the appropriate BCD numbers for an LED display.

**Beat The Clock**

As most EPROMs have 8-bit outputs, the beats-per-minute information will best be stored as two-digit BCD numbers. This will look just the same as two-digit hex numbers to the EPROM but the display will be limited to a maximum beat rate of 99. It was decided to incorporate a 2x switch facility to cater for beat rates up to 198. The clock input of the binary counter is switchable between the clock oscillator and a divide-by-2 flip-flop. If the EPROM is programmed for the divided clock then switching to the normal clock will enable beat rates of 99 to 198 to be measured. The display resolution is obviously 1 beat per minute. Ninety-nine beats-per-minute is 0.606 seconds-per-beat and 98 beats-per-minute is 0.612 seconds-per-beat, so a measuring resolution of about 6ms would be adequate. A 100Hz clock was chosen to make the maths easier. The lowest beat rate that needs to be catered for is in the region of 25 which means that only the first 256 addresses of the EPROM need be programmed.

Ninety-nine beats-per-minute is equivalent to 0.606 seconds-per-beat. During this time, 0.606/.01 (=60) clock pulses would be counted in the binary counter. The first 60 (decimal) addresses of the EPROM should therefore be loaded with hex A1 which will give an overrange indication on the LED display of blank-1. The data for the remaining EPROM addresses may be calculated as follows:

\[
\text{EPROM data} = \frac{60}{\text{EPROM address} \times 0.001}
\]

rounded to the nearest whole number. The calculation is performed in decimal but the data entered in 2-digit hex format.

For those who can't be bothered to work them all out, Table 1 shows all the answers. For those of you who are even lazier, a preprogrammed EPROM is available (see Buylines).

**Making ROM**

A 555 timer is used as the metronome. It is connected in the astable mode and can drive a 50 ohm speaker direct. It gives a series of clicks not unlike a conventional metronome. The speed is adjusted with a potentiometer.

**Inside the prototype Rhyth·ROM — wire integral PSU.**
Fit the resistors and capacitors first. This is always a good idea when CMOS ICs are being used. When fitting the push button ensure that the two notches are in line with the length of the PCB. The 'x1-x2' switch is soldered straight onto the PCB. The leads of the tempo pot must be bent 90° towards its shaft and the pot is then fed from the rear of the PCB through the large hole and then secured with the pot nut on the top of the PCB. The switch section of the tempo pot is hard wired to the PCB with flying leads. Ensure that the LED displays are fitted with the decimal points towards the centre of the PCB. When the PCB is complete connect it to a 5V power supply and the Rhyth-UPM is now ready for calibration. (The prototype included a 5V supply in the case, but this is not necessary).

If you have access to a frequency counter, monitor the speaker output of the Rhythm-UPM. Set the 'x1-x2' switch to x1 and turn the tempo pot slowly until 1 beat-per-second is obtained and the frequency counter reads a period of 1,000 seconds. Adjust RV1 until the LED display of the Rhyth-UPM reads 60 beats-per-minute. If you don’t have a frequency counter, use a watch or clock to adjust the tempo rate.

**Operation**

The Rhyth-UPM is in the metronome mode of operation when the tempo pot is switched from its fully anti-clockwise position. Clockwise rotation of the pot increases the tempo. The LED display shows beats-per-minute directly when x1 is selected. If overrange is shown on the LED display then select x2 and multiply the display by 2. If the musician coincides the first beat of the bar of the music with the clicks of the Rhyth-UPM then the LED display will obviously show bars-per-minute. The speed of the music in ballroom dancing is very important. A modern waltz should be 32 bars-per-minute, a quickstep 48–50.

The thermometer mode is selectable by turning the tempo pot fully anticlockwise so that its switch is operated. The Rhyth-UPM’s speaker will then click when the rhythm button is pressed. Tap along with each beat of the music with the rhythm button and the LED display will show the tempo being played.

In the resistor manufacturing industry, resistor blanks are cut to length by a machine. Metal cups are then fitted by another machine and then a helix is cut by a laser spiral machine. All these machines are of a cyclic nature and production managers are very anxious to know their production rates in pieces-per-minute. The Rhyth-UPM is simply tapped in time with the machine and after a minimum of two taps the production rate is shown.

After a workout in the gym, feel your pulse and tap the Rhyth-UPM to measure your heart rate. The speed at which your heart recovers is an indication of how fit you are. It will probably indicate that you should be spending less time constructing electronic circuits and more time jogging.

---

![Diagram](image-url)

*Fig. 1 The circuit diagram of the Rhyth-UPM.*
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Table 1. EPROM contents, only first 256 bytes are used.

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Fig. 2 Component overlay.

HOW IT WORKS

IC1 is a 555 timer connected in the astable mode. When the charge on C4 has increased so that the voltage at pin 6 increases to 2/3 of VCC, then pin 3 goes low producing a click in the loudspeakers. Pin 7 goes low also which quickly discharges C4 via R7. When the tempo pot is switched to manual or temppometer mode, it is the rhythm button, SW1 which triggers IC1.

The pulses from IC1 are formed into a 1:1 mark space ratio by flip flop IC3a. When the positive going edge from IC3a, the binary counter IC4 is quickly reset to zero via C2, R3 and the contents of the EPROM IC5, are latched into IC6 and IC7. The clock oscillator formed by IC2a and b is then enabled via pin 6 and depending upon the position of the x1-x2 switch, pulses from pin 4 of IC2b or IC3b are fed to the binary counter IC4. When pin 1 of IC3a goes to a 0, counting ceases and the stored count in IC4 selects the appropriate address of IC5. The data from IC5 is allowed into IC6 and IC7 by virtue of pin 5 being at 0. IC6 6 and 7 decode the data and drive the appropriate segments of the LED displays. When pin 1 of IC3a goes to a 1 again the whole process is repeated.

PARTS LIST

RESISTORS (all 1/4W 5%)
R1 220k
R3,6 10k
R4,5 330k
R7 470R
R821 150R
RV1 22k horizontal miniature pre-set
RV2 100k lown pot with switch

CAPACITORS
C1 47n
C2 100n
C4,6,7 16V electrolytic
C5,6,7 100n
C8 1000p

SEMICONDUCTORS
IC1 555
IC2 4011
IC3 4013
IC4 4040
IC5 2516 (or 2716) single rail EPROM
IC6,7 4511
D1 1N4148
LED Displays (two off) HDSP5303 or equivalent
Red LED

MISCELLANEOUS
Printed circuit board
50 ohm loudspeaker
PCB mounted switch, PCB mounted button, case, knob.

BUYLINES

The loudspeaker, button, switch, and LED displays are available from Electrovalue. The pre-programmed EPROM and PCB are available from C P Electronic Services, 87 Willowtree Ave., Durham DH1 1DZ, at £5.20 for the PCB and £5.99 for the EPROM.

ETI NOVEMBER 1985
MODULAR TEST EQUIPMENT

Mike Meakin describes the second of his low-cost, slot-together test gear modules — a 1Hz to 100kHz sine, square and triangular wave generator.

This article describes the construction of a waveform generator which assembles onto a single printed circuit board complete with all the necessary controls and connectors. It is one of a series of test equipment modules which we plan to present over the next few months which together form a versatile integrated test system. Each module is entirely self-contained on one PCB, removing the need for cases and all the other hardware usually required and allowing a high performance test system to be assembled at minimum cost.

The heart of the system is a power supply module which was described in last month's ETI. As well as providing variable voltages to power equipment under test, it also provides the correct voltages to drive each of the other modules and so avoids the costs associated with providing separate PSUs. Ideally the power supply module should be built first, but there is no reason why the waveform generator should not be fed from any other suitable supply if constructors don't require all the facilities of the main PSU or want to assemble the modules in some other order.

The waveform generator produces sine, triangle and square wave outputs variable in frequency from 1Hz to 100kHz and with an output level of up to 5V RMS. The board continues the basic design philosophy of the system by having switches and potentiometers which mount directly onto the board. The switches are designed for preset programming but should still have a life of about 20,000 operations. They can be operated by a small screwdriver. The potentiometers are of a type normally used as presets and are small, PCB mounting and have an integral knob.

While superior performance can be obtained from waveform generators using op-amps and discrete components, it was decided for reasons of simplicity and cost to use an IC. The 8038 was chosen because it is readily available and offers reasonable performance for its cost. However it is not without minor problems. These arise when using the square wave output, which consists of an open collector NPN transistor with its emitter connected to the negative supply.
The 8038 is configured in the manufacturers application circuit and interested readers should refer to the Intersil data sheet. RV2 controls the frequency with RV1 setting the minimum frequency. The component values chosen allow a 1:10 frequency range to be covered in 5 switched overlapping ranges. The output waveforms are buffered and scaled by IC2a to give an output of 5V RMS±7V p-p for the sine wave output and similar peak to peak outputs for the square and triangle waveforms. A passive attenuator, R16-R17, and the output level control RV7 give three switched decade ranges and continuous level control.

The output amplifier consists of a unity gain inverting amplifier which sums the output from RV7 with a switched variable DC offset of ±5V controlled by RV6 and SW3. A47R resistor in series with the output provides additional protection for the output amplifier under short circuit conditions.

The power supplies are decoupled both by 47uF electrolytic capacitors and polyester capacitors. The module consumes about 30mA from each 15V supply rail.

HOW IT WORKS

The manufacturers recommend pulling the output up to the plus 15V supply via a 15k resistor. Both internal and external stray capacitances are charged via this pull up resistor, so the square wave output has slowly rising edges which are particularly noticeable at high frequencies (Fig. 1).

A more serious problem is that when the square wave output is used, spikes appear on the peaks of the sine wave output. These may possibly be due to internal earthing arrangements within the IC itself and the only sure cure is to disconnect the pull up resistor.

Both of these problems can be minimized by using the circuit shown in Fig. 2. The output transistor is now used as a current switch in a virtual earth op-amp circuit. When the transistor is off the output of the op-amp goes to approximately -7.5V due to the 100k resistor connected to the +15V supply. When the transistor conducts a current is drawn from the virtual earth and the output of the op-amp rises to +7.5V to supply this current. Rise and fall times of the square wave output are much improved and this configuration allows the use of a single pole switch, with

BUYLINES

Verospeed and Cricklewood are among those who can supply the enclosed horizontal presets and RS Components stock the carbon track presets with integral knob. RS will only accept orders from trade and professional customers but Crewe-Allen & Co of 51 Scarton Street, London EC2 will obtain parts from them on payment of a small handling charge. The stock numbers are 184-150 for the 10k presets and 184-372 for the 47k preset.

