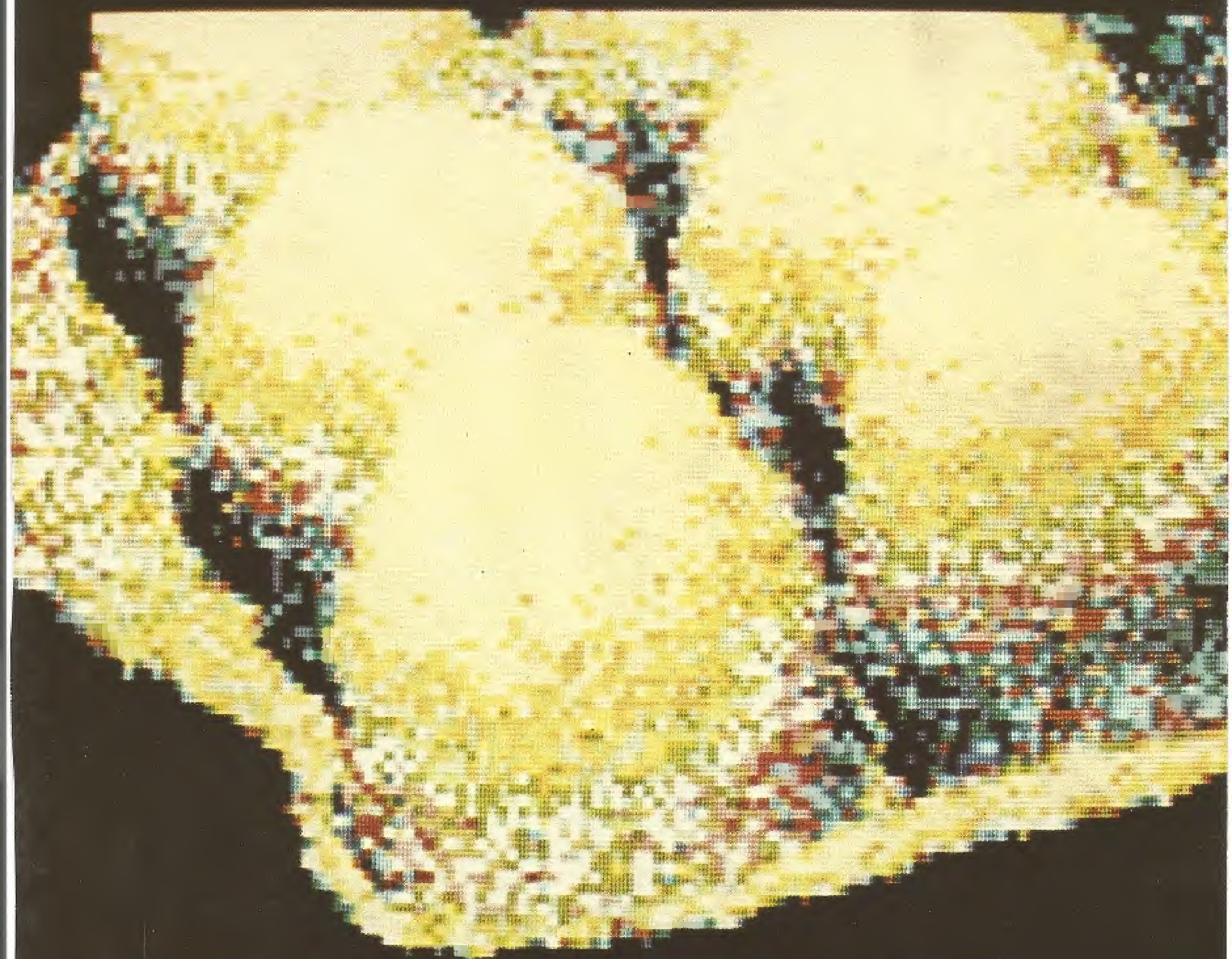
wireless world

Australia A$ 2.40
Canada C$3.25
Denmark DKr 28.25
Germany DM 6.50
Greece Dra 160.00
Holland DFl 8.00
Italy L 3100
Norway NKr 24.00
Singapore M$ 5.50
Spain Pts 240.00
USA $ 3.75

FEBRUARY 1982 70p

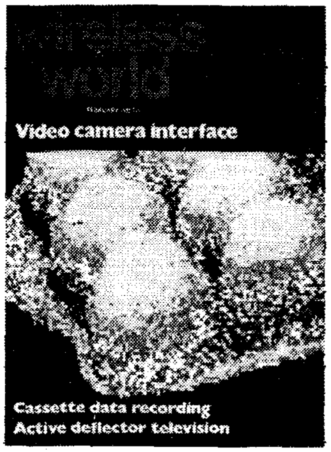
Video camera interface



Cassette data recording

Active deflector television

WW—001 FOR FURTHER DETAILS



Front cover picture is a representation of eggs in a box, obtained by P. Howard with the micro interface to tv camera described in this issue.

IN OUR NEXT ISSUE

BBC microcomputer. The first technical appraisal of the micro to be used in the BBC computer awareness programmes, which started in January. Software and hardware are both examined.

Disc storage systems. A series on the techniques used in disc storage, beginning with an article on the role of the disc drive in computing.

Nickel-cadmium cells. Charging, discharging and storage characteristics are described, and a number of charging circuits are given.

Current issue price 70p, back issues (if available) £1, at Retail and Trade Counter, Units 1 & 2, Bankside Industrial Centre, Hopton Street, London SE1. Available on microfilm; please contact editor.
By post, current issue £1.6p, back issues (if available) £1.50, order and payments to EEP General Sales Dept., Quadrant House, The Quadrant, Sutton, Surrey SM2 5AS.
Editorial & Advertising offices: Quadrant House, The Quadrant, Sutton, Surrey SM2 5AS.
Telephones: Editorial 01-661 3500. Advertising 01-661 3130.
Telegrams/Telex: 892084 BISPRS G.
Subscription rates: 1 year £12 UK and £15 outside UK.
Student rates: 1 year £8 UK and £10 outside UK.
Distribution: Quadrant House, The Quadrant, Sutton, Surrey SM2 5AS. Telephone 01-661 3500.
Subscriptions: Oakfield House, Perrymount Road, Haywards Heath, Sussex RH16 3DH. Telephone 0444 59188. Please notify a change of address.
USA mailing agents: Expediters of the Printed Word Ltd, 527 Madison Avenue, Suite 1217, New York, NY 10022. 2nd-class postage paid at New York.

 ISSN 0043 6062

wireless world

ELECTRONICS
TELEVISION
RADIO
AUDIO

FEBRUARY 1982 Vol 88 No 1553

wireless world

Acting Editor:
PHILIP DARRINGTON
01-661 3039

Technical Editor:
GEOFF SHORTER, B.Sc.
01-661 3500 X3590

Projects Editor:
MIKE SAGIN
01-661 3500 X3588

Communications Editor:
MARTIN ECCLES
01-661 3500 X3589

News Editor:
DAVID SCOBIE
01-661 3500 X3587

Design Editor:
RICHARD NEWPORT

Drawing Office Manager:
ROGER GOODMAN

Technical Illustrator:
BETTY PALMER

Advertisement Manager:
BOB NIBBS, A.C.I.I.
01-661 3130

DAVID DISLEY
01-661 3500 X3593

BARBARA MILLER
01-661 3500 X3592

Northern Sales
HARRY AIKEN
061-872 8861

Midland Sales
BASIL McGOWAN
021-356 4838

Classified Manager:
BRIAN DURRANT
01-661 3106

OPHELIA SMITH
01-661 3033

Production:
BRIAN BANNISTER
(Make-up and copy)
01-661 3500 X3561

Human engineering

While some experience of industry is a distinct benefit when one is called upon to dispense wisdom to industrialists and engineers, the lack of such experience can, evidently, also have its advantages. Prince Charles' recent speech to the Institution of Mechanical Engineers demonstrated well that the objectivity of non-involvement, when combined with perception, can be very valuable.

It has become fashionable to speak of the "British Disease" – a tired and meaningless phrase that has been used to describe almost any shortcoming in any walk of life, from the growth of bingo to an insistence on tea breaks, but the particular manifestation with which the speech was concerned was the constant theme of industrial disaster and its attraction for the news media. An obsession with failure breeds failure.

It also tends to obscure the successes of industry which, as the Prince pointed out, are considerable, and not all from the larger companies. Indeed, it sometimes seems that the larger a company, the less the motivation that can be expected from its staff. There is no lack of will to work and to obtain a better way of life for oneself and one's family – the 7.5% proportion of the 'black economy' testifies to that – but the buffering effect of working in an amorphous organization such as BL or GEC removes much of the incentive to put in more than a contractual amount of effort. The work of one man has only an imperceptible influence on the company's performance: or, at least, that is the inevitable, subjective impression.

There is no lack of successful Japanese companies to prove that the Japanese workforce equates its own fortunes with those of the company to a much greater extent than seems to be the case here, but it is at least possible that differences in national character call for different approaches.

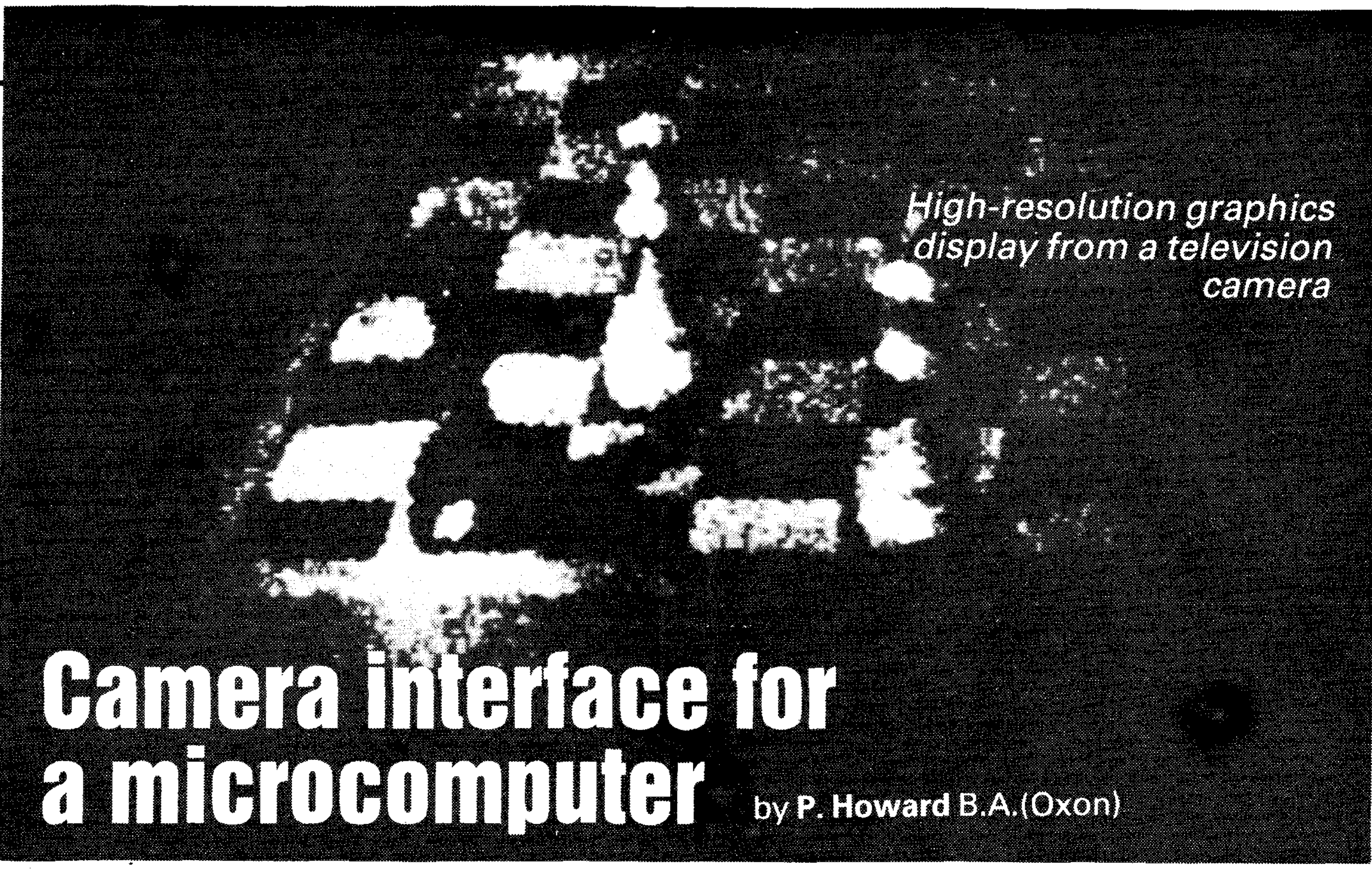
Job satisfaction is a well-worn phrase, but a good deal more than lip-service to the idea is needed if its benefits are to be gathered. Even to hint at a reduction in the number of people performing tedious, unskilled, mind-atrophying tasks by 'closing the loop' in automatic production machinery would be ill-timed, to say the least: a re-deployment of the same number of people in an imaginative way might, however, be practicable and acceptable.

In small companies, many of which are enterprising, enthusiastic and successful, there are few conveyor belts and endlessly repeated, apparently meaningless operations. Single workers or groups are able to produce and see more complete – in some plants entire – pieces of equipment: the finished product is theirs and they are responsible for it. It is hard for an employee on a production line to feel at one with the company that employs him, but considerably easier, and perhaps even more to be desired, to feel pride of achievement in the product itself.

Admittedly, this is only raising the level of repetition, but the contribution to an immediately recognizable product must be more gratifying than the insertion of a few components or even the final testing. People are thinking animals and should not be expected to function as maintenance-free machinery. Over thirty percent of one's waking life from 16 to 65 is spent at work – there must be more to it than a Pavlovian response to the stimulus of a workpiece moving past one's nose.

Tom Ivall

Keen-eyed readers will have noticed that Tom Ivall's name is no longer on our 'masthead'. He has decided to leave *Wireless World* after many years – eight of them in the Editor's chair – to pursue a freelance writing career. We in the editorial team will miss his friendly guidance and persuasive leadership, but hope still to see his work in our pages from time to time. We wish him well in his new career.



High-resolution graphics display from a television camera

Camera interface for a microcomputer

by P. Howard B.A.(Oxon)

Many microcomputers have the facility for displaying high resolution graphics. High resolution in this context means about 300 × 200 points. One problem is that the software required for even a fairly simple diagram is quite extensive and may represent a considerable fraction of the entire program. The interface to be described was designed to enable a picture to be acquired, stored and displayed by the computer in high-resolution graphics, and was required to be relatively inexpensive and reasonably versatile. It presents the computer with picture information from a number of tv frames, taking about five seconds to build up the complete picture. The camera and subject must therefore be stationary for this time.

A tèlevision frame of the CCIR standard consists of two fields each of 312.5 lines, although the first and last few lines of each field do not contain any picture information. A line is 64 microseconds long and picture information is transmitted for about 40 microseconds of this time. The signal from the video camera used was +1 volt peak combined with synchronizing pulses of -1 volt.

The video camera interface (v.c.i.) divides each line into 256 sections, which may be numbered 0 to 255. When it is initialized, it digitizes the voltage of point 0 for each line of the field following the initialization signal. During the next field, point 1 of each line is digitized and the process continues until every point has been converted to a digital number. The interface does not distinguish between the two fields of a tv frame, as this is not necessary for the resolution required, though it could be accomplished if required.

Each point or section is converted to an eight-bit number, the value of which is proportional to the brightness of that section. The time between successive conversions is therefore 64 microseconds. This allows enough time for the computer to accept each number and store it in memory. The time to digitize the complete picture is therefore 256 field periods or 5.12 seconds.

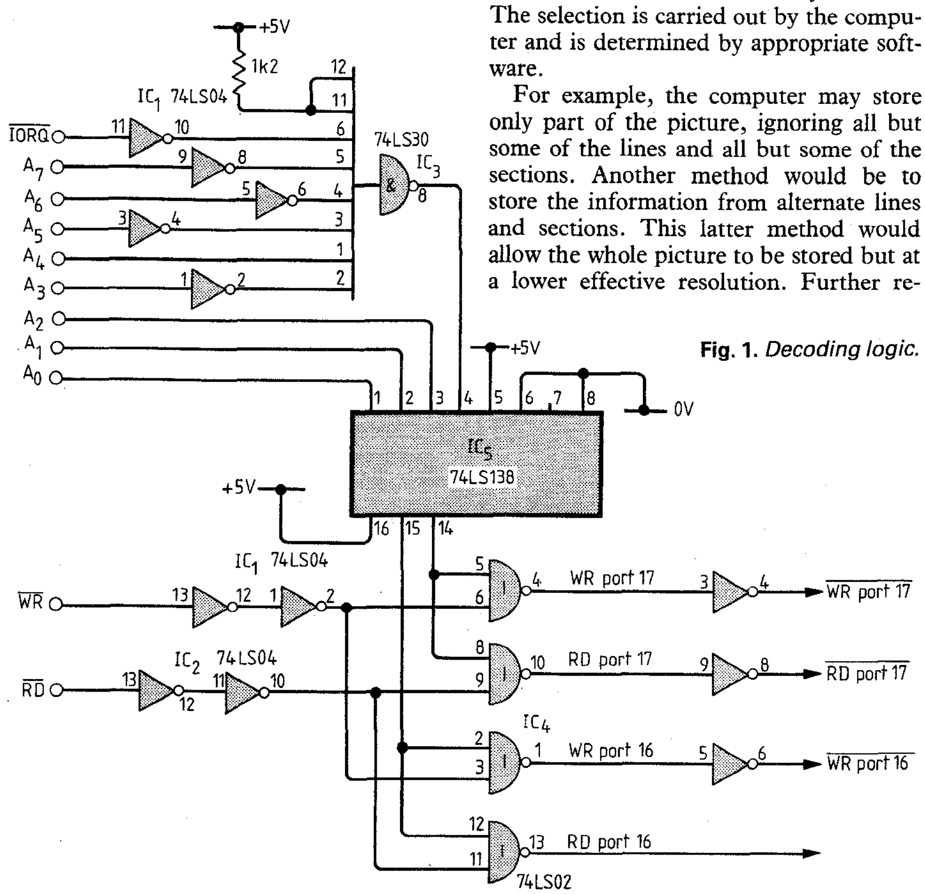


Fig. 1. *Decoding logic.*

Storage and display

Simple arithmetic shows that to store all the information presented by the interface for a picture of, say, 256 lines, each of 256 points would require 65536 bytes. This is more than the entire memory of many small computers. It may be necessary for the computer to store only part of the information with which it is presented. There are various ways in which the information to be stored may be chosen and a combination of methods may be used. The selection is carried out by the computer and is determined by appropriate software.

For example, the computer may store only part of the picture, ignoring all but some of the lines and all but some of the sections. Another method would be to store the information from alternate lines and sections. This latter method would allow the whole picture to be stored but at a lower effective resolution. Further reduction in memory requirements could be achieved by storing only part of the digital number. If the four most significant bits are selected, there will be only 16 different possible contrast levels instead of the 256 that storing eight bits would allow.

A constraint on the complexity of the selection software is that the computer must decide whether a number presented to it by the v.c.i. is to be stored, select the appropriate number of bits to store, decide where it is to be stored and then finally store it, all before the next number is presented. The software must therefore allow the computer to accomplish this within 64 microseconds. A careful check must be made on the execution times of the machine-code instructions. The example programs were written for a Z-80A microprocessor running at 4MHz, arranged to insert an extra WAIT state into every memory read operation.

Once the picture has been stored in memory, it may be displayed or it may be processed in a variety of ways. We shall not discuss picture processing in detail as we have little experience in this field. (Another reason for building the v.c.i. was to learn about the subject.) The software for displaying the picture will depend on the facilities available, but should be fairly straightforward to write. The prototype was interfaced to a Research Machines 380Z and some examples of display software for the high resolution graphics board of this computer are included.

Circuit description

The v.c.i. circuitry can be divided into sections:

- The decoding logic, which is completely straightforward and is the section most likely to require modification to suit different systems. The prototype was decoded to Z-80 input and output ports, as these are easier to decode than memory addresses.
- The converter section comprises a sample-and-hold circuit and the analogue-to-digital converter itself. It also contains synchronization pulse separators which indicate the start of tv lines and fields.
- The timing circuitry sends a signal to the a-to-d converter to start the conversion of a particular section of a line. The v.c.i. runs continuously, but a signal from the computer can restart the timing circuitry at the beginning of the picture.

Decoding logic

It was decided to connect the v.c.i. to the computer via input and output ports as this simplifies decoding. For those unfamiliar with the Z-80, it should be mentioned that these transfer data via the data bus when the IORQ signal is active. The port address appears on the lower eight address lines and the RD and WR signals determine whether data are to be read into, or sent from the c.p.u. If the v.c.i. is decoded to memory locations, all 16 address lines are significant[1].

When the lower eight address lines assume any of the values from 00010000 to 00010111 and IORQ is low, the output of IC_3 goes low and enables the 3 to 8 line decoder IC_5. Depending on the state of the least significant three address lines, one of the outputs of IC_5 will go low. These signals are gated with either RD or WR, depending on whether an input port or an output port is required. Any inputs which are connected to more than one place on the v.c.i. are buffered so that the interface puts a maximum of one l.s.t.t.l. load on any computer line. The circuitry actually decodes eight ports, though not all of these are used and are therefore available for use with other peripheral circuits if required.