The DIL switches used on the prototype were on ERG D516D 1-6 or SW1 and an ERG D516D 1-3 + 1-3 for SW2. ERG will not handle small orders but electronics clubs and other groups prepared to order reasonable quantities could try contacting them directly at Luton Road, Dunstable, Bedfordshire LU5 4LJ, tel 0582-622441. The PCB has extra holes to allow eight position switches to be used instead (the extra positions can be ignored) and RS stock number 337-532 could thus be used as SW1. A standard eight-way DIL switch could be used as SW2 but you would have to make sure that only one switch in each group of four is ever on at one time. Another alternative is to use standard rotary or slide switches and glue or bolt them to the board. The PCB will be available from our PCB Service.

ETI NOVEMBER 1985
disconnection of the pull up resistor when not selected. There is a choice of output amplifier. A TL072CP biFET op-amp can be used for low cost or alternatively an NE5532 dual low-noise op-amp can be used for superior output performance. This IC can deliver over 10V RMS into a 600R load and was designed specifically for audio applications.

**Construction**

Because the waveform generator is designed for use without a case or front panel, the range markings on SW1 and the other legends will all have to be printed on the PCB itself. This is best done by screen printing the boards, but the cost is considerable and we have decided to supply the boards without screen printing in order to keep the price down. This allows constructors to choose their own method of labelling.

The cheapest method would be to mark the boards with a chinagraph pen or something similar. It might also be possible to use embossed tape (eg, Dymo) but the character size would probably be too large to allow much information to be fitted in the available space. Labelling by either of these methods is best done when the board has been assembled.

A better but more expensive method is to use rub-down lettering. This would have to be done before installing any of the components. Begin by cleaning the upper surface of the board thoroughly with a suitable solvent and then spray on a light coat of clear varnish — the type sold in motoring shops for lacquering chrome work is quite good. Wait until the varnish is thoroughly dry then loosely install the components which require labels and note how much space you have around them.

It is best to use white letting rather than black since it shows up well against the board. Most art shops stock sheets of rub down lettering in a variety of typefaces and sizes, but white is less frequently used than black and you may have to order some and wait. The most widely available lettering has a series of parallel broken guide lines on the sheet, and these can be lined up against the edge of the PCB as each letter is rubbed down to ensure that the row of lettering is straight. Check frequently as you work that none of the lettering will be obscured by the components when they are inserted, and when all is finished apply several coats of varnish at suitable intervals and leave to dry.

Installing the components should present no problems provided a reasonable degree of care is taken and the overlay diagram is followed closely. Begin by soldering into place the wire links, the socket, the switches and, if you are using them, the IC sockets, then move on to the resistors, the potentiometers and the capacitors. The hole positions may need to be adjusted with a small file if you are using potentiometers which are physically different from those used on the prototype. Finally, install the two ICs, making sure that they are the right way around.

If you are using the waveform generator with the power supply...
module described last month, solder leads from the appropriate pads on the generator to the thick copper tracks on the supply board. Note that only the ±15 V rails are needed and that the +5 V output from the supply is not used. If you have not built the power supply module, you will need a regulated ±15 V supply in order to operate the waveform generator. The current drawn will be about 30 mA from each rail.

Testing And Setting Up

Set the potentiometers and presets to their mid positions, the function switch to sine, the range switch to 100 Hz – 1 kHz, the level switch to 5 V and the offset switch to off. Apply power and monitor the output, either with a scope or by connecting it to an audio amplifier. For those without access to any other test gear, there is little more that can be done at this stage apart from checking ranges and operation of the offset control. However, in a future article the construction of a universal counter timer/frequency meter will be described which will allow the complete setting up to be carried out.

Those who have a ‘scope and frequency meter should select a square wave output at a frequency of about 100 Hz and adjust the mark space ratio for 50% duty cycle using the preset RV3. The frequency span should be adjusted for a 1:10 range using RV1 and the frequency potentiometer, RV2. There is considerable overlap of the ranges.

Switch to sine wave output and trim the distortion presets until you get the best wave shape. These interact with each other and can best be adjusted by looking at the waveform on a scope. A distortion meter could be used if you have access to one. Finally, check the output levels.

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BRITAIN'S FOREMOST QUALITY COMPONENT SUPPLIERS

ETI NOVEMBER 1985
SWITCHING REGULATOR

To kick-off a new, occasional series on circuits using discrete components only, Paul Chappell brings you a switching regulator that will provide a steady 5V at up to 1A.

I wonder if you can guess how many projects featured in ETI this year have not included at least one IC! Apart from a Tech Tip, which was almost apologetic about the fact, I can't find any at all. Now, I'm not advocating a return to the electronic stone age of cat's whisker and Leyden jars. Contrary to popular belief it really is possible to design circuits with discrete components that work well and don't use an excessive number of parts. The chances are you'll have most of the components in your spares box anyway, so the project won't cost much to build, and if anything burns it will be a 10p transistor and not a £10 IC. The best thing of all about using discrete components is that you can experiment with every part of the circuit, since it's not set solid in a block of silicon. Convinced? Then read on.

Switching Regulators

This month's circuit is a switching regulator. The basic circuit delivers 5V at 1A and has an efficiency of around 80%. With an input of 20V and a small heatsink on the switching FET, the circuit runs under full load at what washing powder packets call 'hand hot'. You can comfortably touch the heatsink without asbestos gloves, which is more than can be said for linear regulators under the same conditions. The output current capability can easily be increased with a suitable choice of FET and some attention to the choke.

The prototype switching regulator. (This is not quite identical to the final circuit.)

No doubt you're all aware of the ideas behind switching regulators so I'll be as brief as possible. The main problem with linear regulators is that they waste a lot of power. If we have an input of 20V and need an output of 5V at 1A, the regulator has to drop 15V and pass a current of 1A plus whatever its internal circuitry needs, and my calculator says that this comes to more than 15W. As the output to the load is only 5W the efficiency of the best linear regulator in the world under these conditions is going to be less than 25%.

A switching regulator, on the other hand, can have an efficiency of 100% in theory. When a switch is on it passes current but has no voltage across it. When it is off, it has a voltage across it but no current flows. In either of these conditions V x I = 0, so no power is lost. Naturally the 100% efficiency can never be achieved, partly because the regulator's internal circuitry requires some current, but mainly because of imperfections in the switch. The most serious imperfection is the time taken to switch from one state to the other, during which period it will be wasting power. Sensible efficiencies to aim for are between 70 and 85%.

One way a switch can be used to drop an input voltage is shown on the diagram. In the diagram, Vin is the input voltage, Vout is the output voltage, and LOAD is the load. The switch is controlled by the control circuit, which decides whether the switch is on or off. The switch is made of a FET, which is a type of transistor that can switch on or off very quickly. The switch is controlled by a signal from the control circuit, which is typically a digital signal. When the signal is high, the switch is on, and when the signal is low, the switch is off.

The switching regulator is a very efficient way of converting input voltage to output voltage, and it is used in many applications where high efficiency is required. The switching regulator is more complex than a linear regulator, but it is also more powerful and can handle larger loads.

Fig. 1 Basic switching regulator configuration.
ETI

Fig. 2 Inductor current.

Fig. 3 (left) Basic drive circuit and
Fig. 4 (right) improved bootstrap drive circuit.

in Fig. 1. When the switch is closed, current begins to flow in the inductor and Vout starts to rise. When the switch is open, the inductor will produce a voltage in the direction needed to maintain the flow of current, so the right hand side will become positive with respect to the left hand side. As the voltage at the right hand side can't change instantly, because of the capacitor, the left hand side will go negative until it is 'caught' by the diode at 0V or thereabouts. Current will continue to flow into the load, the circuit now being completed by the diode, but will reduce as the energy stored in the inductor leaks away. If the switch is continuously opened and closed, the current through the inductor will vary about a mean level, and as long as the switch is operated frequently enough it will never stop flowing altogether. The inductor effectively smooths out the pulses from the switch.

HOW IT WORKS

Fig. 5 Circuit of the switching regulator.

Q5, D3, L1 and C5, 6, and 7 correspond to the parts shown in Fig. 1 with Q5 taking the place of the switch. Q3 acts as a constant current source giving 1mA into the emitters of Q1 and Q2. The portion of the current that each transistor will take depends on the relative voltages of their bases. If Q1 base is at a much lower voltage than Q2 base, Q1 will take the current. If Q2 base is lower than Q1 base, Q2 gets the current. There is a small linear region when the base voltages are almost equal and the transistors share the current, but otherwise one transistor or the other gets the lot. A portion of the output voltage is tapped off by R1V and applied to the base of Q1 while Q2 base is held at a constant voltage by D1. (I used a band-gap reference diode there, by the way, but in this application a zener would be just as good.) R1V is set so that when the output is at exactly 5V, the voltage at the slider equals the voltage of the reference diode. If the voltage at the output rises, so will the voltage of Q1 base, so current will be steered into Q2, turning on Q4 which in turn takes the gate of Q5 low. This turns off Q5, the current in the inductor begins to decay, and when the inductor current becomes lower than the load current the output voltage will begin to drop. When the output voltage drops below 5V, Q1 base goes low with respect to Q2 so current is steered into Q1 and away from Q2. Q4 then turns off, allowing the bootstrap action of C4 and R5 to take the gate of Q5 high. Q5 turns on, the current in L1 builds up, the output voltage eventually rises slightly and the whole cycle takes place again. The entire circuit will therefore be in a state of continuous oscillation—a around 100kHz in the prototype.

The action of the bootstrap components is described elsewhere. C2 and R3 serve to pull Q1 and Q2 quickly through their linear region to improve the comparator action.

For a given load, the output voltage will depend on the mean current and therefore on the ratio of 'on' to 'off' times of the switch. With the addition of some circuitry to sense the output and keep it at a constant level by operating the switch, we have a switching regulator.

The Switch

I chose an n-channel MOSFET as the switching element more for reasons of economy than anything else. P-channel devices tend to cost around twice the price of their n-channel counterparts, so it makes sense to avoid them where possible. The circuit of Fig. 3 would work as a means of driving the FET, given a high enough input voltage to allow Q1 to be driven hard into conduction and a small enough value of R1 to charge the gate capacitance within a reasonable time. (The faster Q1 turns on, the lower will be the power it will dissipate, so the higher the efficiency of the regulator.)

The circuit of Fig. 4 gives an improvement in turn-on time. The idea is that when the drive transistor Q2 is conducting, the FET will be turned off, its source
voltage will be at 0V or thereabouts, and C1 will charge up to the full input voltage. When Q2 turns off, Q1 conducts and its source voltage rises to the input voltage. The voltage across C1 remains, and as the voltage at Q1 source rises, so does the voltage at the other end of C1. This reverse-biases the diode and lifts the voltage at the top of R1 to almost twice the input voltage. This gives a considerable improvement in turn-on time and allows the circuit to be used with smaller input voltages than would otherwise be possible.