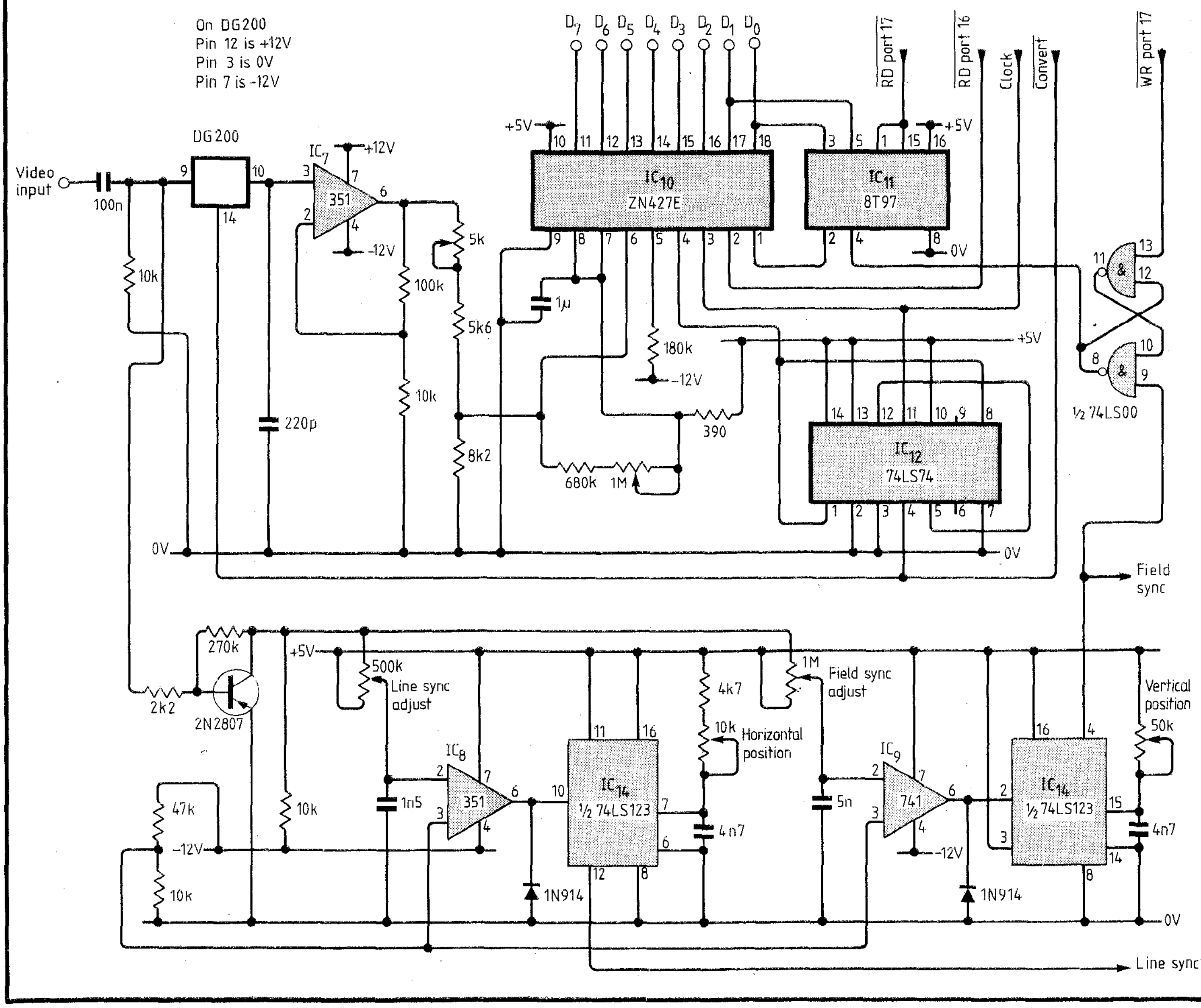


Fig. 2. *Analogue-to-digital converter and sync. separators.*

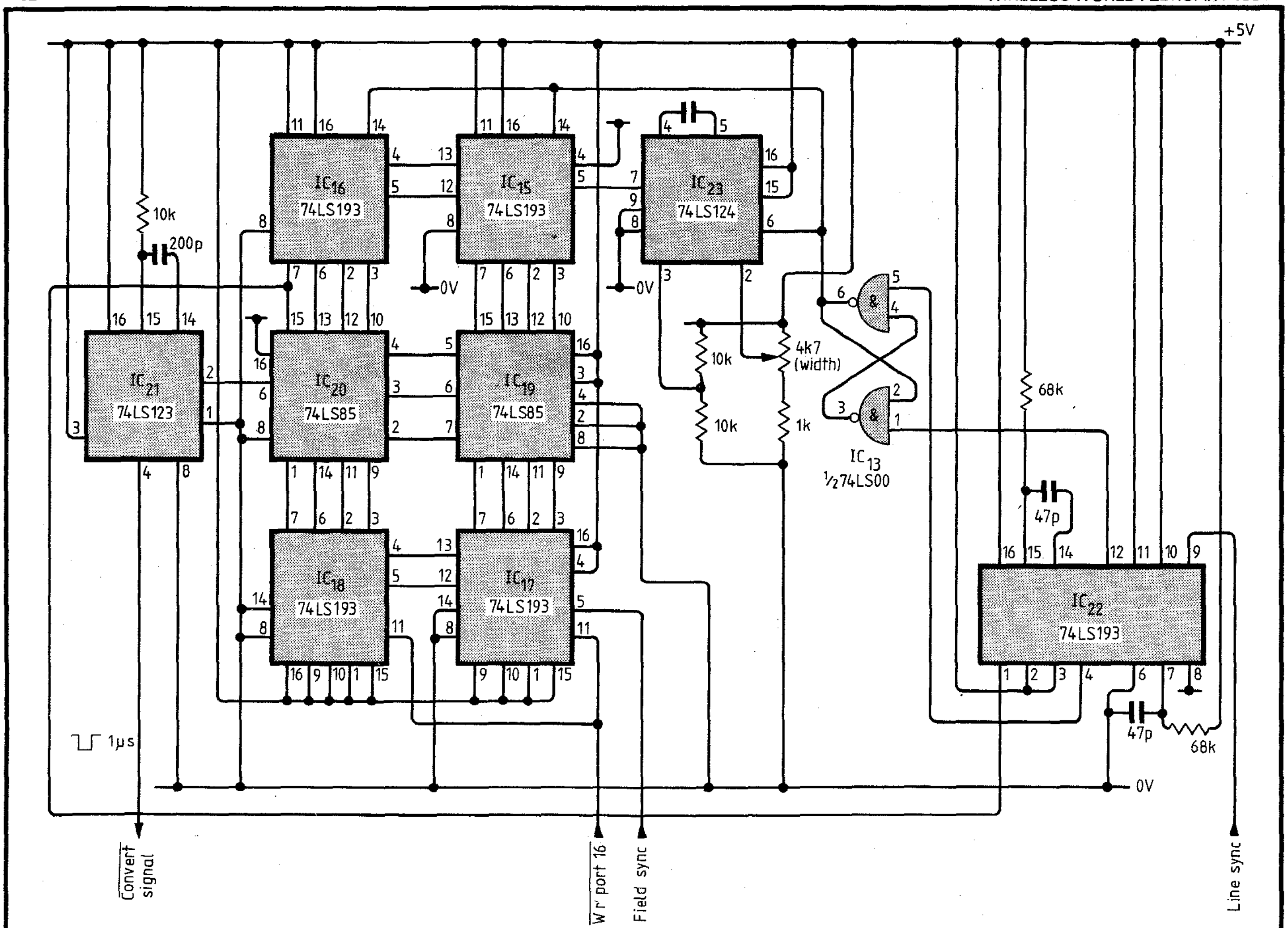


Fig.3. *Timing circuitry.*

Analogue-to-digital converter

The combined video signal from the camera is fed to the input of a sample and hold circuit, consisting of a c.m.o.s. switch, capacitor and f.e.t. input operational amplifier. The size of the capacitor chosen allows it to charge to the value of the input voltage in the 1 microsecond for which the c.m.o.s. switch is turned on. It must maintain this charge for the 18 microseconds conversion time of the a.d.c. The output of the sample and hold operational amplifier is connected through a level setting potentiometer to the a.d.c. which is connected as suggested in reference 2. The ZN427E analogue to digital converter used has tri-state outputs which are connected to the data bus and enabled when input port 16 is read.

To separate the synchronizing pulses, the video signal is first amplified by the transistor Tr_1 and then passed to two operational amplifiers used as voltage comparators. These are unable to change state until the capacitors on their inputs have charged through the input resistors. The values of these resistors and capacitors are chosen so that IC_8 will change state during line pulses, but the capacitor on IC_9 will not charge rapidly enough to change state except during the longer field pulses. The outputs of these comparators are arranged to trigger monostables with controllable delays (IC_{14}).

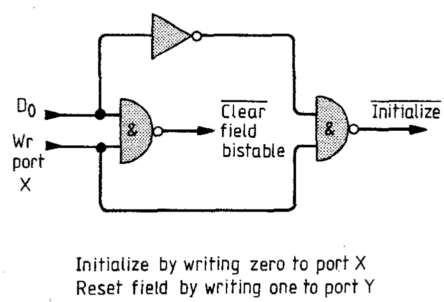


Fig. 4. *Modification to enable both control signals to operate from one output port.*

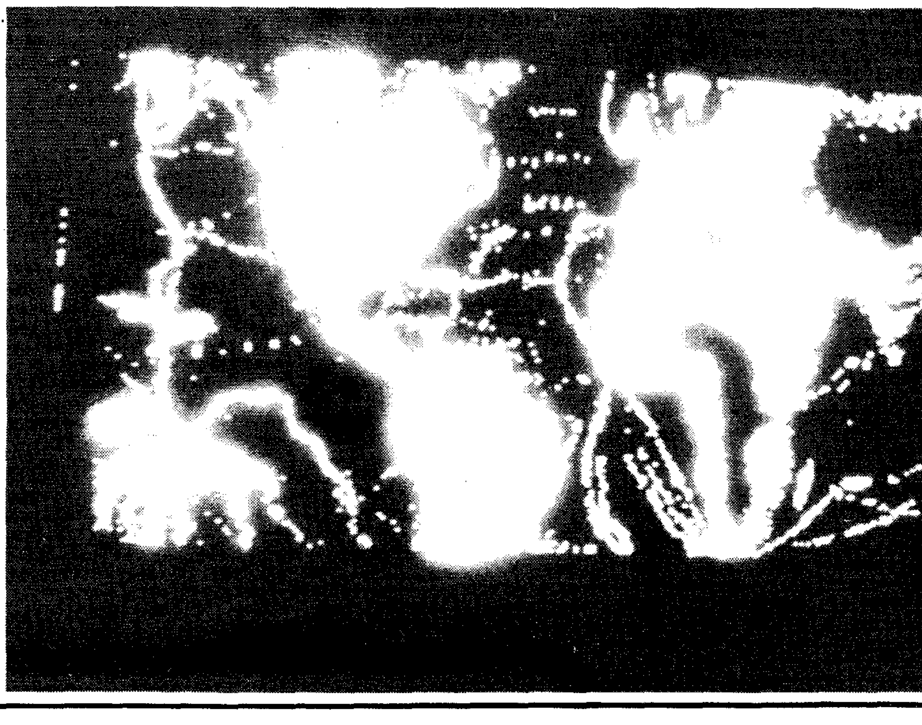


Fig. 5. *A map of the world with trade routes, digitized from an atlas – high resolution.*

Timing circuitry

The number of the section to be digitized is stored in the binary counter formed by IC_{17} and IC_{18}. This is incremented by each field pulse. This counter therefore stores the number of field pulses since the picture was started. As one section of each line is converted every field, this is also the number of the section to be converted during the current field. The contents of this counter are compared with those of another counter formed by IC_{15} and IC_{16}. This comparison is performed by IC_{19} and IC_{20} which are magnitude comparators, which have two sets of inputs and generate an output when both sets have the same logical values. The second counter is drive from a voltage-controlled oscillator, IC_{23},

Fig. 6. *A medium resolution picture showing the increased number of shades. The software used has compressed this picture vertically.*

Fig. 7. *Assembler listing of high resolution acquisition routine.*

```
                      ;Camera Board Software
                      ;Version 1.2
                      ;256 dots by 128 lines

             *FORMFEED OFF

49DE =                FIRST = 18910   ;Start of Picture Store
69DE =                LAST  = 27102   ;End of Picture Store
0020 =                LINES = 32      ;Number of Lines/4
0069 =                HILAST= LAST/256;High part of LAST address
00DE =                LOLAST= LAST-HILAST*256;Low part

0000 21DE49           LD HL, FIRST ;Set HL to start of picture store
0003 3600     NXCLR:  LD (HL),0    ;Clear memory location
0005 23               INC HL       ;Go to next location
0006 7C               LD A, H      ;Check to see if we
0007 FE69             CP HILAST    ;are at end of picture
0009 20F8             JR NZ, NXCLR ;store...
000B 7D               LD A, L
000C FEDE             CP LOLAST
000E 20F3             JR NZ, NXCLR ;...and go back if not.
0010 21DE49           LD HL, FIRST ;Set HL to start again
0013 0E00             LD C, 0      ;Reg C contains number
                                   ;of lines digitized/4

0015 D310             OUT (16),A   ;Initialize VCI
0017 D311     FRMCLR: OUT (17),A   ;Reset field bistable
0019 DB11     WTSOF:  IN A, (17)   ;Check to see if
001B E602             AND 2        ;field bistable is set
001D 28FA             JR Z, WTSOF  ;and wait until it is
001F 0614             LD B,20      ;These lines merely
0021 DB11     PAUSE:  IN A, (17)   ;provide a delay
0023 E601             AND 1        ;of 20 lines so that
0025 20FA             JR NZ, PAUSE ;a more central part
0027 DB11     PAUSE2: IN A, (17)   ;of the picture is stored.
0029 E601             AND 1        ;This is done by waiting
002B 28FA             JR Z, PAUSE2;for the conversion complete
002D 05               DEC B        ;line to go low and high
002E 78               LD A,B       ;twenty times.
002F FE00             CP 0


0031 20EE             JR NZ, PAUSE
0033 1802             JR WT1
0035 18E0     ISLE:   JR FRMCLR ;Allows relative jump from end of routine.
0037 DB11     WT1:    IN A, (17) ;Accept control signals
0039 E601             AND 1        ;Select  End of Conversion
003B 20FA             JR NZ, WT1 ;Wait for it to go low
003D DB11     WTN1:   IN A, (17)
003F E601             AND 1
0041 28FA             JR Z, WTN1 ;...and then high again
0043 DB10             IN A, (16) ;Accept byte from ADC
0045 E6C0             AND 0C0H     ;Select top two bits
0047 CB3F             SRL A        ;Put them in lowest
0049 CB3F             SRL A        ;positions
004B CB3F             SRL A
004D CB3F             SRL A
004F CB3F             SRL A
0051 CB3F             SRL A
0053 B6               OR (HL)      ;OR with rest of currentt picture byte
0054 77               LD (HL), A ;Put new result back into memory.

0055 DB11      WT2:      IN A, (17) ;Now do exactly the same
0057 E601                AND 1      ;with three more sections
0059 20FA                JR NZ, WT2
005B DB11      WTN2:     IN A, (17)
005D E601                AND 1
005F 28FA                JR Z, WTN2
0061 DB10                IN A, (16)
0063 E6C0                AND 0C0H
0065 CB3F                SRL A         ;...except that second section
0067 CB3F                SRL A         ;is put in next lowestt
0069 CB3F                SRL A         ;two bits...
006B CB3F                SRL A
006D B6                  OR (HL)
006E 77                  LD (HL),A
006F DB11      WT3:      IN A, (17)
0071 E601                AND 1
0073 20FA                JR NZ, WT3
0075 DB11      WTN3:     IN A, (17)
0077 E601                AND 1
0079 28FA                JR Z, WTN3
007B DB10                IN A, (16)
007D E6C0                AND 0C0H
007F CB3F                SRL A          ;...third section in
0081 CB3F                SRL A          ;second highest pair...
0083 B6                  OR (HL)
0084 77                  LD (HL), A
0085 DB11      WT4:      IN A, (17)
0087 E601                AND 1
0089 20FA                JR NZ, WT4
008B DB11      WTN4:     IN A, (17)
008D E601                AND 1
008F 28FA                JR Z, WTN4
0091 DB10                IN A, (16)
0093 E6C0                AND 0C0H
0095 B6                  OR (HL)        ;...and fourth in top
0096 77                  LD (HL), A   ;two bits.
                                        ;This memory location
                                        ;is now filled, so we must
0097 23                  INC HL         ;start on the next one.
0098 0C                  INC C          ;Reg C is also incremented
0099 79                  LD A, C        ;and used to see if
009A FE20                CP LINES       ;enough lines have been digitized.
009C 2099                JR NZ, WT1     ;If not then go back.
009E 7C                  LD A, H        ;See if all picture memory
009F FE69                CP HILAST      ;has been used...
00A1 2004                JR NZ, RESETL
00A3 7D                  LD A, L
00A4 FEDE                CP LOLAST
00A6 C8                  RET Z          ;If so then EXIT
00A7 0E00      RESETL:   LD C,0         ;Reset reg C
00A9 188A                JR ISLE        ;Go and wait for next field


49DE FIRST   0017 FRMCLR  0069 HILAST  0035 ISLE    69DE LAST
0020 LINES   00DE LOLAST  0003 NXCLR   0021 PAUSE   0027 PAUSE2
00A7 RESETL  0037 WT1     0055 WT2     006F WT3     0085 WT4
003D WTN1    005B WTN2    0075 WTN3    008B WTN4    0019 WTSOF

No errors
```

which is started after a delay by the line sync. pulse and stopped when the counter has cycled once. The period of the v.c.o. is approximately 170ns, allowing the counter to cycle in about 40 microseconds. This corresponds to the duration of the picture information in a video line. Adjusting the speed of the v.c.o. alters the fraction of the video line digitized and can therefore be used as a 'width' control. The position on the video line may be adjusted by altering the monostable delay (IC_{14}) before the counter is started.

The magnitude comparator produces an output when the two counters are equal. It will therefore produce a 170ns pulse once per line which will be progressively delayed as the field counter is advanced. This is too short to allow the sample-and-hold switch to turn on, so monostable IC_{21} lengthens it to about 1 microsecond. This signal is also used to start the analogue to digital conversion.

The field counter may be initialized by a signal from the computer on output port 16. This resets it to 255 (its maximum count) ensuring that it will start at zero after the next field pulse. The computer also needs to be supplied with information about the number of fields that have elapsed since the v.c.i. was initialized. This is achieved by a bistable circuit which is reset by a signal from output port 17 and set by the next field pulse. The computer can read the state of the bistable by examining the second least significant bit (bit 1) of input port 17.

Construction

The v.c.i. was constructed on a Vero prototyping board which was connected via an edge connector to the 380Z power supply and bus. Few special precautions were taken over the layout and construction. Leads from the v.c.o. and the magnitude comparators were kept short, as were the video input leads. Power supply decoupling was extravagant. About twenty 0.01μF capacitors were used, distributed around the board to ensure that each integrated circuit was decoupled by a capacitor within 3 cm of its supply pins.

Hardware modifications

There is a number of ways in which the circuitry could be modified. Perhaps the most useful would be to connect the data and control signals to a single input port. This would mean reducing the number of available bits from the a.d.c. to a maximum of six, as two bits are needed for the field and conversion complete signals. This would perhaps not be an important limitation: it would enable the interface to be operated via the 380Z user port, for example. It would also be necessary to arrange for the v.c.i. reset and field bistable reset to be operable from a single output port. The decoding circuitry would then be largely eliminated. A suggestion for how this could be achieved is shown in diagram 4.

To decode the interface at different ports, it is only necessary to change the arrangement of inverters on the address lines. An inverter is included on each line that is low when the v.c.i. is accessed.

Obtaining a better resolution

The field counter is advanced by one after every field, so each frame will provide the information for two sections. If the field pulses are divided by two before they reach the counter, the resolution will be increased by a factor of two.

To increase the number of sections per line, it is merely necessary to increase the length of each of the field and section counters and the magnitude comparator that connects them. The v.c.o. speed will also need to be increased. Before embarking on such a modification, it would be advisable to ensure that the counters and comparators could operate at the increased speeds.

Fig. 8. *Assembler listing of medium resolution acquisition routine.*

```
                    ;Camera Board Software
                    ;Version 2.2
                    ;160 dots by 96 lines

               *FORMFEED OFF

49DE =              FIRST = 18910    ;Start of Picture Store
67DE =              LAST  = 26590    ;End of Picture Store
0030 =              LINES =    48    ;Number of Lines/2
0067 =              HILAST= LAST/256 ;High part of LAST address
00DE =              LOLAST= LAST-HILAST*256 ;Low part

0000 21DE49         LD HL, FIRST  ;Set HL to start of picture sto
0003 3600   NXCLR:  LD (HL),0     ;Clear memory location
0005 23             INC HL        ;Go on to next location
0006 7C             LD A, H       ;Check to see if we
0007 FE67           CP HILAST     ;are at end of picture
0009 20F8           JR NZ, NXCLR  ;store...
000B 7D             LD A, L
000C FEDE           CP LOLAST
000E 20F3           JR NZ, NXCLR  ;...and go back if not.
0010 21DE49         LD HL, FIRST  ;Set HL to start again
0013 0E00           LD C, 0       ;Reg C contains number
                                  ;of lines digitized/2
0015 D310           OUT (16),A    ;Initialize VCI
0017 D311   FRMCLR: OUT (17),A    ;Reset field bistable
0019 DB11   WTSOF:  IN A, (17)    ;Check to see if
001B E602           AND 2         ;field bistable is set
001D 28FA           JR Z, WTSOF   ;and wait until it is.
001F 0614           LD B,20       ;These lines merely
0021 DB11   PAUSE:  IN A, (17)    ;provide a delay
0023 E601           AND 1         ;of 20 lines so that
0025 20FA           JR NZ, PAUSE  ;a more central part
0027 DB11   PAUSE2: IN A, (17)    ;of the picture is stored.
0029 E601           AND 1         ;This is done by waiting
002B 28FA           JR Z, PAUSE2  ;for the conversion complete
002D 05             DEC B         ;line to go low and high
002E 78             LD A.B        ;twenty times.
002F FE00           CP 0
0031 20EE           JR NZ, PAUSE
0033 DB11   WT1:    IN A, (17)    ;Accept control signals
0035 E601           AND 1         ;Select 'End of Conversion'
0037 20FA           JR NZ, WT1    ;Wait for it to go low
0039 DB11   WTN1:   IN A, (17)
003B E601           AND 1
003D 28FA           JR Z, WTN1    ;...and then high again
003F DB10           IN A, (16)    ;Accept byte from ADC
0041 E6F0           AND 0F0H      ;Select top four bits
0043 CB3F           SRL A         ;Put them in lowest
0045 CB3F           SRL A         ;positions
0047 CB3F           SRL A
0049 CB3F           SRL A
004B B6             OR (HL)       ;OR with rest of current picture byte
004C 77             LD (HL), A    ;Put new result back into memory.
004D DB11   WT2:    IN A, (17)    ;Now skip a line so only
004F E601           AND 1         ;alternate lines are digitized
0051 20FA           JR NZ, WT2
0053 DB11   WTN2:   IN A, (17)
0055 E601           AND 1
0057 28FA           JR Z, WTN2
0059 DB11   WT3:    IN A, (17)    ;Repeat the process
005B E601           AND 1         ;with another section
005D 20FA           JR NZ, WT3
005F DB11   WTN3:   IN A, (17)
0061 E601           AND 1
0063 28FA           JR Z, WTN3
0065 DB10           IN A, (16)
0067 E6F0           AND 0F0H      ;...except that it is put
0069 B6             OR (HL)       ;in the upper four bits.
006A 77             LD (HL), A
006B DB11   WT4:    IN A, (17)    ;Skip another line
006D E601           AND 1
006F 20FA           JR NZ, WT4
0071 DB11   WTN4:   IN A, (17)
0073 E601           AND 1
0075 28FA           JR Z, WTN4
0077 23             INC HL        ;Go to next memory location
0078 0C             INC C         ;Reg C is incremented
0079 79             LD A, C       ;and used to see if
007A FE30           CP LINES      ;enough lines have been digitized
007C 20B5           JR NZ, WT1    ;If not then go back.
007E 7C             LD A, H       ;See if all picture memory
007F FE67           CP HILAST     ;has been used...
0081 2004           JR NZ, RESETL
0083 7D             LD A, L
0084 FEDE           CP LOLAST
0086 C8             RET Z         ;if so then EXIT
0087 0E00   RESETL: LD C,0        ;Reset Reg C
0089 188C           JR FRMCLR     ;Go and wait for next field

49DE FIRST   0017 FRMCLR  0067 HILAST  67DE LAST    0030 LINES
00DE LOLAST  0003 NXCLR   0021 PAUSE   0027 PAUSE2  0087 RESETL
0033 WT1     004D WT2     0059 WT3     006B WT4     0039 WTN1
0053 WTN2    005F WTN3    0071 WTN4    0019 WTSOF

No errors
```

Fig. 9. *Assembler listing of high resolution display routine.*

```
                          ;Camera Board Software

                          ;This routine draws the picture
                          ;from the information stored
                          ;by the CAMERAH routine.

                          ;High Resolution Version

49DE =                    FIRST = 18910
FB00 =                    PORTO = 0FB00H
FB01 =                    PORT1 = 0FB01H
0000 =                    FRAME = 0
0001 =                    LINE  = 1

0000 2100FB               LD HL, PORTO
0003 3603                 LD (HL), 3  ; Set HR mode
0005 3607                 LD (HL),7   ; Open Video
0007 21DE49               LD HL, FIRST ;Start of video store
000A 0E00                 LD C,0 ; Stores X posn/4
000C 0600                 LD B,0 ; Stores Y posn

000E 1600        NXY2:    LD D,0 ; Info for HRG
0010 EF2E                 CALR TWOBIT
0012 B2                   OR D ; Put two bits in D
0013 57                   LD D,A
0014 EF3D                 CALR NEXTX
0016 EF28                 CALR TWOBIT
0018 CB27                 SLA A
001A CB27                 SLA A
001C B2                   OR D
001D 57                   LD D,A
001E EF33                 CALR NEXTX
0020 EF1E                 CALR TWOBIT
0022 CB27                 SLA A
0024 CB27                 SLA A
0026 CB27                 SLA A
0028 CB27                 SLA A
002A B2                   OR D
002B 57                   LD D,A
002C EF25                 CALR NEXTX
002E EF10                 CALR TWOBIT
0030 CB27                 SLA A
0032 CB27                 SLA A
0034 CB27                 SLA A
0036 CB27                 SLA A
0038 CB27                 SLA A
003A CB27                 SLA A
003C B2                   OR D
003D 57                   LD D,A ; Now we have a byteful
003E 181E                 JR PLOT ; in register D
0040 C5          TWOBIT:  PUSH BC
0041 78                   LD A,B ; Take appropriate two
0042 E603                 AND 3  ; and put them in low
0044 47                   LD B,A ; bits of reg A
0045 7E                   LD A,(HL)
0046 2807                 JR Z,ENDBIT
0048 CB3F        SHIFT:   SRL A
004A CB3F                 SRL A
004C 05                   DEC B
004D 20F9                 JR NZ,SHIFT
004F E603        ENDBIT:  AND 3
0051 C1                   POP BC
0052 C9                   RET
0053 7D          NEXTX:   LD A,L ; Adds 32 to HL
0054 C620                 ADD 32
0056 6F                   LD L,A
0057 7C                   LD A,H
0058 CE00                 ADC 0
005A 67                   LD H,A
005B C9                   RET
005C 18B0        NEXTY:   JR NXY2
005E E5          PLOT:    PUSH HL
005F 58                   LD E,B
0060 CB3B                 SRL E
0062 CB3B                 SRL E ; Take top half of Y
0064 CB3B                 SRL E ; address and send to
0066 CB3B                 SRL E ; port 1
0068 2101FB               LD HL,PORT1
006B 73                   LD (HL).E
006C 3E0F                 LD A,15
006E A0                   AND B
006F 69                   LD L,C
0070 CB25                 SLA L ; Take low 4 bits of X
0072 CB25                 SLA L ; and high 4 bits of Y
0074 CB25                 SLA L
0076 CB25                 SLA L
0078 B5                   OR L
0079 6F                   LD L,A ; to form low video address
007A 61                   LD H,C
007B CB3C                 SRL H
007D CB3C                 SRL H ; High four bits of X
007F CB3C                 SRL H
0081 CB3C                 SRL H
0083 3EF0                 LD A,0F0H ; and Base address
0085 B4                   OR H
0086 67                   LD H,A ; to form high part
0087 7A                   LD A,D
0088 54                   LD D,H
0089 5D                   LD E,L
008A 2100FB               LD HL,PORTO
008D CB46                 BIT FRAME,(HL)
008F 2808                 JR Z,W3
0091 CB4E        W1:      BIT LINE,(HL)
0093 28FC                 JR Z,W1
0095 CB4E        W2:      BIT LINE,(HL)
0097 20FC                 JR NZ,W2
0099 00          W3:      NOP
009A 12                   LD (DE),A
009B E1                   POP HL
009C 7D                   LD A,L ; Get back original HL value
009D D660                 SUB 96
009F 6F                   LD L,A
00A0 7C                   LD A,H
00A1 DE00                 SBC 0
00A3 67                   LD H,A
00A4 04                   INC B ; Increase Y by one
00A5 78                   LD A,B ; If divisible by 4 then
00A6 E603                 AND 3
00A8 20B2                 JR NZ,NEXTY
00AA 23                   INC HL ; increase HL address
00AB 78                   LD A,B ; Is this end of vertical line?
00AC FE80                 CP 128
00AE 20AC                 JR NZ,NEXTY
00B0 0C                   INC C ; If so then increase X
00B1 116000               LD DE,96
00B4 19                   ADD HL,DE
00B5 0600                 LD B,0
00B7 79                   LD A,C
00B8 FE40                 CP 64 ; Have we finished???
00BA 20A0                 JR NZ,NEXTY
00BC 2100FB               LD HL,PORTO
00BF 3603                 LD (HL),3
00C1 C9                   RET

004F ENDBIT  49DE FIRST   0000 FRAME  0001 LINE   0053 NEXTX
005C NEXTY   000E NXY2    005E PLOT   FB00 PORTO  FB01 PORT1
0048 SHIFT   0040 TWOBIT  0091 W1     0095 W2     0099 W3
```