Components For The Regulator

Most components in the circuit are working under conditions that are not too demanding, and following Auntie Static's advice (ETI September, 1985) I have stuck to general purpose devices. There are a few exceptions.

D3 must have a low forward voltage drop and short recovery time. I chose a Schottky barrier rectifier but a silicon fast recovery device would also do the trick, although the much higher conduction voltage drop of these would mean lower efficiency for the regulator. Components you can't use are signal diodes (when Q5 is off, D3 takes the entire load current) or ordinary rectifier diodes like 1N4001 (the reverse recovery is painfully slow.)

D2 must also have a fast recovery, but as all the requirements are met by a general purpose 1N914 or 1N4148 there is no need for anything exotic here.

The choke L1 is working under difficult conditions. It must pass the entire DC component of the output current as well as dealing with any fluctuations, so care must be taken to avoid saturation of the core. Two other points to watch if you want to make any changes to L1 are that it is one of the components which determines the oscillation frequency of the circuit, and secondly that it can radiate a good deal of EM interference. Remember that the circuit is producing powerful oscillations at low radio frequencies, so interference with other equipment can easily become a nuisance.

Finally, you may wish to choose a different FET. Some desirable characteristics are a low drain-source resistance (r ds(on)) short rise time (tr) and fall time (tf), high forward transconductance (gfs) and low input capacitance (Ciss).

Experiments, Modifications, Etc

If you have never built a switching regulator before, don't just take my word for it that the components should be such-and-such. Replace D3 with a 1N4002 and see what happens!

The circuit could be improved by better output filtering. You should bear in mind that there's no point in increasing the size of the capacitors on the output. About 20mV of ripple is needed to switch the circuit and increasing the capacitor value will just slow down the frequency of oscillation. Any additional filtering will have to be 'tacked on the end' of the circuit.

There are two fairly important things that the circuit lacks: a soft start and overload protection. The idea of a soft start is that the input voltage should be established before the regulator begins to deliver current to the load, to ensure that the regulator will function correctly. The easiest way to achieve this is to introduce a time delay, as shown in Fig. 7. This method is not ideal because the regulator would switch on even if the input only rose to 4V and stopped, but it will cope with normal powering up and most fault conditions. As Vin rises, the transistor in parallel with Q4 holds Q5 off. When Vin reaches its final
value, the capacitor will continue to charge through the resistors until the negative plate drops below the voltage needed to hold the transistor on. The regulator can now start working. The diode to ground discharges the capacitor quickly when Vin falls. The diode in series with the transistor base is to increase the amount of ripple that can be present on Vin without switching the transistor on and off.

The circuit of Fig. 8 is a simple form of over-current protection. At a little over 1A the extra transistor will begin to turn on, limiting the voltage applied to the gate of Q5. This circuit is not ideal as the resistor will waste power at all times. More important, the action of the circuit is to reduce the switching efficiency of Q5 which will get rather hot under overload conditions.

A circuit with possibilities is shown in Fig. 9. The extra transistor allows Q1 to work as normal when Vout is high. If Vout drops, conduction in the transistor is reduced and R begins to take the collectors high. A point will quickly be reached where Q1 can no longer pass any current; Q2 gets it all and holds off Q5 by the normal action of the circuit. Q1 collector will end up at a voltage higher than its emitter. As transistors work ‘back to front’ to some extent, the effective circuit will be as shown in Fig. 10 with Q1 supplying extra current to hold Q2 hard on. The regulator will be locked off and will not supply any current at all.

Naturally, arrangements will have to be made to ensure that the extra transistor turns on when power is first applied. The regulator will then be reset after overload by turning the supply off and on again. With a little ingenuity the transistor could be switched by a sudden drop in the output voltage rather than at a certain voltage level. As good text books say, I leave these as an exercise for the reader. . . .

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**Etiquette**

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**PROJECT:** Regulator

![Fig. 8 Simple current limit.](image)

![Fig. 9 A possible protection circuit.](image)

![Fig. 10 No, the labels aren't wrong.](image)
This one is definitely wacky, but it's also an ingenious and reliable way of measuring large-value capacitors. Thank God it's Faraday, we say, and thanks to designer Ray Bold.

The instrument is capable of measuring capacitors in the range 1µF to 100,000µF, and was inspired by the purchase by the author of a 'goody bag' containing a huge number of unmarked electrolytics, and the prospect of a long tedious exercise using a bridge to measure them. A handy electromechanical counter and some development work led to the construction of a prototype instrument from which this design derives.

Construction is straightforward. The meter uses a bench supply rated at 12V DC and the full load current is in the region of 200mA.

**Theory**

If a capacitor, which is initially discharged, is charged from a constant current source, the voltage across it will change linearly with time. The time taken to charge to a given voltage will be dependent on the size of the capacitor and the magnitude of the charging current. Expressed mathematically,

\[ CV = It \]

or

\[ t = \frac{CV}{I} \]

If we fix V and I, then t will be a function of C, and if we then arrange to measure t, we will also be measuring C.

In this design, a constant current of either 9µA, 90µA, 900µA or 9mA is supplied to the capacitor under test (CUT). The voltage across the capacitor is monitored by a window comparator and, while the voltage is within the limits of the window comparator, a counter operates at approximately 10Hz to give an

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**Fig. 1** Charging time between voltage limits is proportional to capacitance under constant current conditions.
indication of the capacitor's size. Figures 1 and 2 help to illustrate the theory.

Circuit Description

The circuitry around Q1 (Fig. 3) forms a constant current source with the current being selected by SW1. The capacitor under test is connected to the terminals marked CUT. SW2 and R2 remove any initial charge on the capacitor. When SW2 is opened the capacitor charges at a constant current and its voltage increases linearly with time. Since the relationship is CV = It, for a given current, the voltage will rise between two limits over a time determined by the value of the capacitor.

IC1, IC2 and associated components form a window comparator whose output goes low when the voltage across CUT lies within certain limits. The limits can be adjusted by RV1, so providing a means of calibration. The window comparator gates the astable built around IC3 which counts at the rate of approximately 10Hz, driving the counter via Q2 and Q3 as long as it is gated on.

D3 and C1 decouple the supply to the sensitive parts of the circuit and protect them from the interference generated by the counter. D4 suppresses the spikes generated by the counter. Figure 3 shows the voltages at various parts of the circuit during a measurement cycle.

Construction

The printed circuit overlay is shown in Fig. 4 and is mounted using two of the mounting posts in the case, drilled 3mm. The board is drilled 3.5mm and 3mm set screws are used. Another 3mm set screw, with nut and spacer is inserted just above IC1. Take care to mount polarised components and semiconductors correctly, and solder in the semiconductors last. Component values are not critical but if R3 to R6 are close tolerance so much the better. Electrolytics have large tolerances which vary with age and temperature, so extreme accuracy is unnecessary.

HOW IT WORKS

The Millifaradometer consists of four basic sections.

a) A constant current generator based on Q1 and associated components.

b) A window comparator built around IC1 and IC2.

c) A gated astable multivibrator around IC3a and IC3b.

d) A counter and driver transistors Q2 and Q3.

With SW2 closed, the capacitor to be measured is connected across the terminals marked CUT. The switch and 150k resistor ensure that the capacitor is discharged at the outset. When SW2 is opened the capacitor charges up via Q1, and the voltage increases linearly at a rate dependent on the current and size of capacitor.

When the bottom of the ‘window’ (set by the divider chain R7, R8, R9, R11) is reached, the output of the window comparator goes low gating on the astable multivibrator IC3a and b. This runs at approximately 10Hz and drives the counter via Q2 and Q3. After a time, determined by the value of the charging current, the size of capacitor and the width of the window (set by RV1), the output from the comparator goes high and the counter stops.

The charging currents and window width are so arranged that on range 1 the voltage across a capacitor of 10µF will change by 1V/s and during this time the counter will count to 10. A capacitor of 20µF will take twice as long and count to 20 in 2 seconds on range 1. On the X 1000 range a capacitor of 47,000µF will take 4.7 seconds to change its charge by 1V and the counter will register 47.

BUYLINES

The counter used in the prototype was supplied by MS Components, Zephyr House, West Norwood, London SE27 9HL, (01-670 4466), catalogue no. 9036, price £5.75. Its dimensions are: 45.7 x 45.7 x 50.5mm (depth to end of terminals), cut-out size 37x23mm. The case was also supplied by MS Components, catalogue no. 4108, price £2.71. Its dimensions are: 170 x 143 x 56mm. No other components should pose any problems and the PCB will be available from our PCB Service.

Fig. 2 Circuit diagram of the ETI Millifaradometer.

Fig. 3 Voltages at key points in the circuit.
Fig. 4 Component overlay of the ETI Millifaradometer.

Fig. 5 Front panel layout of the ETI Millifaradometer.

**PARTS LIST**

**RESISTORS** (All 1/4w 5% carbon film)
- R1 3k3
- R2 15R
- R3 220k
- R4, 9 22k
- R5 2k2
- R6 220R
- R7 180k
- R8 15k
- R10 3M9
- R11 470k
- R12 10k
- RV1 10k horizontal skeleton preset

**CAPACITORS**
- C1 1000µ; 16V electrolytic
- C2 100n polyester

**SEMICONDUCTORS**
- IC1, 2 3140
- IC3 4001
- Q1, Q2 BC214L
- Q3 ZTX500
- D1 to 4 1N4001
- ZD1 2V7 400mW zener diode

**MISCELLANEOUS**
- SW1 Single-pole 4 way rotary switch
- SW2 SPST toggle switch

Counter, 12V, 10 impulses per second with reset; case (console type); PCB; terminals; pointer knob; red and black 4mm plugs; connecting wire; set screws; nuts, washers, transfers.

Flying leads are used to connect the instrument to the bench power supply or 12V battery.

The panel layout is shown in Fig. 5. If the case is supplied with a protective film on the panel, leave this on until all marking out (with a ballpoint pen) and drilling is completed. Peel off the film just prior to lettering and assembly. This will reduce the risk of damage to the finish. Take care not to mount the counter any lower than indicated as clearance below is minimal. Even so, the terminals may need carefully bending a little, supporting them with pointed nose pliers near the body. After applying the lettering, the finish can be protected with two thin coats of clear lacquer.

**Setting Up**

Find a capacitor of between 10 and 20 µF. If possible, measure it accurately on a piece of commercial equipment. Reset the counter to zero and connect the capacitor. With SW1 on the X1 range switch SW2 to test. The...
counter should operate. Note the reading and repeat the procedure adjusting RV1 to give consistently accurate results. With large value capacitors there is a delay before counting begins while the bottom of the window is reached.