Fig. 10. *Assembler listing of medium resolution display routine.*

```
                          ;Camera Board Software

                          ;This routine draws the picture
                          ;from the information stored
                          ;by the CAMERAM routine.

                          ;Medium Resolution Version

49DE =                    FIRST = 18910
FB00 =                    PORTO = 0FB00H
FB01 =                    PORT1 = 0FB01H
0000 =                    FRAME = 0
0001 =                    LINE  = 1

0000 2100FB               LD HL, PORTO
0003 36A3                 LD (HL), 0A3H  ; Set MR mode
0005 36A7                 LD (HL),0A7H   ; Open Video
0007 21DE49               LD HL, FIRST ;Start of video store
000A 0E00                 LD C,0 ; Stores X posn/4
000C 0600                 LD B,0 ; Stores Y posn

000E 1600        NXY2:    LD D,0 ; Info for HRG
0010 EF0E                 CALR IVBIT
0012 B2                   OR D ; Put four bits in D
0013 57                   LD D,A
0014 EF24                 CALR NEXTX
0016 EF08                 CALR IVBIT
0018 CB27                 SLA A
001A CB27                 SLA A
001C B2                   OR D
001D 57                   LD D,A ; Now we have a byteful
001E 1825                 JR PLOT ; in register D
0020 7E          IVBIT:   LD A,(HL)
0021 CB40                 BIT 0,B
0023 2808                 JR Z,ENDBIT
0025 CB3F                 SRL A
0027 CB3F                 SRL A
```

```
0029 CB3F             SRL A
002B CB3F             SRL A
002D E60F    ENDBIT:  AND 15
002F 5F               LD E,A ; Shift high bits
0030 E60C             AND 12 ; two places
0032 CB27             SLA A
0034 CB27             SLA A  ; to the left
0036 B3               OR E
0037 E633             AND 33H
0039 C9               RET
003A 7D      NEXTX:   LD A,L ; Adds 48 to HL
003B C630             ADD 48
003D 6F               LD L,A
003E 7C               LD A,H
003F CE00             ADC 0
0041 67               LD H,A
0042 C9               RET
0043 18C9    NEXTY:   JR NXY2
0045 E5      PLOT:    PUSH HL
0046 58               LD E,B
0047 CB3B             SRL E ; Take top half of Y
0049 CB3B             SRL E ; address and send to
004B CB3B             SRL E ; port 1
004D 2101FB           LD HL,PORT1
0050 73               LD (HL),E
0051 3E07             LD A,7
0053 A0               AND B
0054 CB27             SLA A
0056 69               LD L,C
0057 CB25             SLA L ; Take low 4 bits of X
0059 CB25             SLA L ; and low 3 bits of Y
005B CB25             SLA L
005D CB25             SLA L
005F B5               OR L
0060 6F               LD L,A ; to form low video address
0061 61               LD H,C
0062 CB3C             SRL H
0064 CB3C             SRL H ; High four bits of X
0066 CB3C             SRL H
0068 CB3C             SRL H
006A 3EF0             LD A,0F0H ; and Base address
006C B4               OR H
006D 67               LD H,A ; to form high part
006E 7A               LD A,D
006F 54               LD D,H
0070 5D               LD E,L
0071 2100FB           LD HL,PORT0
0074 CB46             BIT FRAME,(HL)
0076 2808             JR Z,W3
0078 CB4E    W1:      BIT LINE,(HL)
007A 28FC             JR Z,W1
007C CB4E    W2:      BIT LINE,(HL)
007E 20FC             JR NZ,W2
0080 00      W3:      NOP
0081 12               LD (DE),A
0082 E1               POP HL
0083 7D               LD A,L ; Get back original HL value
0084 D630             SUB 48
0086 6F               LD L,A
0087 7C               LD A,H
0088 DE00             SBC 0
008A 67               LD H,A
008B 04               INC B ; Increase Y by one
008C CB40             BIT 0,B
008E 20B3             JR NZ,NEXTY
0090 23               INC HL ; increase HL address
0091 78               LD A,B ; Is this end of vertical line?
0092 FE60             CP 96
0094 20AD             JR NZ,NEXTY
0096 0C               INC C ; If so then increase X
0097 113000           LD DE,48
009A 19               ADD HL,DE
009B 0600             LD B,0
009D 79               LD A,C
009E FE50             CP 80 ; Have we finished???
00A0 20A1             JR NZ,NEXTY
00A2 2100FB           LD HL,PORT0
00A5 36A3             LD (HL),0A3H
00A7 C9               RET

002D ENDBIT  49DE FIRST  0000 FRAME  0020 IVBIT  0001 LINE
003A NEXTX   0043 NEXTY  000E NXY2   0045 PLOT   FB00 PORT0
FB01 PORT1   0078 W1     007C W2     0080 W3

No errors
```

Software description

In very broad terms, the behaviour of the picture acquisition software is as follows:

1. Initialize the v.c.i.
2. Wait for the start of field signal.
3. Wait for the conversion complete signal.
4. Accept the picture information and store it.
5. Repeat steps 3-5 until enough lines have been accepted.
6. Repeat steps 2-6 until all the sections have been accepted.

In practice, some of these steps are somewhat complicated to implement, especially if memory is restricted and only part of the information is to be stored. When writing the software it is important to remember that steps 4 and 5 are time critical and must not take more than 64 microseconds to execute. The exact program required will depend on so many factors that it would be impossible to discuss all the possibilities. We will content ourselves with some examples which may need to be modified or completely rewritten for different implementations of the v.c.i. The programs are documented and are self-explanatory. They are written in relocatable code, in other words they do not need to occupy any particular area of memory. However they use the relative call instruction CALR, which is not a Z-80 instruction but is interpreted by a routine within the 380Z monitor program. If the system does not possess an equivalent facility, CALL instructions will have to be substituted. This will necessitate re-assembly of the program for a specific area of memory.

Display software

The description of the software required for displaying a picture will be confined to that used for the RML high resolution graphics board. The steps involved in displaying a picture are:

1. Initialize the High Resolution Graphics board by setting it up for the appropriate resolution mode and clearing the graphics memory.
2. Set up the 'Colour Lookup Table' so that the brightness of the displayed spot is proportional to the number representing that spot.
3. 'Open' the graphics memory so that it may be written to during video line and frame blanking periods.
4. Collect the picture information for each byte of graphics memory and store it in the appropriate memory location. Each byte will contain the brightness value of either two (in 'medium resolution' mode) or four (high resolution) picture elements.

RML have already written routines for stages 1 and 2 as extensions to BASIC and it was therefore decided to use these rather than duplicating their effort.

The whole display software can be written as a BASIC program, but this is extremely slow, taking several minutes to display a single picture. An assembly language routine was therefore written which can be inserted into memory and called from a controlling BASIC program. Two versions of this routine are required; one each for high resolution and medium resolution pictures. The reader is referred to the RML High Resolution Graphics manual for an explanation of h.r.g. addressing.

References

1. Z80 Microcomputer Devices Technical Manual MK3880 Central Processing Unit, Mostek Corporation, 1977
2. Data Sheet R/4052 A/D Converter I.C., RS Components Ltd, 1980
3. The TTL Data Book for Design Engineers (Second Edition), Texas Instruments Incorporated, 1976
4. High Resolution Graphics Reference Manual, Research Machines Limited, 1980

Radar explores the ionosphere

New incoherent scatter radar system in northern Scandinavia

by I. Berkovitch, Ph.D.

"I am told" said the King of Sweden "that if I press this red button, something dramatic will happen". And, sure enough, in response to the signal the 32-metre diameter dish of the EISCAT u.h.f. radio telescope at Kiruna in North Sweden obediently turned and tilted to pick up echoes of signals transmitted from Tromso in Norway. The occasion was the inauguration of the *E*uropean *I*ncoherent *Scat*ter facilities simultaneously at three sites linked by radio – Kiruna, Tromso and Sodankyla in Finland.

EISCAT is an advanced radar system designed to study the upper atmosphere at high latitudes. It is jointly supported by Finland, France, Germany, Norway, Sweden and the UK. But what is "incoherent scatter"? At the lower frequencies of radar systems operating in the MHz range, nearly all of the wave energy directed to the ionosphere is returned to Earth. This is known as coherent total reflection. But at higher frequencies, using exceptionally strong radar signals, very weak echoes are obtained from ionospheric electrons that can be picked up with a large radio telescope and amplified with a very sensitive receiver. Most of the energy of the transmitted waves escapes into space but a minute fraction returns. The principles – and difference in behaviour – are shown in Fig 1. The method is called "incoherent scatter radar" (ISR).

The physicists working on the project emphasise the magnitude of the problem of detecting these signals by comparing it with obtaining a radar signal from a small coin at a distance of several hundred kilometres.

There are already five ISR laboratories active in other parts of the world. But this new £12 million group of installations is claimed to be a second generation facility advancing the technique, opening up new fields of upper atmosphere research and located in a region of special interest. There are two independent radar systems. A v.h.f. system has both a transmitter and receiver only in Tromso. This will scan in the magnetic meridian and up to 20° either side to the east or west. A u.h.f. system (at 933MHz) has a transmitter at Tromso and receivers at Tromso, Kiruna and Sodankyla. All three of the u.h.f. radio telescopes can look at the same volume of the upper atmosphere at the same time. The sampling height is of course determined by the place where the transmitter and receiver beams intersect. By measuring the scattered signal in three different directions (see Fig. 2), EISCAT can make a three-dimensional measurement of the velocity of ionised material in the upper atmosphere.

Quantities that can be measured include electron density and electron temperature, ion temperature, ion composition, plasma bulk velocity and the magnetic field. Measuring these quantities makes it possible to study such phenomena as exchange of mass and energy between the ionosphere and the magnetosphere, the field aligned plasma flow and the atmospheric electric currents. And ISR data will be combined with other observations from satellites, rockets and other sources in such studies as the relationship between the magnetosphere and the ionosphere.

The auroral or polar cap regions are considered especially suitable for such work since they are the boundary regions between the magnetic field of the Earth and the magnetic field of interplanetary space. More detailed measurements of the aurorae will now be made than were ever before possible.

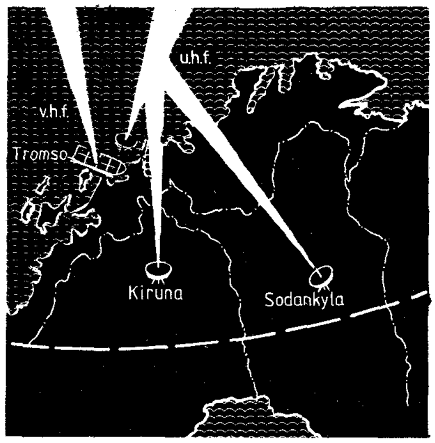
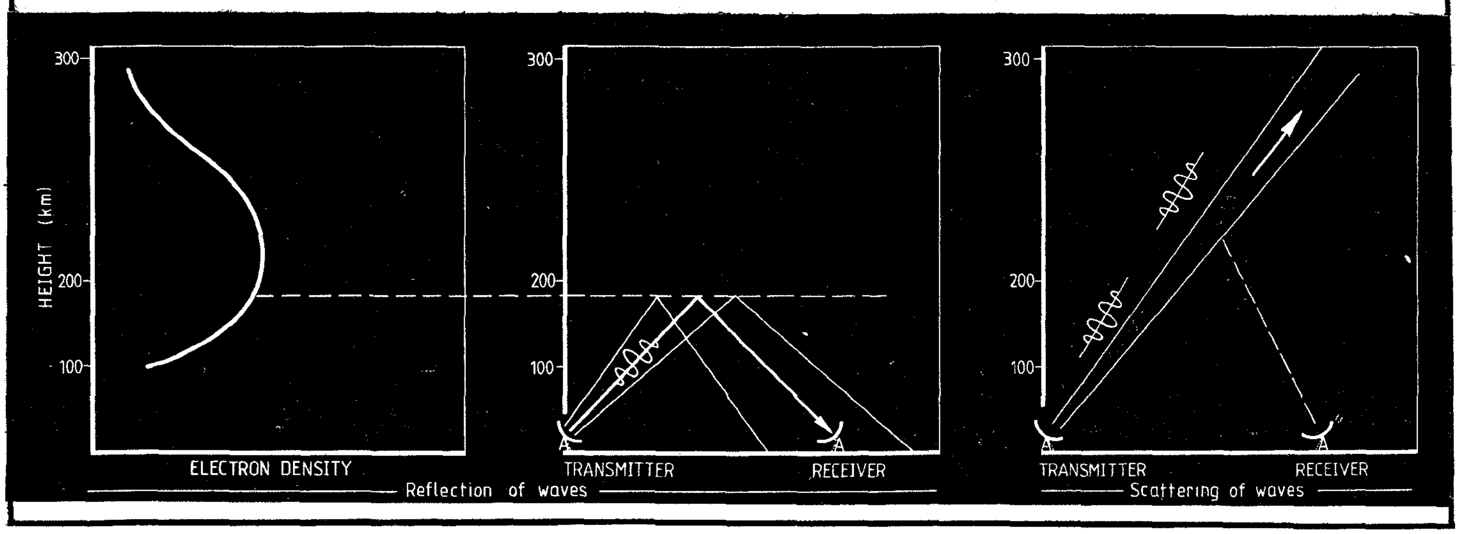


***Fig. 2.** Arrangement of radar systems in the EISCAT scheme. In the v.h.f. system the transmitter and receiver are in Tromso. The u.h.f. system, however, has a transmitter in Tromso and receivers at Tromso, Kiruna and Sodankyla. Both transmitters have a repertoire of pulse modulation waveforms to match the differing requirements for space, time and frequency resolution in various parts of the ionosphere.*

In an opening speech, Professor Sir Granville Benyon, chairman of EISCAT Council, pointed out that they had set up arrangements between scientific organisations, not governments, that would provide a flexible framework for scientists from the six countries to work together.

Operating for 48 hours a week, the facility will devote about half its time to common programmes adopted by a scientific advisory committee. The first of these will be measuring and mapping the temperature of the aurorae as a function of latitude by scanning procedures. The remaining time will be allotted to special programmes proposed by individual countries with relative time allowed in the proportion of their contributions to overall costs.

***Fig. 1.** Illustrating the difference between reflection of electromagnetic waves from the ionosphere (left) and the scattering of waves (right). Also shown is a graph of the density of free electrons resulting from ionization of the Earth's upper atmosphere by solar radiation.*



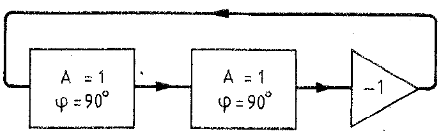
Phase-shifting oscillator

Low distortion design improves on Wien bridge

by Roger Rosens, Ing.

The use of a thermistor to stabilize an oscillator can lead to third harmonic distortion, especially at low frequencies. The circuit described here includes a simple network which virtually eliminates the third harmonic component. The result is an oscillator with a very flat frequency characteristic and very low distortion (typically 0.0005%)

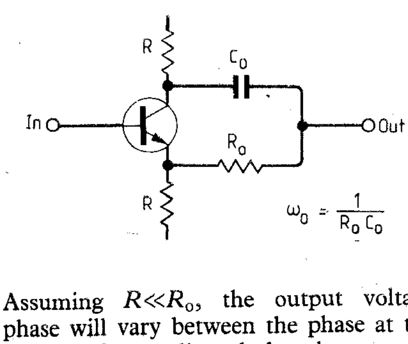
When a simple variable-frequency generator is required to give a low distortion sine wave, the commonly used circuit is the Wien-bridge oscillator. In its elementary form, this circuit requires only one op-amp as the active device. Using the kind of audio op-amp now available it is, however, possible to build other attractive circuits with only a little more complexity. Compared with the Wien, the phase-shifting oscillator presented here shows a flatter frequency characteristic and a significant reduction of the third-harmonic distortion caused by the stabilizer thermistor at low frequencies. The circuit is based on two 90° phase-shifting networks, followed by an inverter stage, giving a total loop phase shift of 360°.



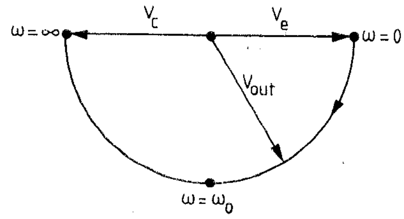
Operation of the phase-shifting network

The phase-shifting network is in fact a first order all-pass filter, the transfer function of which is defined by $F(p)=p-\omega_0/p+\omega_0$, where ω_0 is the corner frequency. This function has a constant magnitude equal to 1 at all frequencies, while the phase shift varies from 0° to 180°. The phase shift attains 90° at the corner frequency ω_0; this will thus be the oscillation frequency.

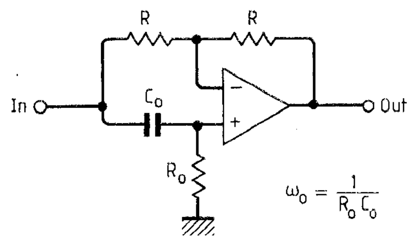
The first-order all-pass function can be realized with the following circuit:



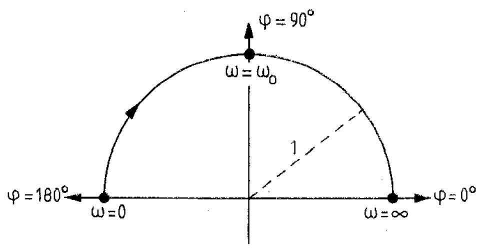
Assuming $R \ll R_0$, the output voltage phase will vary between the phase at the emitter (for $\omega=0$) and the phase at the collector (for $\omega=\infty$), which gives a phase variation of 180°.



An improved version of the all-pass circuit replaces the transistor with an op. amp.



The transfer function of this circuit is $F(p)=p-\omega_0/p+\omega_0$. The magnitude of thus always 1 and the phase angle is given by $\phi=180°-2 \arctan \omega/\omega_0$. The polar plot is:



The oscillation frequency can be adjusted by varying R_0 or C_0. Since there are two all-pass networks used in the oscillator circuit, a two-ganged element will be required to adjust the frequency.

The use of all-pass networks in an oscillator circuit has two important advantages:
– Stable amplification factor (equal to 1), regardless of the equality between the ω_0 of the all-pass circuits.
– Consequently, there is no need for close matching of the ganged element. The oscillator will have a very flat frequency characteristic while it is possible to use a low cost ganged potentiometer for the frequency adjustment.

Basic circuit diagram

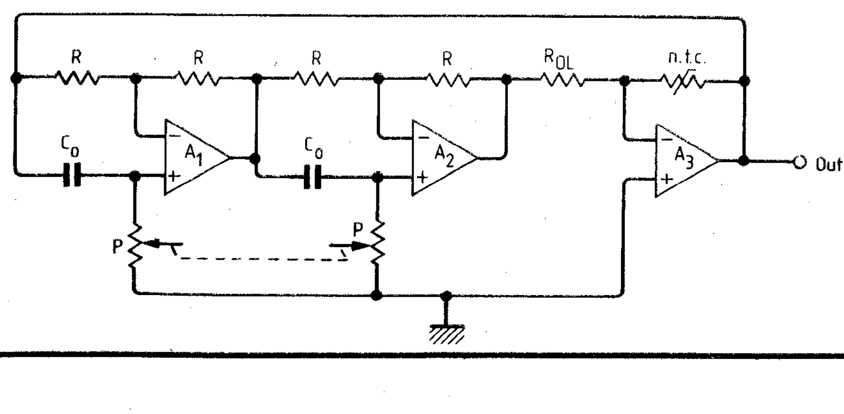
The complete oscillator circuit is quite simple (Fig. 1) The oscillation frequency is adjusted with P. The output level is stabilized with a thermistor (n.t.c.). Theoretically, the operating point is fixed at $R_{ntc}=R_{ol}$. If A_1 and A_2 are in the same package, their input bias currents will be about equal so that the offset voltages, caused by the voltage drops over P, will cancel each other at the output of A_2. Hence, the dc voltage on the thermistor will be negligible. This is important because this dc voltage causes second harmonic distortion, especially at low frequencies. For the same reason, the maximum resistance value of P must be limited to ≈100kΩ.

The circuit has two further interesting features:
–it can deliver three different sine waves of equal amplitude with relative phases of 0°, 90° and 180°.
– the frequency-adjustment potentiometers are both connected to ground. Compared with the Wien-bridge oscillator, this makes it easy to convert the circuit into a programmable oscillator. This is done by replacing the potentiometers by fixed resistors which may be switched by f.e.ts. The f.e.ts would all have their sources connected to circuit ground, which would make their gate drive very simple.

Distortion considerations

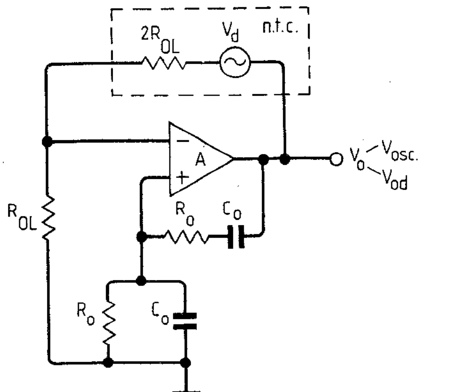
Two kinds of distortion are produced in the circuit:
– distortion generated by the active components.
– distortion generated by the amplitude stabilizing mechanism concerning the distortion in the op amps, a figure of

Fig. 1. *The basic phase-shifting oscillator circuit.*

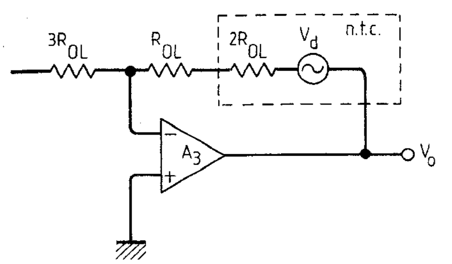


<0.01% can be obtained easily by choosing a quality audio op. amp.

The distortion introduced by the thermister is more difficult to reduce because a compromise has to be made between low distortion, fast settling time and good temperature stability. The n.t.c. distortion varies inversely with the settling time and the frequency while it is almost proportional to the temperature rise of the n.t.c. (see appendix 1). As is known, the relative temperature coefficient of the oscillator voltage is equal to $-1/2\Delta T$. Since a certain amount of thermistor distortion must be tolerated, it is important to reduce its effect on the output voltage as much as possible. This can be done by using an oscillator circuit with good frequency selectivity. One can calculate (see appendix 1) that the distortion generated in the n.t.c. consists mostly of third harmonic. We can now compare the output distortions between the Wien-bridge and the phase-shift oscillators. Let v_o be the oscillator output voltage, composed of: the fundamental, v_{osc}, and the distortion, v_{od}; and let v_d be the (3rd harmonic) distortion voltage generated by the n.t.c. v_d can also be defined as $d_3 v_{ntc}$ in which d_3 = distortion figure of the n.t.c. and v_{ntc} = oscillator voltage on the n.t.c. With the Wien bridge the circuit is:



For the phase-shifting oscillator, we can re-arrange the circuit so that v_{ntc} and v_{osc} are the same as on the Wien bridge circuit and this output stage results:



The output distortion can be determined in two ways: – by direct calculation of the transfer function v_{od}/v_d. Putting $=1/R_oC_o$, this results in:

$$\frac{v_{od}}{v_d}=\frac{-p^2+3p\omega_o+\omega_o^2}{p^2+\omega_o^2}$$

for the Wien bridge circuit, and

$$\frac{v_{od}}{v_d}=\frac{p^2+2p\omega_o+\omega_o^2}{2(p^2+\omega_o^2)}$$

for the phase-shift circuit. We can use the relation derived by Thomas Philips (*Electronic Engineering*, April 1981). If $F(p)$ is the transfer function of the frequency selective network, the distortion transfer function of the nth harmonic is given by:

$$\frac{v_{od(n)}}{v_{d(n)}}=\frac{F(j\omega_o)}{F(nj\omega_o)-F(j\omega_o)}$$

For the Wien bridge, $F(p)=p\omega_o/(p^2+3p\omega_o+\omega_o^2)$ and $F(j\omega_o)=1/3$,

$$\text{Thus } \frac{v_{od}}{v_d}=\frac{1/3}{F(nj\omega_o)-1/3}$$

For the phase-shift network, $F(p)=(p-\omega_o)(p+\omega_o)$ and $F(j\omega_o)=-1$.

$$\text{Thus } \frac{v_{od}}{v_d}=\frac{-1}{F(nj\omega_o)+1}$$

Of course the two methods give the same results. For the 3rd harmonics we find:

$$\frac{v_{od}}{v_d}=\frac{\sqrt{145}}{8}\approx 1.5$$

for the Wien-bridge circuit and, since

$$v_d=d_3v_{ntc}=d_3\times 2/3 v_{osc},$$

$$\frac{v_{od}}{v_{osc}}\approx 1.5\times 2/3 d_3\approx d_3$$

$$\frac{v_{od}}{v_d}=\frac{\sqrt{100}}{16}\approx 0.6$$

for the phase-shift network and thus

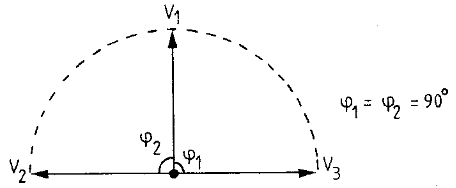
$$\frac{v_{od}}{v_{osc}}\approx 0.6\times 2/3\approx 0.4d_s$$

Conclusion: For similar operating conditions, the output distortion of the phase-shifting circuit is two and a half times less than that of the Wien-bridge. Since the phase-shifting circuit has no amplitude selectivity, this result is at first sight surprising. In fact, good harmonic suppression is a consequence of the circuit's "phase selectivity".

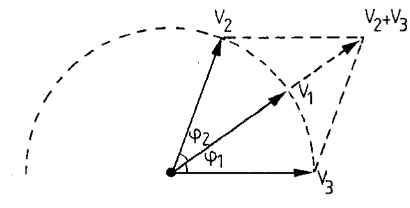
Additional circuit, to further reduce distortion

Choosing a practical compromise of the different circuit characteristics, the output distortion for the described circuit is 0.1% at 20Hz, decreasing to <0.01% above 100Hz. Further attempts to reduce these figures resulted in an additional circuit that virtually eliminates the third harmonic distortion generated by the ntc Let v_1,v_2 and v_3 be the voltages at the outputs of A_1 and A_2 and A_3.. The relationship of these voltage is given by the following diagrams:

for the fundamental:



– for the 3rd harmonic



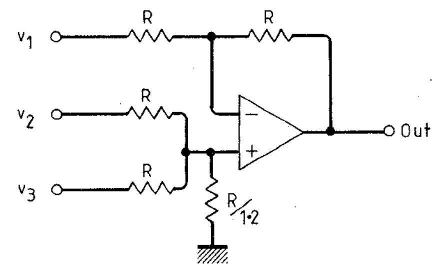
We can easily find that:

$$\phi_1=\phi_2=180°-2\arctan 3\approx 37°$$

$$v_2+v_3=1.6v_1, \text{ or}$$

$$v_1-\frac{v_2+v_3}{1.6}=0$$

This means that the third harmonic distortion can be eleiminated with a simple adder circuit. A suitable design is:



With regard to the fundamental, this circuit has no influence: v_2 and v_3 cancel each other, so that $v_{out}=(-)v_1$.

In practice, due to component tolerances, the distortion cannot be completely eliminated. The main source of error comes from the difference of ϕ_1 and ϕ_2, derived from the matching difference between the all-pass networks. When using 1% components and a ganging tolerance of 1dB for the dual potentiometer, the reduction of the 3rd harmonic is about 20 times. Since the distortion decreases with the frequency, the 1dB ganging tolerance is only required around the maximum resistance setting of the potentiometers.

Practical circuit and measured characteristics

The basic circuit has been optimized for the audio range 20Hz-20Hz. The selected op-amp is the NE5532, a dual circuit with low noise, low distortion and a still fair voltage gain of 2200 at 10kHz. (Some tests were also made with the TL072 but the results were not as good). With the addition of the distortion cancelling circuit, the distortion figure which was 0.1% at 20Hz falls to <0.005% over the whole frequency range.

The lower distortion limit is about 0.0002% (at 1000Hz). the final circuit diagram is shown in Fig. 2.The power supply for the circuit is ±12V to ±15V. The resistors are 1% metal film from the E96 series. Approximate values of the E24 series will also do the job. The range selecting capacitors should be preferably 1% polystyrene types. (For the 820 nF, selected polycarbonate capacitors were used with good result). The choice of the n.t.c. type was determined by the available op-amp current, the allowed distortion and the reqired output level. A 68kΩ, 20mW from Philips (code number 2322 634 32683) was selected. The operating point of the thermistor lies at about 3.4V and 910Ω which gives a dissipation of about 12mW and a minimum output voltage of 5V (typically 5.4V). The 100pF capacitor in the output stage compensates for a small lift in the frequency characteristic at the high frequency end of the range.

surements were carried out with fixed 1% resistors.