**In Use**

If in doubt about the value of a capacitor, set the range switch to a high range. The counter may count a couple of times or fail to count at all, which will indicate that a lower range is required. Starting on a low range may mean watching the counter for an excessive time until it stops. When counting finishes, simply add the number of zeros indicated by the range switch and that is the value of the capacitor.

To test another capacitor, switch the test switch up, replace the tested capacitor with the one to be tested, zero the counter and select a likely range. Then switch the test switch to test and watch.

After finding the value of a capacitor (bearing the +50%/-20% tolerance in mind) it is frequently possible to find the voltage rating by referring to catalogues, as these often give data including the physical size of capacitors.

The Milifaradometer can also be used to check electrolytics in faulty equipment. It should be remembered that a capacitor which is very leaky will charge slowly and seem to have a large value because of the current shunted through its own internal resistance. Capacitors which behave in this way should have their leakage checked with a multimeter set on resistance.
CHORUS UNIT

Designed to complement the Noise Gate described in our July issue, Ian Coughlan’s Chorus unit offers the maximum of versatility in the minimum of space.

Two of the most popular effects available to musicians are double-tracking and stereo double-tracking, better known simply as echo and chorus respectively. The design to be described offers both of these effects in a simple, compact, battery-operated unit. Double-tracking can be produced by placing a large, heavy metallic plate close to the vocalist or musician, to reflect sound back to the microphone. The result is a much richer, fuller sound. Strictly speaking, this ought not to be called echo since the time-delays involved are too short for the ear to interpret them as such. The subtle phase-differences produced will be interpreted quite correctly as a ‘a large, heavy object close to the sound-source!’ Plates are still popular in the studio and many electronic delay units seek to reproduce their sound or perhaps several types of ‘plate sound’.

Chorus is an attempt to emulate the sound of another instrument or voice playing or singing along in perfect harmony with the first. To understand what is involved in producing this effect, consider for a moment what an extra instrument or voice would sound like — exactly the same as the original. Either heard singly would be indistinguishable from the other, but heard together they will interact, producing subtle phase shifts which our ears then interpret as a second instrument or voice.

The chorus effect can be produced by delaying a portion of the audio signal, slowly varying this part of the waveform.

1C2a is connected as a high-impedance input buffer, and R7 and C5 provide high frequency pre-emphasis, or boost, to the input signal. Part of this signal goes to Mix control R1V, and part goes through 1C2b which has a gain of about 0.3 and then through a low-pass filter built around IC3a. This filter prevents high-frequency components from reaching the BBD line, IC4. IC4 requires two DC bias voltages, and these are provided by the divider chain R12, R13 and R14. The delay-line also requires two anti-phase clock signals, and these are supplied by IC3 which is a phase-locked loop IC, but is used in this design as a simple voltage controlled oscillator. IC4, a TDA1097, is specified for a supply voltage of no less than 12 volts. Because this is a compact battery-operated unit and the supply must be regulated, 5V is all it gets. It will work, but the performance suffers. In particular the attenuation from input to output, typically 0dB at 12V, is very much higher at the reduced supply voltage.

1C3b is configured as another low-pass filter, and gets rid of most of the clock-frequency from the output of the delay-line. Q1, a field-effect transistor, functions as a simple switch to gate the signal through to the next stage depending on whether or not the effect is selected. The next stage is an amplifier with sufficient gain to compensate for the attenuation of the delay-line. The output from this stage goes to the Mix control. It can be seen that at one end of the Mix potentiometer is the un-delayed signal and at the other is the delayed signal. The position of the Mix control determines the proportion of each that appears at the output.

The 1n0 capacitor (C21) across the feedback resistor of IC2c will reduce the high-frequency response of this stage but remember that high-frequencies were boosted at the input stage, so the overall response is fairly flat.

IC6 is the sweep-generator. IC6a is an integrator, and IC6b is connected as a Schmitt. If the voltage on IC6 pin 7 is of a sufficiently high level, pin 1 will also be high. This will cause pin 7 to ramp downwards at a rate determined by R32, C23, C24 and the Rate control. When the voltage is low enough, it will cause IC6b to switch, sending its output low. This will cause IC6 pin 7 to ramp upwards, and the cycle repeats itself. Unfortunately, the linear ramp which IC6 produces is of little use in a chorus. The sweep-generator is used to vary the clock-frequency of the BBD line, and hence the delay-time. There’s no problem when the delay-time changes from, say, 7 to 10 milliseconds, but if the delay-time changes from 17 to 20ms in the same time, the ear hears a not-very-musical ‘whoop’. What is needed is some way of slowing down the rate-of-change of the delay-time as it approaches the 20ms end of the range.

The solution relies on the fact that bipolar transistors have a very non-linear switch-on characteristic at low levels of base current. This characteristic is used to turn the linear output of the ramp generator into something approaching the sawtooth waveform shown in Fig. 3. At the 7ms end of the delay-time range, the rate-of-change is high but inaudible; at the other end, the rate-of-change is much slower. RV3 and RV4 are used to adjust the shape of the waveform.

When the width control is fully clockwise, it can be seen that the signal present at IC6d will appear at the input of the VCO, and therefore the delay-time will sweep over the entire range. As the width control is turned counterclockwise, the input to the VCO is derived more and more from IC2d, whose output is set by the Manual control, RV5. Thus the Width control provides the option of a fully swept delay-time, a fully manual delay-time, or anything in between.

All of the op-amps are supplied with +9V except IC6c, which is supplied with +5V. The delay-line and its clock generator are also supplied with +5V. A 78L05 provides the +5V supply. Some parts of the circuit require half the battery voltage and others require +2.5V. These supplies are provided by potential dividers.
delay, and mixing the delayed signal with the undelayed signal. The two signals are identical, but mixed together they will interact, sometimes adding together and sometimes cancelling out. Our ears are fooled into thinking that what they are hearing is a second instrument or voice.

The Circuit

The heart of the Chorus is a Bucket Brigade device (BBD), in this case a TDA1097 which has 1536 stages or 'buckets'. To get an idea of how a BBD device works, imagine a line of people trying to put out a fire by handing buckets of water along the line, then imagine that they don't pass the buckets but each pour the water from their bucket into the next person's bucket. If we ignore spillage, then it's clear that the contents of the first bucket will eventually find their way into the last bucket, delayed by the time taken to pass along the line.

In an electronic 'bucket-brigade' the buckets are capacitors, the men are transistors and the water is a voltage level. The voltage level on the input to the BBD line will eventually appear at the output, delayed by the time taken to pass through all the 'buckets'. The delay time is dependant on the clock frequency applied to the BBD line.

The TDA1097 requires a two-phase clock and this is supplied by the voltage controlled oscillator section of a 4046 phase-locked loop chip. Two clock cycles are required to shift the input signal through each stage, so the delay time can be expressed as:

$$t(d) = \frac{N}{2f}$$

where N is the number of stages and f is the clock frequency. For chorus, a delay range of seven to 20 milliseconds is about right, and this gives a minimum clock frequency of 38.4kHz and a maximum of 109.7kHz. The TDA1097 is only specified to 100kHz, but no problems were encountered with the prototype.

A BBD line is essentially a sampling device, and as such introduces the problem of the clock-signal interfering with the audio-signal. The clock signal will
inevitably find its way to the output, albeit at a much lower amplitude than the audio signal. The clock frequency never falls below 38.4kHz so it will not be audible. The problem is that as the harmonic or noise frequency components of the input signal approach half the clock frequency, the lower sideband of the clock frequency will become audible. For example, if noise components in the range 10kHz to 40kHz are present at the input to the BBD line, they will mix with the clock signal to produce sum and difference signals. The difference signals will be in the range 1.6kHz to 28.4kHz. Trying to filter out these difference signals is obviously impractical, since doing so would also get rid of most of the audio signal.

The solution is to put a low-pass filter immediately before the input to the BBD line, with a cutoff frequency of around 6kHz. In this way, difference signals can only be produced above 32kHz, and a similar low-pass filter on the output of the BBD line effectively gets rid of those.

A cut-off frequency of 6kHz may seem a bit brutal, but the chorus effect ceases to be audible above this, and in any event, only the delayed portion is so affected: the undelayed signal is not filtered. The filters in this design have an actual cut-off of 6.2kHz and a slope of -20dB/octave. In addition, pre-emphasis on the input and de-emphasis on the output endow this unit with a good noise performance.

In order to achieve adequate chorus effects, the delay time must vary steadily rather than being constant. This is made

**Fig. 2 Component overlay for the chorus unit PCB.**
possible by a sweep facility which repeatedly swings the delay period from its maximum to its minimum and back again. The sweep rate can be varied from once every ten seconds to ten times a second, and the width of the sweep is continuously variable from full sweeps between the limits of 7 ms and 20 ms delay and no sweep at all (i.e., a constant delay period). In the latter condition, the delay period can be manually set to permit the 'plate' effects described earlier. Further versatility is provided by a Mix control which allows the delayed and undelayed signals to be mixed in any proportions. This makes it possible, for example, to use the full amount of sweep but still achieve a very subtle effect.

Power to the unit is provided by PP3 size battery (preferably alkaline), and a socket allows connection to an external supply. The unit is switched on by inserting a mono jack plug into the input socket, and Effect or Bypass mode is selected by the built-in footswitch or by a remote switch connected to the REM socket. An LED indicates when the unit is in Effect mode.

**Construction**

Commence assembly by installing the wire link, the four jack sockets, and if desired, sockets for ICs 4, 5 and 6. Sockets cannot be used for ICs 2 and 3 or they will interfere with the potentiometers when the unit is assembled. The jack sockets must be of the recommended type if they are to fit correctly into the prepared holes in the PCB.

Continue assembly by soldering into place the resistors, capacitors, and presets, taking care that the capacitors near the connector end of the board are mounted flat so as to clear the potentiometers. Next fit the diodes, transistors, and ICs 1, 2 and 3. Cut to length three pieces of ordinary insulated connecting wire, and solder them between the points shown on the PCB overlay, then fit the two battery-guide pillars and the battery connector. Connect the four potentiometers, R28 and the LED to the PCB using insulated wire. Lastly, fit ICs 4, 5 and 6. The board can now be tested.

An oscilloscope is essential if the chorus unit is to be accurately set up, so if you don’t own one you will have to borrow or otherwise acquire one before proceeding further.

Connect the oscilloscope input to IC5 pin 2 and check that a square wave signal is present with an amplitude of about 5 V peak-to-peak. Turn the Width and Manual controls fully anti-clockwise and adjust RV6 until the frequency of the square wave is about 38.4 kHz. Turn the Manual control fully clockwise and check that the frequency rises to about 109 kHz. If either of these frequencies are outside the range of adjustment of RV6, try altering the value of R43.

Connect the oscilloscope input to IC6 pin 7 and check that a triangular waveform is present with an amplitude of about 2 V peak-to-peak. Rotate the Rate control (RV2) and check that the frequency varies from about 0.1 Hz when it is fully anti-clockwise to 10 Hz when it is fully clockwise. Move the oscilloscope probes to IC3 pin 14 and check that the waveform present is similar to that shown in Fig. 3. Make any
necessary adjustments, using RV3 to set the amplitude and RV4 to alter the offset.

Return the oscilloscope connections to IC5 pin 2 and set the Width control fully clockwise and the Rate control fully anti-clockwise. The frequency of the signal should be slowly changing between the previously-set limits of 38.4 kHz and 109 kHz. Carry out any fine tuning required using RV3 and RV4 and the setting up is complete.

If you want a particularly compact unit it is best to use the recommended box, but if you are unable to find the correct type or for any other reason wish to use a different box, choose one that is slightly larger than our prototype so as to avoid cramping the components.

Refer to Fig. 4 and drill the necessary holes as accurately as possible. Take particular care with the holes for the switch and the sockets since these must line up with the components on the PCB. If you have any doubts about your skills in this direction, try drilling the holes a little smaller than is required in the first instance, then offer up the PCB to check that they coincide and make any necessary adjustments with a small file before enlarging the holes to their final diameter.

When drilling is complete, rub down the outside of the box with glass paper to deburr the holes and prepare the surface. Clean the box thoroughly and then prime and paint it allowing suitable drying times. Loosely assemble the potentiometers and knobs as a guide and use rub-down lettering to apply the legends. Remove the fittings, lightly buff the surface to remove any fingerprints, etc, then apply a coat of clear varnish and leave to dry.

Glue a piece of foam rubber inside the box to prevent the battery rattling around. Mount the switch in position through the appropriate hole in the case but do not tighten the fixing nut. Mount the potentiometers through the top panel of the case. Place fibre washers onto each of the sockets on the PCB, then offer the board up to the case, socket end first. Loosely assemble the socket securing nuts from the outside to stop the board slipping back through, jiggle the switch and the PCB until the switch pins appear through the holes in the board, then solder them to the pads and tighten the switch into position. Complete the construction by adding the switch cap, the knobs and the base plate and tightening the securing nuts on the sockets. Don't forget to install a battery!

Apply an input signal of about 1kHz at a few hundred millivolts to SK2. Inserting the jack plug should turn the unit on. Connect up an oscilloscope so that you can alternatively observe the input signal and the output signal from SK4. Better still, if you have a dual-beam oscilloscope use it to monitor both at once. Press the footswitch if necessary until the Effect LED is on and check that the output waveform closely resembles the input waveform. Press the switch again and the output waveform should start to subtly alter in shape. Operating the various controls should influence the nature and extent of the alteration.

Having got the pretend stuff out of the way, you can now plug in a real instrument, hook the unit up to an amplifier and get chorused away!
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Components

EMI ELECTRO-MECH INDUSTRIES
ENLARGER EXPOSURE METER

Bring a little light to your darkroom operations with this meter, designed by Doug Bollen

Despite fully automatic cameras and simplified colour processes, many amateur photographers still perform prodigious feats of mental arithmetic in dark places every time they make some minor adjustment to their enlarger. With monochrome prints it is a fairly simple matter to run off a test strip in open dishes to assess exposure before creating each finished print, and there is no colour balance to bother about. A colour test strip, on the other hand, can take all of fifteen minutes to produce, plus time to dry the drum. It uses a disproportionate amount of expensive colour chemicals, and several colour test strips may be needed to establish correct exposure and colour filtration for just one frame of a film.

There is no substitute for human evaluation in photography but a good exposure meter can reduce time and wastage of materials by establishing a reference point against which differing enlarger magnifications and film densities may be judged. The ideal exposure meter would have equal scale divisions which are readable in complete darkness, adjustment for a wide range of photographic paper sensitivities, good accuracy over the entire visible spectrum, high sensitivity to low light levels, no dependence on temperature or supply voltage, a very small light sensitive cell area and, of course, a very low cost.

Stop On Sight
It is in the nature of light sensors that they stubbornly defy attempts to achieve equal scale divisions, even when using moving-coil meters with specially shaped poles. Another problem is that light sensors tend to have a wide tolerance, typically ±40%, which makes conformity to a precalibrated scale on a mass produced instrument difficult. On casual inspection it might be thought that the requirement is for a linear time scale, say 1-100 seconds, but that is not the case. Exposure time and light intensity on cameras and other photographic instrument scales both proceed in a geometric series (because that is the way human vision works) where each basic division represents a doubling or halving of time or light. Light itself is not so easy to measure because a large number of variables have to be allowed for. However, a unit commonly employed to assess relative light levels is the arbitrary ‘stop’ for which doubling a reference exposure figure means halving the light and is designated +1, while halving the figure means doubling the light and is said to be −1. The overall accuracy of camera plus film stock is seldom better than 1/3 stop either way. An exposure meter should at least equal this performance over a substantial portion of its scale. In practice, the exposure meter can be conveniently calibrated in seconds in steps of 1, 2, 4, 8, 16 and so on and in equivalent stops, say −4 to +4.

A popular type of low-cost exposure meter circuit is shown in Fig. 1, where the voltage output from a Cadmium Sulphide cell (Light Dependent Resistor) and load resistor, R, is compared with the voltage at the slider of a conveniently calibrated potentiometer, RV. A comparative reading is taken at null point and adjustments are made by varying the enlarger lens aperture. This technique neatly avoids the need to calibrate the exposure.

The prototype exposure meter
plot was made of LDR voltage (Fig. 2) against R for measured cell resistance corresponding to lens f-stops. Doubling the R value produces a shift of one stop without changing the meter scale span, so a variable resistor for R will act as a paper sensitivity control. The plot also shows that the meter will cover 8 stops without too much cramping of extreme scale gradations, plus an additional 4 stops with adjustment of R. This is an arithmetic range of 1:4096 in light intensity.

**Construction**

The exposure meter is simple enough for a photographer with a limited knowledge of electronics to build. The complete novice should, however, study the art of soldering and pay particular attention to the orientation of polarised components (IC1, C1, C2, C3, LED1, LED2 and B1). If any of these are the wrong way round, malfunction or disaster may follow! Make sure that RV1 and RV2 connections are correct or calibrated scales may be reversed (Figs. 4 and 5).

The PCB will fit inside a 3.5" wide by 2" deep slotted box, but if this size is not available, drill holes at the corners of the PCB and bolt to the bottom of any suitable box with stand-off spaces and the LEDs pointing straight up. The dial for RV2 should be as large as possible. Obtain a knob with a screw-on skirt and replace the skirt with a disc of perspex. A paper disc should be temporarily placed above the perspex to allow calibration (Fig. 1). It can be sandwiched between the perspex and an aluminium disc for protection after calibration. Aluminium was also used for the LED shield (Fig. 6). The knob for RV1 can be a standard 0 – 10 component. The LDR is simply glued inside a slim light-tight box over a hole slightly smaller than the LDR case.

**Testing And Calibration**

Assemble the instrument, switch on and check that both LEDs are working. Plug the LDR, set RV2 to mid-track and point the LDR towards a source of low light. Adjust RV1 for balance. If balance cannot be obtained, the light level may be too high. Check that RV2 is rotated anti-clockwise to balance if there is a reduction in light, and RV1 similarly clockwise.

Lightly glue a rectangle of white card with a 1/4" hole over the LDR aperture (this can later be reduced or increased if required). Place the sensor on the enlarger baseboard, set the head to mid-height and the lens to maximum aperture. Set RV1 to 2. Now click stop the lens down and trim RV1 slightly to obtain a null. Mark the RV2 setting with a soft pencil and the figure zero. Without touching RV1, click stop the lens up and down and balance with RV2. At each balance point mark the paper disc. You should now have part of the scale marked. Return RV2 to mid-position zero and adjust head height and click stops to fill in the remainder of the scale (leaving RV1 where it is). Slight errors will occur where there is overlap with previous dial markings. Average these errors out over several successive settings and different overlaps. It should be possible at this stage to detect if a particular lens f-stop deviates from true. If errors greater than 1/3 stop are encountered, or if best accuracy is required, the exposure meter can now be double checked with a camera lens.

You'll need a camera with manual override and a means of holding the shutter open (‘B’ or ‘bulb’ setting). Tape the LDR box in the film plane with the sensor aperture central. Place the camera on the enlarger base board with the lens up and the focus on near. Use a piece of plain white paper to cover the lens. Do not strain the open camera back. Adjust camera f-stop, enlarger magnification and click-stop for
null with RV2 at mid-point zero. Now vary the camera f-stop to check RV2 calibration.

When satisfied that the paper dial is accurately calibrated in stops — 4 to +4 for the CdS cell you are using, it only remains to make an ink copy and interpolate intermediate settings from Table 1.

Using The Meter

There are many ways of using an enlarger exposure meter but only one will be described here: ‘pegging the mid-tone’. After obtaining a test strip of a reference negative or transparency and deciding on the correct exposure for a given magnification and aperture — say, 16s at stop 0 — measure the blackest and whitest areas of a new subject in stops and determine the mid-point. For example, a transparency may give +2 and −3 so the mid-point would be −0.5, which gives an exposure time of 16s x 2−0.5 or 16x√2, which equals 11.3s. Notice that the arithmetic mean, (16s x 20.5 + 16x√2)/2 = 33s, is hopelessly wrong.

Don’t forget to allow for reciprocity failure (decrease of paper sensitivity) when making very long exposures at high magnifications or through very dense negatives. The correction is typically (x1.5) at 40s (x2.3) at 100s and (x3.5) at 200s.

<table>
<thead>
<tr>
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<tr>
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</table>

Table 1. Stops related to exposure times.

The circuit of Fig. 3 is the result of refinements made to the rudimentary LDR bridge circuit. RV1 is now the variable load resistor for the LDR. High input resistance op-amp buffers IC1a and IC1b ensure that there is negligible loading of either the LDR or calibrated pot RV2 when taking measurements. C1 and C2 filter out unwanted mains ripple. The purpose of R1 and R2 is to reduce the input level to IC1a and IC1b to prevent ‘bottoming’ of amp outputs. IC1c acts as a simple comparator routing the brilliance of red and yellow leds, LED1 and 2. At null point the overall light output changes from red to yellow, or visa versa, thus providing continuous ‘off baseboard’ illumination for RV1 and RV2 scales. When the battery voltage falls to about 6V the yellow led will extinguish to indicate battery low. The circuit of fig. 3 is virtually unaffected by battery voltage and so needs no regulator. The overall temperature coefficient of the circuit, including LDR, is +1/4 stop for a 10° C rise and maximum sensitivity has been estimated at around 0.1 lux. Depending on the light source efficiency of the enlarger used, CdS cell aperture can be fixed between 4 to 10 mm dia.