– The large bandwidth of the NE5532 requires some precaution: the wiring must be very short and capacitive loads should be avoided. During the tests, the connection of the oscilloscope through a coax cable caused h.f. oscillations. The remedy is to load the circuit only via a series resistor ≥100Ω. Preferably, a 600Ω (R_{out}) will be chosen in order to obtain a standard generator impedance.

Appendix 1: distortion generated in the n.t.c.

The resistance of an n.t.c. resistor is given by the exponential law: $R = Ae^{B/T}$. When subjected to an ac voltage, the n.t.c. temperature will vary cyclically and hence its resistance will be modulated. This means that the instantaneous voltage/current relationship will be non-linear; in other words, some distortion has been generated. The amount of distortion can be calculated starting from the following basic

When we apply to the n.t.c. a sine wave voltage with an r.m.s. value, v_o: $v_{ntc}=\sqrt{2}v_o\cos\omega t$. We define the corresponding operating point by P_o, R_o and T_o which are related by:

$$P_o=\frac{v_o^2}{R_o}=\delta\Delta T=\delta(T_o-T_{amb}) \qquad (4)$$

By using (4), (3) can also be written as:

$$P\mathrm{d}t=H\mathrm{d}T=P_o\mathrm{d}t$$

$$\text{or } \frac{(\sqrt{2}v_o\cos\omega t)^2}{R}\mathrm{d}t=H\mathrm{d}T+P_o\mathrm{d}t \qquad (5)$$

(1) can be transformed into $lnR=lnA+B/T$ and after differentiation:

$$\frac{\mathrm{d}R}{R}=\frac{B}{T^2}\mathrm{d}T \qquad (6)$$

For small variations of the n.t.c. temperature, R and T may be approximated in the equations (5) and (6) by R_o and T_o; this gives:

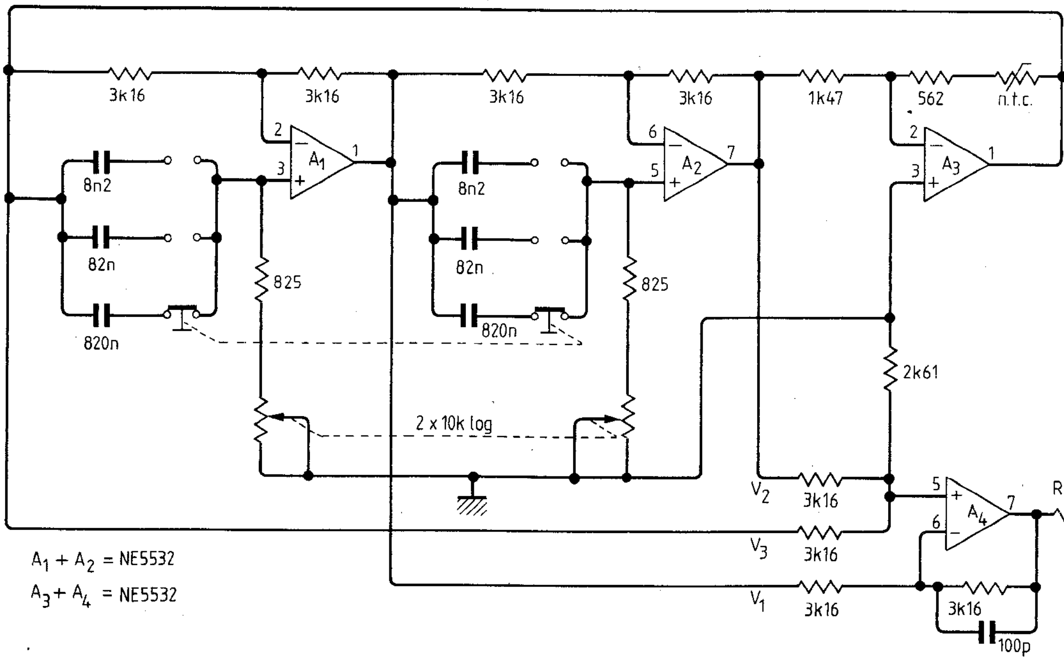


Fig. 2. *The complete circuit for an audio oscillator.*

The circuit characteristics, as measured on the breadboard model, are:

– level flatness (20Hz-20KHz): 0.04dB

– temperature dependence: −0.03dB/K

– harmonic distortion (R_{load}≥1kΩ) : <0.004% (typically 0.0005%)

The signal characteristics at the outputs of op-amps A_1, A_2 and A_3 are:

– level flatness : 0.03dB at the output of A_3

– harmonic distortion : 0.06dB at the output of $A_1$and A_2

: 01.% at 20Hz decreasing to 0.01% above 1000Hz

Remarks

– During the development of the circuit, consumer grade potentiometers were used. At some resistance setting, these potentiometers introduced a lot of noise and signal distortion due to the poor contact resistance. Therefore, the distortion mea-

expressions:

$$R=Ae^{B/T} \qquad (1)$$
$$P=v^2 \qquad (2)$$
$$P\mathrm{d}t=H\mathrm{d}T+\delta\Delta\mathrm{d}t \qquad (3)$$

where R=resistance of the n.t.c.

A,B=(nearly) constants depending on the n.t.c. type

T=absolute n.t.c. temperature (in K)

P=power dissipated by the n.t.c.

v=voltage across the n.t.c.

H=heat capacity of the n.t.c. ceramic material (in J/K)

δ=dissipation factor of the n.t.c. (in W/K)

ΔT=temperature increase of the n.t.c. caused by the power dissipated in it.

$$H\mathrm{d}T=\frac{2v_o^2}{R_o}\cos^2\omega t\,\mathrm{d}t-P_o\mathrm{d}t$$

$$=P_o(2\cos^2\omega t-1)\mathrm{d}t$$

$$H\mathrm{d}T=P_o\cos^2\omega t\mathrm{d}t \qquad (7)$$

$$\text{and } \frac{\mathrm{d}R}{R_o}=-\frac{B_2}{T_o}\mathrm{d}T \qquad (8)$$

Eliminating $\mathrm{d}T$ between (7) and (8) results in:

$$\frac{\mathrm{d}R}{R_o}=-\frac{BP_o}{HT_o}\cos2\omega t\,\mathrm{d}t$$

and, after integration:

$$\frac{R-R_o}{R_o}=-\frac{BP_o}{2\omega HT_o^2}\sin2\omega t$$

$$\text{or } R=R_o 1-\frac{BP_o}{2\omega HT_o^2}\sin 2\omega t$$

The current is given by:

$$i=\frac{v_{ntc}}{R}$$

$$=\frac{\sqrt{2}v_o\cos\omega t}{R_o\left(1-\frac{BP_o}{2\omega HT_o^2}\sin 2\omega t\right)}$$

which is nearly equal to

$$\frac{\sqrt{2}v_o}{R_o}\cos\omega t\left(1+\frac{BP_o}{2\omega HT_o^2}\sin 2\omega t\right)$$

$$=\frac{\sqrt{2}v_o}{R_o}\left(\cos\omega t+\frac{BP_o\sin\omega t}{2\omega HT_o^2}\ \frac{\sin 3\omega t}{2}\right)$$

The current is thus composed of the fundamental and of a third harmonic. This would be the same if a voltage, composed of a fundamental and a 3rd harmonic, were applied to a fixed resistor R_o. For the fundamental component, the term is negligible with regard to the term $\cos\omega t$; so, the third harmonic distortion can be approximated by

$$d_3=\frac{BP_o}{4\omega HT_o^2}$$

$$\text{or } d_3=\frac{B\delta\Delta T}{4\omega H(T_{amb}+\Delta T)^2} \qquad (9)$$

Table 1. *Distortion measurement results.*

Frequency (Hz)		110	263	520	1092	2636	5224	9564
Harmonic components (dB)	H2	−104	−112	−117	−122	−119	−113	−108
	H3	−117	−124	−121	−117	−116	−115	−114
	H4	−121	−124	−124	−123	−125	−123	−123
	H5	−119	−120	−121	−120	−118	−118	−119
	H7	−125	−128	−130	−126	−128	−126	—

Measurements were made using an HP3580A spectrum analyser preceded by a passive notch filter, giving a measuring limit of −130dB.

This function is zero for $\Delta T=0$ and $\Delta T=\infty$. Its maximum is reached for $\Delta T=T_{amb}$ (in K). For small values of ΔT, the distortion is almost proportion to ΔT. The expression $B\delta/H$ can be seen as a measure for the distortion proper to a certain type. For the used n.t.c., $B=3900$k, $\delta=0.11$mW/K and $H=0.5$mJ/K.

Using (1), expression (9) can be transformed to:

$$d_3=\frac{1}{4\omega\tau}\left(-\frac{T_{amb}}{B}\ln\frac{R_{amb}}{R_o}\right)\ln\frac{R_{amb}}{R_o} \qquad (10)$$

Where $\tau=H/\delta$ thermal time constant of the n.t.c.

R_{amb}=n.t.c. resistance at the ambient temperature

R_o=n.t.c. resistance at the operating point.

In the particular case when R_o is only slightly less than R_{amb} we have

$$\ln\frac{R_{amb}}{R_o}=\ln\left(1+\frac{R_{amb}-R_o}{R_o}\right)$$

$$\approx\frac{R_{amb}-R_o}{R_o}$$

and (10) becomes $d_3\approx\frac{1}{4\omega\tau}\ \frac{R_{amb}-R_o}{R_o}$,

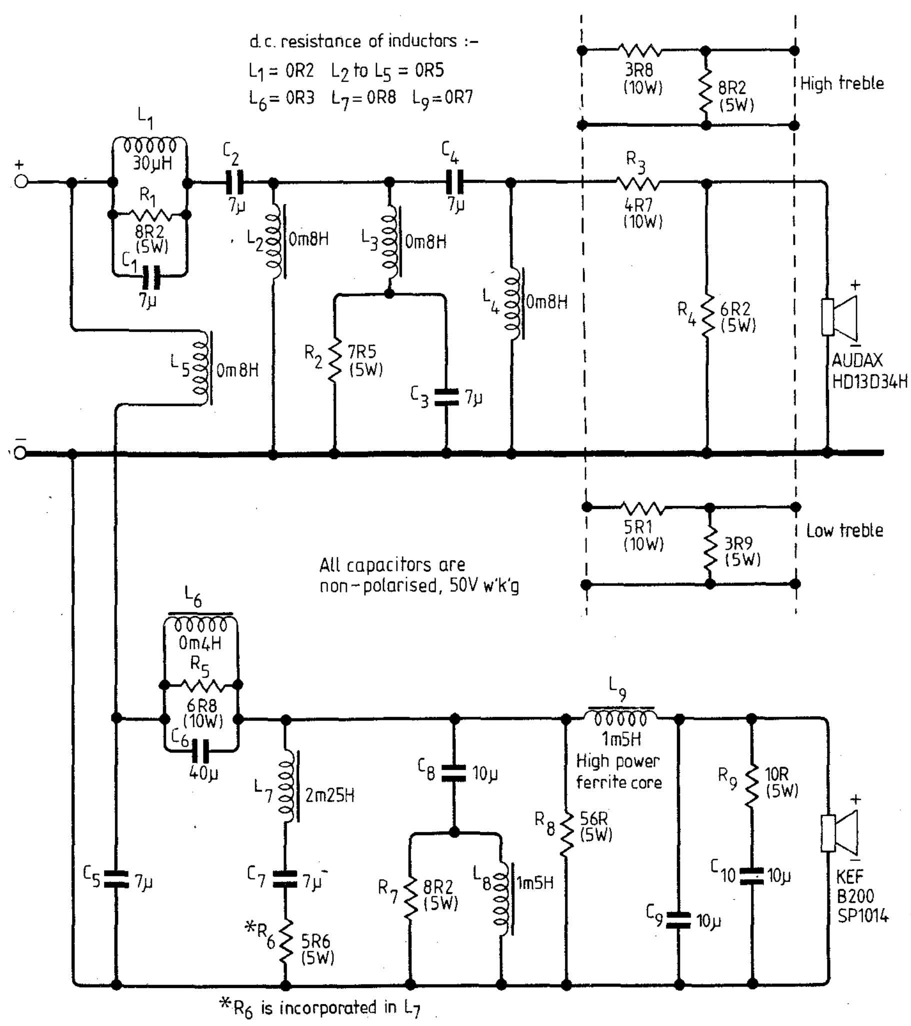
which conforms to the analysis of Dr F. N. H. Robinson (*Int. Journal of Electronics*, No. 2, 1980). In our circuit, the calculated n.t.c. distortion is about 0.13% at 20Hz which would give a distortion figure of 0.05% at the output of A_3. The measured distortion is 0.1%. The reason for this difference has not been determined exactly, though it looks as if H decreases at increasing frequency. This could be explained by the spherical shape of the n.t.c. material which causes a non-uniform current density and hence, especially at higher frequencies, a non-uniform temperature variation inside the n.t.c.

Book-shelf loudspeaker improvements

An article by J. Wilkinson describing the design and construction of a high-quality book-shelf loudspeaker was originally published in the October 1977 issue and improvements to the design followed in the June 1979 issue. Subsequent testing has prompted further small improvements.

Three small component changes in the crossover circuit have been made. One of these, namely changes in value of R_3 and R_4, has resulted from critical listening and comparison tests and gives a few dB attenuation in all three switch settings to compensate for room reflections of the tweeter's output. Changes in the values of R_5 and R_6 give a little extra dip in the crossover's output response curve at around 1kHz to compensate for a peak in the woofer's response curve at this frequency. Connecting the input of the low-pass filter before, instead of after L_1 gives a virtually inaudible improvement in performance but is nevertheless the best option from a theoretical viewpoint.