Any reasonable quality enlarger lens, or camera lens, can be used as a calibration source for the exposure meter, as will be outlined later. The enlarger exposure meter is perfectly suitable for spot measurements from black and white negatives and integrated measurements (through a ‘diffuser’) of ‘average subject’ colour shots. Spot measurements of coloured areas can also be made, but this requires some expertise, and three readings for each spot measurement through primary colour filters to achieve a meaningful result.

<table>
<thead>
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<th>PARTS LIST</th>
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<td>RESISTORS (all 1/4W, 10%)</td>
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<td>R3, 5</td>
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<td>C3</td>
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<td>IC1, 2</td>
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<tr>
<td>LED1</td>
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<tr>
<td>LED2</td>
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<tr>
<td>LDR</td>
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<td>MISCELLANEOUS</td>
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<tr>
<td>PL1, SK1</td>
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<tr>
<td>B1</td>
</tr>
<tr>
<td>SW1</td>
</tr>
<tr>
<td>PCB, connecting wire, PCB pins, spacers, battery clip, knobs, boxes, perspex, paper and aluminium discs to suit box (see text for details), small piece of aluminium, 14-pin DIL socket.</td>
</tr>
</tbody>
</table>

Fig. 3 Full circuit of exposure meter.

Fig. 4 Component overlay

ETI NOVEMBER 1985
PROJECT: Exposure Meter

Fig. 5 Front panel and LDR connections.

Fig. 6 Light shield.

Fig. 7 Specimen scale.

COUNTERS

TF800
Bench/Portable: large 8-digit LED display. Frequency range 5Hz to 600MHz. Resolution 0.1Hz. Sensitivity 10mV rms. Timebase accuracy ±2ppm. Battery life 200 hours. Frequency average period, totalize & reset. 2 ranges. Self-tapping screw. Complete with mains adaptor.

TF200
Bench/Portable: 8-digit Liquid Crystal Display. Frequency range 10Hz-200MHz. Resolution better than 1ppm. Sensitivity typically 10mV rms. Timebase accuracy 0.3ppm. Battery life 200 hours. Frequency average period, totalize & reset. 2 ranges. External clock facility. Complete with batteries.

TF040
Bench/Portable: 8-digit Liquid Crystal Display. Frequency range 10Hz-40MHz. Resolution 1Hz. Sensitivity 40mV rms. Timebase accuracy ±0.5ppm. Battery life 80 hours. Frequency, totalize and reset. 2 gate times. Complete with batteries.

PFR200A
Pocket size: 8-digit LED display. Frequency range 20Hz-200MHz. Resolution 0.1Hz. Sensitivity typically 10mV rms. Timebase accuracy ±0.5ppm. Battery life 10 hours. Frequency, 2 ranges, 4 gate times.

TFR800 PRESCALER
Frequency range 40MHz-600MHz. Sensitivity 10mV rms. Powered directly by TF200 or TF040 (leads supplied).

TFR1000 PRESCALER
Frequency range 100MHz-1000MHz. Sensitivity 25mV rms. Will extend TF200 and PFR200A capability beyond 1GHz.

For further information contact Thandar Electronics Ltd. London Road, St Ives, Huntingdon, Cambs PE17 4JH. Telephone (0480) 68746. Telex 32250.

ETI NOVEMBER 1985

www.americanradiohistory.com
Beat your own drum with this status cymbal. Electronic noises off, and on, by D. Stone.

The circuit associated with Q1 is the white noise generator, amplifying the noise produced in the diode D1 as a result of reverse leakage current. A germanium diode is used as the leakage current is higher than that of silicon for a given voltage, producing a higher noise signal level. This amplified signal is coupled by capacitor C1 to a further amplifying stage built around the op-amp, IC1a. IC1b is a constant bandwidth, bandpass filter. The filter centre frequency is set by the dual-ganged potentiometer, RV1. This is a classic second order bandpass filter of the multiple-feedback tuned type, which tunes the white noise into variable frequency range noise. SW2 is to choose either the filtered noise or the unfiltered white noise. The filtered noise is inaccurately described as pink.

The voltage controlled amplifier (VCA) is built around the 3080E transconductance op-amp, IC2. Output current is a function of the control current fed to pin 5 of the package and the difference in voltage between the two input pins. The output current of the device is converted into a signal voltage by R27 and the signal is capacitively coupled to the output by C13.

The signal from the microphone is first amplified by Q2 and then fed to a pulse amplifier built around IC1c. This section inverts the signal and amplifies it to give a negative going pulse at its output whenever a sound is picked up. The duration for which the pulse remains negative depends, to some extent, on the volume of the input signal to the mic. This gives some sensitivity to the impact of a beat. The negative going pulse is led to Q3 which, with C12 and its associated circuit, forms a simple envelope generator.

When the pulse is received by Q3, the transistor turns on. C12 charges up rapidly through the transistor, D2 and R25, giving the fast attack which is characteristic of a drum. The transistor then turns off and C12 discharges through R25, RV2, D3 and R21. This gives a variable decay, considerably slower than the attack. The voltage is converted into a current by R26 and fed to pin 5 of IC2, the VCA.

C6 provides necessary supply capacity to eliminate any power thump which may find its way into the circuit when the drum is struck. C7 provides high frequency supply decoupling. A false signal earth is supplied in the form of a decoupled 4.5V rail. This rail is formed by C3, C4, R4 and R5, and eliminates the need for a two battery split rail supply.

There is a spare op-amp on IC1 available, should any adventurous constructor feel the need to expand the unit in some way.

HOW IT WORKS

Construction

The construction of the unit is straightforward if the recommended PCB is used. The assembly should follow the usual format of passive components first, followed by the semi-conductors and integrated circuits. The use of IC Sockets is recommended to prevent damage to the chips by overheating and to ease the removal of chips should this become necessary. The microphone was fitted to the circuit board near IC1 in the prototype with double sided sticky fixing pads.

The prototype unit was housed inside an 8” Tambour. The Tambour has a removable drum skin which allows easy changing of the battery. Tambours can be bought from any good music shop. A base was fitted to the Tambour

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The prototype unit was housed inside an 8” Tambour. The Tambour has a removable drum skin which allows easy changing of the battery. Tambours can be bought from any good music shop. A base was fitted to the Tambour
The PCB fits neatly inside a tambour case.

with white modellers' 'Plasticard'. The whole unit was then sprayed with enamel paint and labelled with rub down letters. As an alternative housing, for a number of these units, matching two litre ice-cream cartons could be used, and give a 'space-age' effect when painted in suitably futuristic hues. Mounting the units in playing positions is left to the reader.

Fig. 1 Circuit diagram of the crash cymbal.

Fig. 2 Component overlay of the crash cymbal.
THE SECOND LINE OF DEFENCE

Vivian Capel concludes his reflections on domestic security with a design for an alarm system which represents a radical departure from most current thinking — it's simple and effective and there's not an IC to be seen.

This alarm system was born of experience and developed through six models over several years. It was originally inspired by an unwelcome visitation, after which — following the time-honoured practice of bolting the stable door after the horse had fled — the designer went to purchase a commercial system, only to discover snags in all the alarms then on offer.

The considerations that went into the design were all dealt with. The upshot was to go for a battery-operated device, with a low current drain in the on-guard or stand-by mode and a moderate current drain when the alarm was sounding. For reasons that have already been well-rehearsed in these pages, the designer settled for a conventional loop circuit (perimeter) device in which the loop would consist of a number of magnetic door and window switches wired in series.

Defensive Postures

The current requirement is met in the alarm. In some systems, currents of several milliamps are common. In this circuit, the 'stand-by' current is around 0.1mA and the batteries, two 9V6 lantern units in series, with chocolate block connectors screwed on to the stranded end of the terminals for wiring up, can last up to two years with the alarm set every night.

The main circuit contains only one active component — a BC108 transistor. This configuration was arrived at after much pruning of an original, considerably more complicated, circuit. The object was to increase reliability without sacrificing any required feature or diminishing performance.

The circuit includes a facility for pressure pads to be placed at strategic points around the premises. These are normally open devices, so they cannot be included in the loop which triggers the alarm by being broken. As with most similar systems, the pressure pads are included in a separate circuit and are, therefore, optional.

The problem of leaving premises which have been armed is overcome in this circuit by means of a large-value capacitor fitted across the section of the loop which covers the entrance. This will give a delay of about eight seconds after the door is opened, and once the alarm has been set, before the bell is triggered.

Obviously, this delay will operate on entry as well as exit and the alarm simply resets itself when the door is closed. This is useful to allow people out and in without the alarm going off, but might just allow a lucky or carefree burglar to stroll through the door with impunity. The obvious solution is probably the best: insert a key-switch in series with the delay capacitor and located in the front door. This should defeat all but the most determined burglar. The key-switch will take the capacitor out of circuit, thus removing the delay.

The alarm also features a panic-button facility, useful in case of attack in the home or for the elderly or infirm.

After considerable research and testing, the Tann Synchro B6D12 bell was found to be the most suitable available. Current consumption averages out at 95mA, which gives a total consumption for two bells in

Fig. 1 Circuit diagram of the alarm. Look, no ICs...
parallel of less than a torch bulb. The sound output is very high — about 88dB at 3m. Of course, any 12V bell or siren of suitably low current consumption and high sound output may be used in this application. RS stock a number, including an own-brand bell which is specified at 80mA and 88dB at 3m (for the £5.29 unit), and a useful low current siren which is specified at 20mA and 107dB at 1m (RS 249-924). Maplin also stock a 12V bell (YK 85)G with a claimed current consumption of only 60mA. Remember that the 3m output figure is always 10dB less than the 1m figure. RS components can be ordered through any dealer with an account or through Crewe-Alien Ltd., whose address is given in Buylines. A good guide to bell or siren quality is British Standard BS 5839: ‘Fire detection and alarm systems in buildings’. You should not have to pay much more than £12 for any suitable bell or siren.

False Alarms

Safeguards against false alarms are designed into the circuit, and, in that respect, it is deceptively simple. One common problem with perimeter loops is that they can act as untuned aerials picking up all manner of signals which might be rectified by the semiconductor input and trigger it. Two bypass capacitors are wired across the base-emitter junction of the main circuit transistor, one a large value for lower frequency RF and one a small, non-inductive type for higher frequencies. This gives a high degree of immunity from unwanted signals.

Another capacitor wired across the main loop itself gives a slight delay when the alarm is triggered by a break in the loop. With the stated value (100µF) this amounts to 0.8 seconds, enough to ensure that temporary breaks in the circuit — caused, for example, by the wind or by rattling doors moving magnets and their corresponding reed switches apart — do not trigger the alarm.

Components And Construction

The specified reed relay and key switches were all obtained from RS and are listed as such in the parts list. Flush contact and surface mounting reeds for alarm applications are also obtainable through Maplin (FK77), £1.35, and YW47B, £1.95, respectively). Door loops (YW48C, £2.39), window foil (YW50E, £1.19), foil terminators (YM51, 65p), pressure mats (YB91Y, £3.55) and stair pressure mats (FK79L, £2.25) are also available from Maplin. In fact, Maplin will supply almost all the hardware you require, including panic buttons, heat detectors, vibration sensors and junction boxes.

HOW IT WORKS

The base of Q3 is tied to the emitter by the two loops, and is non-conductive in the stand-by mode. The only current passing in this condition is through R1, the 120k base resistor, which across 12 volts is 0.1mA. C4 and C5 are the RF bypass capacitors.

SW3d connects Q3 emitter to the negative rail when the alarm is on. If the main loop is open, C1 starts to charge through R1 and takes 0.8 seconds. The base becomes positive and Q3 turns on, energizing the relay, R1, in its collector circuit. One pole of the relay switch connects the bottom end of R1 to the negative rail when the alarm is on. SW3c, thus latching it on and cutting out Q3. The other pole switches the supply to the bells through SW3a and SW3b.

When the loop is opened, the same action occurs but C2 charges, and being ten times the value of C1, takes 8 seconds to do so. If the loop is closed within this time it discharges C2 (through R6, for current limiting) which will start to charge again with any re-opening of the daly loop. SW1 is a normally-closed test button in series with the loops. Pressing it gives immediate triggering. SW2 switches C2 out of the circuit, thus avoiding the delay.

The open circuit which serves the normally-open pressure mats and panic buttons, connects the relay bottom end to the negative supply, thus energizing it directly. In the event of Q3 failing and not triggering when an intruder enters, the pressure mats and panic buttons would still be on guard.

SW3c, which at first glance seems unnecessary, serves to delatch the relay when switching between positions, since the switch must be break-before-make. Without this there would be no way of switching off after the test button has been used without turning the circuit off.

In the test position, SW3a and b place the two bells in series with an LED and current limiting resistor R4. On triggering the circuit with the test button, the LED will light only if both bells and their wiring are continuous. The current is insufficient to sound them. As the relay, switching, and Q3 must also be in order to light the LED, it will not light before the button is pressed, so the loop, too, is tested. Only pressure mats and panic buttons are not checked this way.

Q1 and Q2 are connected as a darlington pair across the relay coil and Q3 base-emitter junction via one pole of the relay switch. The relay is switched at 6 volts, obtained from the 12V supply via R3 which forms a potential divider with the relay coil (375Ω). When the relay is first energized, Q1 and Q2 do not conduct and have no effect. However, C3 starts to charge slowly through R2. When the voltage across it rises, Q1 is biased on, the transistors begin to conduct and shunt the relay coil reducing the voltage across it. When the voltage drops to about 2V, Q4 also becomes biased, switching off the bells. The darlington pair are also disconnected from the negative supply, and this causes C3 to discharge quite rapidly into the base circuit of Q1.

The circuit is now reset and ready for any further action. If the loop is still open triggering will be almost immediate, so the bells will continue to ring. Time for the bells to switch off is about 10 minutes. If a shorter or longer time is required the value of C3 should be decreased or increased accordingly.
them go on the PCB except the loop and panic switches, LED and test button. Connections are made using PCB terminal strips.

**Operation**

After installation, shut all controlled doors and windows and turn all shunt switches to the open position. Turn the switch to ‘test’. The LED should not light up. If it does, the loop is not continuous. Check and correct. Once the LED does not light up, leave the switch on ‘test’ and press the ‘test’ button. This gives a silent test of the bells. The LED should light on this test to show that the bell circuits are continuous.

**PARTS LIST**

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<td>R2 6M8</td>
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<td>D1 1N4001</td>
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<table>
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<th>MISCELLANEOUS (see text)</th>
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<tr>
<td>SW1 normally closed push-button switch</td>
</tr>
<tr>
<td>SW2 key switch</td>
</tr>
<tr>
<td>SW3 4-pole, 3-way rotary switch or key switch (see text)</td>
</tr>
<tr>
<td>RL1 double-pole DIP reed relay (see text)</td>
</tr>
<tr>
<td>Two 12V low-power bells or sirens, 4 off 4-way</td>
</tr>
<tr>
<td>PC terminal blocks, case, PCB, bell pushes, pressure pads, flush or surface mounted magnetic switches and connecting wire as required.</td>
</tr>
</tbody>
</table>

**FYXILINES**

The problem components are all discussed in the text at some length. The addresses of the suppliers mentioned there are:

- RS Components, PO Box 99, Corby, Northants, NN17 9RS (0536 201234)
- Farnell Electronic Components, Canal Road, Leeds LS12 2TU (0532 636311)
- Maplin Electronic Supplies, PO Box 3, Rayleigh, Essex, SS6 8LR (0702 532911).

Maplin sell to individuals and they have a number of shops whose addresses are given on the Maplin ad in this issue. RS and Farnell components should be obtainable through any supplier who maintains an account with these companies. RS components are also available on order from Crewe Allen, 51 Scrutton St, London EC2.
DIGITAL SOUND SAMPLER

In the latest episode of this everyday story of sample folk, Paul Chappell looks into the technique we'll be using to convert sound into numbers.

The first stage in designing any digital sound sampler is to determine how to do the analogue-to-digital conversion. For all the reasons rehearsed in ETI, September 1985, we’ve decided to use a companding A-to-D converter. This will give us effective 12-bit resolution in an 8-bit system — good enough for the quality reproduction of sampled sounds. Last month’s article, ‘The Sample Life’, goes into some of the finer theoretical points of the companding A-D converter, so I won’t repeat them here.

The conversion circuit looks a bit complicated (Fig. 1) and you will note that it uses a companding D-A converter, the Am6070 or DAC88. Why didn’t we use a single IC? The answer, my friends, is that nobody makes one, so we have to make our own from the DAC and a few other ICs.

Conventional Conversion

The circuit is not particularly revolutionary. To tell you the truth, I pinched it straight from a data sheet since I can’t see any way to improve it. The only change I have made is to specify a precision comparator in place of the 311 type suggested in the data. In normal (non-companding) A-to-D conversion, the only results of comparator offset voltage would be to shift the DC level of the converted signal by a few tens of mV and to cause the onset of clipping to occur marginally earlier than otherwise. Nothing to worry about.

Fig. 1 Circuit diagram of the companding A-to-D converter.

HOW IT WORKS

The SAR, IC4, is a general purpose register for A-D conversion and does not make any special provision for sign and magnitude representation, so a little extra circuitry is needed to use it to set the sign bit correctly. Conversion begins when a logic 0 is applied to START, IC4 pin 10. The SAR sets its outputs to 01111111 for the first trial on the next rising edge of the clock. At the same time, the Q output of IC3a, half a 7474, is set to 0 (because CONVERSION COMPLETE will be low after the previous conversion cycle and will not go high until the clock falls again).

The low on the IC3a, pin 5, sets the ENCODE/DECODE (E/D) input of the converter, IC5, pin 1, to ‘decode’, cutting off any current to the encode outputs (IC5, pins 14 and 15). This means that the comparator, IC1, is comparing the input voltage directly with ground (via R1) and will switch high if the input is positive and low if it is negative. If you work through the logic levels on the EXCLUSIVE OR gates — IC2a and IC2b — you’ll see that this is passed unchanged to the input of the SAR, IC4 pin 7.

The next time the clock goes high, this value is clocked into the MSB position of the SAR (IC4 pin 14), giving the correct sign bit. At the same time the next bit is set to 0 for the first voltage level trial. The 7474 will also change state because CONVERSION COMPLETE, IC4 pin 2, will have gone high. The encode outputs will now be allowed to pass a current as the D-A conversion takes place, and the conversion continues as described.

One slight complication is that if the input voltage is negative, we begin the trials with the lowest possible voltage (highest magnitude but negative sign) and work upwards. This means that the comparator output must be inverted. The EXCLUSIVE ORs (IC2) take care of this.
In companding conversion, however, it's a different story. A DAC88 is going to cost you around £16 and you are paying your money for the precision with which it will handle small signals. To make the best use of it, you want small signals to be firmly centred on the high resolution section, or chord, of the DAC's conversion characteristic. As this chord consists of quantisation steps spaced 0.025% of full scale apart, the entire chord will only stretch 3.75mV for a 1V FS. Beyond this point the quantisation steps double in size, resolution is halved and quantisation noise doubled. A small voltage offset cuts the small-signal performance by 50%.

I am a bit dubious about using a 311 in this situation. It has a maximum voltage offset of 7.5 mV and even if this is trimmed out it will still drift with temperature and time. Another reason for being doubtful about the 311 is that for a fairly fast conversion rate — a 1MHz clock, say — its slow response time is pushing the performance of the whole loop quite close to its limits. If we do a quick sum of the delays around the loop, the total of the gate delay for the 7486, setup and propagation delay in the 2502 comes to roughly 80ns; settling time for the DAC will be about 50ns, leaving a little over 400ns for the 311 to respond. Typical response time for the 311 is given as 200ns, which seems OK. Looking into the situation a little more closely (and remembering Auntie Static's warning about 'typical' figures) we find that this will be very much slower in conditions of small overdrive voltage — the time when we are particularly concerned that it should be at its best.

The contribution of R1 and R2 to any error is far less than it may seem at first sight. With an ideal converter, changes in value would only alter the effective gain of the converter. With a real comparator, offset voltages can be introduced by variations in the input bias currents. These can be kept to a minimum by keeping R1 and R2 low in value: 1% tolerance is perfectly adequate to keep offset from this source far below that from other causes.

Approximate Truth

The converter circuit produces an output in sign and magnitude form. The most significant bit of the binary word will be a 1 if the instantaneous voltage is above 0V, 0 if it is below. The remaining seven bits represent the 'distance' of the input from 0V.

Just suppose for a moment that we have a voltage which is rising from 0V to full scale, which we'll say is 1V just to make the sums easy. As the voltage is positive, the sign bit will be a 1. The next three bits represent the chord number. Initially we will be on the highest resolution chord, so they will all be 0. The remaining four bits represent the step within the chord and at 0V these will also be 0.

The first chord goes in steps of 0.025% of full scale — 250µV steps with a 1V FS, so that when the input reaches 250µV bit 0 will be set to 1. As the input voltage rises, the lower four bits continue to count up in 250µV steps until at 3.75mV they are all set to 1. As you would expect, the next step is for the count to advance from 10011111 to 10010000, but instead of 250µV it takes a 500µV rise to get there because we are now in the second chord and the step size has doubled. The lower four bits count in 500µV steps until 10011111, the end of the second chord, and it now takes a 1mV rise to reach the next count; we are in the third chord. The fourth chord goes in 2mV steps — and so on, until at a count of 11111111 we reach the sixteenth step of the eighth chord, which will be the 1V full scale.

The circuit performs the conversion by a process called successive approximation. In essence this consists of first setting up the highest possible number in the SAR, or successive approximation register, (ignoring the sign bit for a moment, this will be 11111111). Starting with the most significant bit, each bit is taken low in turn, the code is converted by the DAC and the resulting analogue voltage is compared with the input. If the input is higher, the bit being tested is set back to a 1 again. If the input is lower, it is left at 0. In this way, after testing and setting all seven bits, the closest number above the value of the input will be left in the SAR and the conversion will be complete.

A Quick Sample

As a quick sample, let's take a four-bit linear conversion with a full-scale of 1.5V. Let's suppose the input is 701mV (you'll see why later). The first step is to set the MSB to 0, giving 0111. This is converted to an analogue voltage (700mV) and compared with the input. The input is higher, so the MSB is set back to a 1. The next bit is then set to 0 giving 1011, the analogue voltage (1V) is left at 0. The next trial is with 1001; the input is still lower so this bit too is left at 0. The final trial is with 1000; the input is still lower so the LSB is also left at 0. The conversion is complete and we have our digital code: 1000.

If you've got a calculator and enough patience, you can convince yourself that this will also work with companded conversion. In fact it will work with any non-linear conversion as long as the DAC is monotonic.

Thankful

You may also have noticed that we rejected the first trial of 700mV, and ended up with a code representing 800mV for an input of 701mV. As I mentioned before, the successive approximation technique gives the nearest value above the input. In linear conversion it is usual to give a half step offset to the converted wave to restore the correct DC level. In companding conversion the situation is more complicated as the offset will alter from chord to chord. Thankfully, the DAC88 takes care of this internally.

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The foil pattern for the Modular Test Equipment waveform generator board.

The pattern for the Millifaradometer board.
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The chorus effect unit board.

The Rhythm Chip foil pattern.
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The switching regulator foil pattern.

The foil pattern for the Second Line of Defence alarm system.
TRAINS OF THOUGHT

For many years railway modellers have been installing 12V lamps in their coaches connecting them to power pickups on the wheels and running line. Although the trains are very effective if the room lighting is turned down to simulate dusk. This arrangement, however, suffers from three disadvantages:

1. Lamp brightness varies with train speed and is nil when trains are stopped;
2. The lighting cannot be turned off;
3. Modern coaches mostly use polystyrene construction, with horrific consequences if it comes in close contact with a hot lamp!

Using command-control systems such as Hornby's Zero-1 eliminates the first problem and makes constant brightness possible. An accessory control module overcomes the second problem but the third is more difficult.

For those of us who use conventional (albeit electronic) control system, two methods of coach lighting are possible. The first is to superimpose on the track a high-frequency AC voltage, eg. 6V at 30 kHz. The high frequency has minimal effect on the loco motors which are primarily inductive devices. Placing a 470n capacitor in series with each lamp effectively isolates it from the DC—or quasi DC—output from the controller, while the AC passes ready through—470n offers only about one degree of resistance at 30kHz—and lights the lamps. Switching on and off the AC supply will control the lights with minimal effect on the locos.

There are just two problems—both potentially disastrous. First, building a circuit that is capable of delivering up to 1A, 6VAC at 30 kHz is not easy. If you can get away with a small output transistor like the 4F51 all well and good, but watch out for thermal runaway with larger types such as the ubiquitous 2N3055. Push-pull output at such frequencies is best accomplished using the rather more expensive (but delightful to handle) VMOS devices.

Secondly, although the AC cannot interfere with the loco motors directly it can interfere with the controller and vice-versa. Turning the controller on full can short-circuit the AC voltage so that the lamps fail and the AC generator goes into smoke. If the controller is an electronic type with a feedback loop, the AC gets into the feedback loop where it can cause all sorts of interesting phenomena, including making the trains dance jigs!

The solution is a choke in series with the controller and AC generator, but you will need a push-pull VMOS output and a split power supply.

The alternative approach is to run the coach lamps off an array of rechargeable batteries, eg., NiCads, concealed somewhere in the train and recharged from the track voltage (when available). Switching on and off is best accomplished by on-train reed switch/bistable combinations activated by temporarily placing magnets near the passing train.

The problem this time is regulating the voltage picked up from the track down to the 6V or so that can be delivered by NiCads. Normal voltage regulators generate far more heat than can be dissipated within the confines of a polystyrene coach; the best solution is a switching regulator which runs cold. Such circuits can be devised quite simply using the inexpensive 555 timer IC. The batteries are best concealed in a brake van such as the Lima full broke.

There is no simple answer to the problem of stopping the lamps from melting the train! It is best to keep the lamps as far from polystyrene structures as is practicable and under-run them by using a lower-than-rated supply. The light will be yellowish but this doesn’t really matter and it does extend the life of the lamps. Now that blue LEDs are available (if expensive) it should not be too long before white-light LEDs are with us. That should make life a lot easier for those who are engaged in modelling includes light entertainment.

Roger Amos

OPEN CHANNEL

As you read this column, a decision is being made regarding the future of British and European television systems which could bring them into broad alignment with other systems throughout the world. The benefits of a single world standard in television broadcasting are fairly obvious: programmes made in one country could be retransmitted in another without reformatting of any kind, television receivers would not be unique to individual countries but could be transported (and sold) anywhere without modification. This all sounds great in theory, but what about the practice?

Present world television systems fall into three types. The European systems are based on the PAL standard (Phase Alternation by Line), the French SECAM system (Sequentiel Couleur A Memoire), and the American NTSC system (National Television Systems Committee). The quality of picture in the three systems varies considerably: the NTSC standard, for instance, is badly affected by weather conditions, with colour hue changing as the phase of the received signal changes.

Until now, television standards have been designed to provide a system which is only as good as the customer wants. British television systems, for example, have progressed in steps—first there was a 405-line system, then the present 625-line system. In both cases, television receivers could be built using the available technology at a price the viewer was prepared to pay. Considering the complexity of television receivers, it is quite remarkable that they are as cheap as they are—due to mass-production, no doubt.

Given the steps forward in electronics, micro-miniaturisation, and communications since the three main standards were initiated, it seems sensible that any new standard should be more complex. But dare we aim for the ultimate performance possible in any television system given the present technology, or should we just create a new standard which again provides only what the viewer wants, but which we may later discard when further technological developments have taken place?

The Japanese think that the first type of standard is required. They have put a great deal of research and development into a standard high-definition television (HDTV), which uses a picture of 1125 lines and creates a viewed image of remarkable detail and clarity. This standard is so good that the Americans have been persuaded to adopt it too, which means that virtually the whole world will use HDTV—apart from Europe perhaps.

The problem with Europe is that a standard has already been developed which, although not producing quite such good picture quality as HDTV, still gives enormous quality improvements over existing PAL standards. As far as most of the broadcasting authorities in Europe are concerned, this new standard (C-MAC) is good enough for the foreseeable future of European television broadcasting. By the time C-MAC has outlived its capabilities, new technological developments will probably allow an even greater jump forward (3D television? 'smellyvision'?!) than can be obtained by

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ALF'S PUZZLE

Whenever we are puzzled by some complex bit of electronics theory, Alf is always on hand to help. 'What's a sine wave, Alf?' we ask. 'A pure frequency,' Alf ripostes without hesitation. 'What are other waves made of?' we inquire, searchingly. 'Sine waves,' Alf replies confidently.

The other day, our Alf was trying to repair a faulty square wave generator. One of the outputs was normal, the other was distorted. Imagine the surprise with which our basement brains responded when he added the two waves together and found they made a sine wave!

Alf knows that a square wave contains more harmonics than it knows what to do with, and reckons that the distorted version must have an harmonic or two knocking about as well. How on earth could all those frequencies add up to a wave with no harmonics at all?

What made matters worse was that his generator was producing 2KHz square waves and the sine wave was at 1KHz, a frequency which shouldn't have been there at all. Where on earth was it coming from?

The answer to last month's puzzle: If you connect a load between +ve and -ve in a power supply circuit, you're okay. The voltage remains fairly constant across the smoothing capacitor and gets topped up each half cycle of the mains. That's the situation with the 12V supply on its own, but try and connect a resistor from the 12V rail to an even higher voltage (say the 24V rail in last month's diagram), and it's a different story. This is exactly what Alf did by providing the 24V rail and connecting the LED and associated circuitry from it to the 12V rail. Now the bias on the 12V supply becomes reverse biased and there is nothing to stop the voltage of the original capacitor across the 12V supply rising as high as it likes. This is what happens in Alf's circuit when the switch is pressed.

HDTV now.

In effect, C-MAC pushes forward the present PAL system, utilising it more efficiently. It's not really a new standard at all, merely an adaptation of what we presently have. Nevertheless, C-MAC pictures, in an enhanced form, are of excellent quality and compare well with high-quality 35mm cinema film. The system can give a wider aspect ratio-5:3-and stereo sound.

But that's not all. Enhanced C-MAC can be transmitted easily over DBS satellite channels: HDTV as yet can't. The bandwidth of an HDTV signal (30 MHz) is greater than the total bandwidth of proposed satellite capabilities. The HDTV engineers say they will be able to compress the signal, but it is unlikely that any great savings in bandwidth will be possible - the most we can hope for is that a single satellite channel will be able to transmit a single HDTV signal whereas at present three PAL channels can be broadcast over one DBS channel.

Another technical problem arises because Europe has a mains supply system which alternates at 50 Hz while American and Japanese systems alternate at 60 Hz. Mains alternation frequency drastically affects television reception because the receivers rely on the mains frequency for various timing controls. As a result, the worldwide 'standard' isn't as standard as you may at first think.

Also, we have to consider what will happen if Europe chooses the HDTV route. The Japanese are already known to be gearing up mass-production of HDTV receivers. If all television receivers worldwide will be to that standard, they will literally be able to flood the whole market with their equipment, instantly. The European manufacturers may not have that opportunity. And what about the cost; prices as high as £500 or more have been quoted for HDTV receivers. C-MAC television receivers would be priced at around the same figures as existing PAL receivers.

So, what's the answer? Well, the decision is actually out of our hands, and is about to be made for us. By the end of October a meeting of the CCIR (the International Radio Consultative Committee) will decide which system is to be adopted in Europe. Even though HDTV is undoubtedly a better standard than C-MAC, and even though a worldwide standard is useful. I personally don't think it is the best solution for Europe.

Keith Brindley

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