Extensive listening tests have also revealed a slight deterioration in sound quality caused by the 'anti-reflection' fillet attached to the bass-unit sub-baffle. The best solution is to replace the wood with 1/2in bituminous felt or similar material. A modified printed-circuit board, all the necessary components and the speakers can be obtained from Falcon Acoustics Ltd, Tabor House, Norwich Road, Mulbarton, Nr Norwich, Norfolk NR14 8TT.



*R_6 is incorporated in L_7

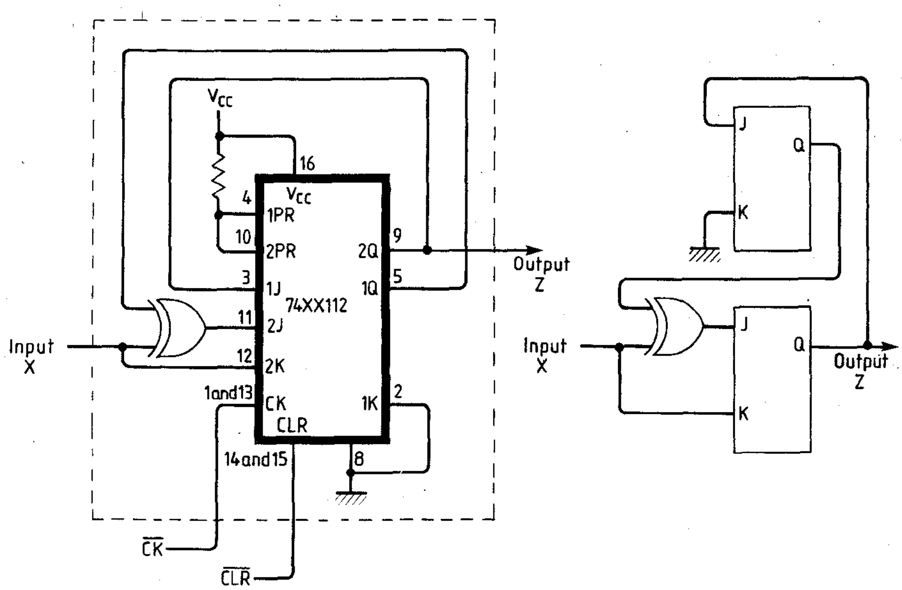
Circuit Ideas

Serial conversion to or from two's-complement

Digital number crunching is made easier if the two's-complement notation is used because it includes a unique representation of the number zero and has the ability to add or subtract numbers without concern for their sign. However, a problem can arise when a two's-complement output must be interfaced to a conventional system. This circuit converts to and from two's-complement with only two i.cs, regardless of word size.

The design is based on the algorithm that two's-complement can be formed by leaving all least-significant zeros and the first one unchanged, and complementing the remaining digits, i.e. start at the l.s.b., progress to the first 1, complement all digits after the 1. This can be achieved using a dual J-K flip-flop and one exclusive-OR gate. The number to be converted is fed serially to X, one exclusive-OR delay before the clock, and the clear line is synchronized to indicate the beginning of the word.

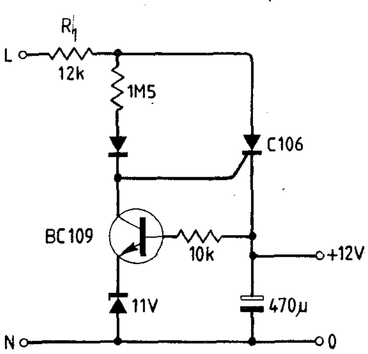
J. Okun
Santa Clara
California
USA

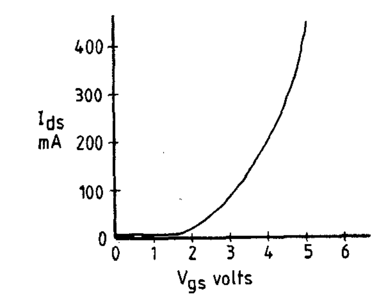


Low-loss power supply

This mains power supply is intended to drive circuits which are on continuously and normally require little power, but occasionally demand a larger current. Unlike a conventional series dropper or Zener diode controlled supply, when the load is off, no current is drawn from the mains.

When the voltage across the capacitor drops below 12V, the transistor turns off and allows the next half cycle of mains to trigger the s.c.r. which then recharges the capacitor. The circuit was developed for c.m.o.s. which normally consumes around 1μA, but occasionally needs 10mA to operate a relay. Resistor R_1 limits the maximum current available from the mains supply.

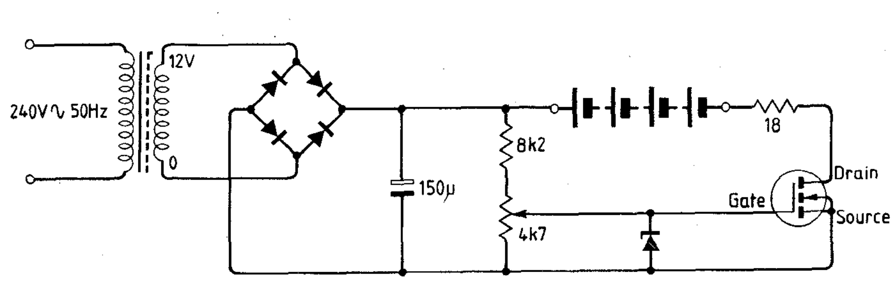




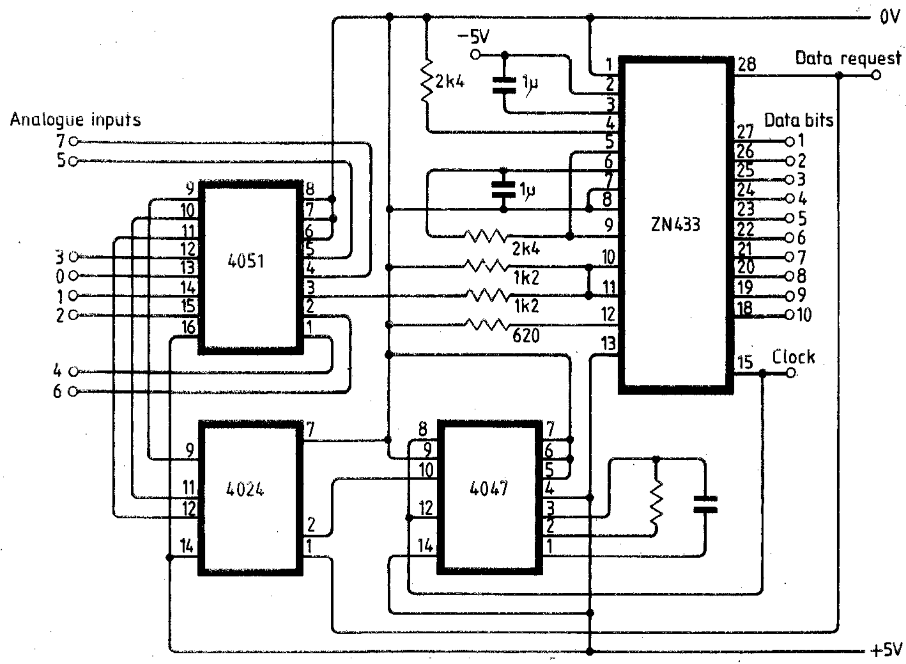
Nickel-cadmium battery charger

A simple constant current charger for NiCd batteries can be constructed using a power v.m.o.s. f.e.t. instead of the usual bipolar device. By varying the gate-to-source voltage from about 1.5 to 3V, the drain-to-source current I_{ds} can be varied from 0 to 100mA. The Zener diode shown across the gate and source is contained within the VN10KM package. No circuit parameters are critical, but the supply must be several volts above the maximum voltage of the battery to be charged.

A. C. Dickens
Leicester



Multiplexed 10-bit a-to-d converter



If analogue data is to be fed to a computer it must first be processed by an a-to-d converter. 8-bit devices, although fast, provide poor resolution while 12-bit types are slow and require sequential transfer of two words to an 8-bit bus. The ZN433 10-bit device provides a good compromise and can be connected to the user port of a Pet computer. However, as the available output lines on the Pet are limited, two features have been added to allow multiplexed inputs. A single line is used to provide the data request, which triggers on a falling edge, and the multiplexer counter increment, which triggers on a leading edge. Secondly, asynchronous use of the counter is achieved by wiring a monostable, with a period of two clock cycles, to the counter reset (stopping the clock resets the multiplexer to its first channel).

The software is written in 6502 assembly language for a 3032 Pet, and uses the i/o facilities of the user port which comprises 8 data bits ($ E84F), 2 data bits ($ E810), an output line ($ E840) and the CB2 output line. Before entering the program, the USR function and CB2 must be initialised by setting $0000:4C, $0001:80, $002:03, $E84C:CC. The result of the conversion is then available by using the Basic instruction

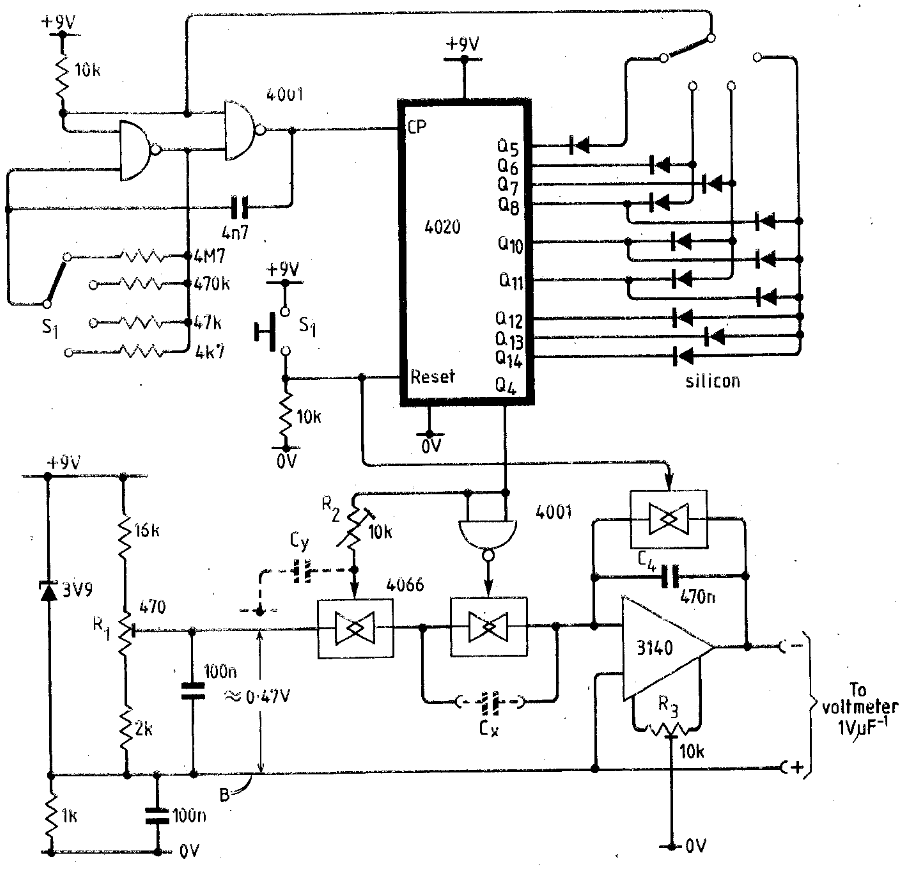
A = USR(B):PRINT A

where B is the multiplexer channel number. The USR routines, which are part of the computer operating system, appear in different locations for different models, and the appropriate entries are shown below.

D. A. Hills, E. D. Harvey
and S. F. Brown
Nottingham

CBM model	1000	2000,3000	4000,8000
floating to fixed $	D6DO	D6D2	C92D
fixed to floating $	D278	D26D	C4BC

Capacitance meter

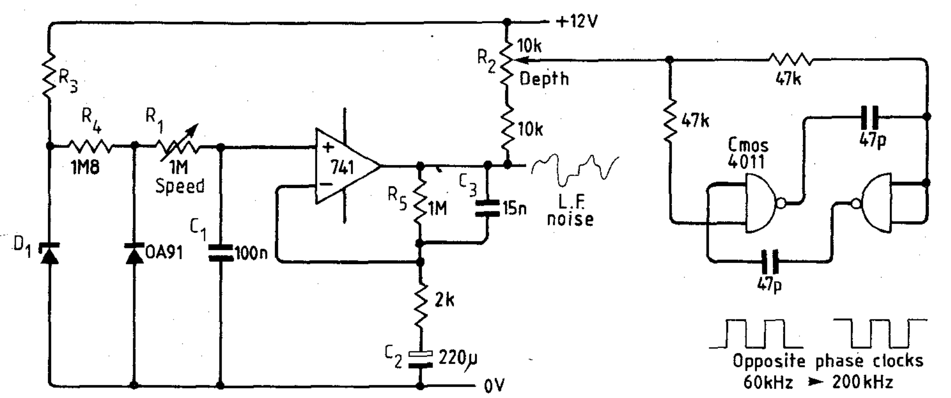


A direct reading of capacitance can be obtained on a d.v.m. with this circuit. The op-amp forms a switched capacitor integrator where the capacitor to be measured is repeatedly charged to V_{ref}. The number of charge/discharge cycles is determined by S_1, and the reference voltage is set by R_1 with respect to point B. When S_2 is pressed, the binary counter is reset and C_4 is discharged, which sets the output to 0V. When S_2 is released, the counter is enabled and counts to 1, 10, 100, 1000, at which point the oscillator is disabled and the counter stops. The timing resistors are scaled to give roughly equal measurement periods. After a measurement, the stored voltage at the integrator output is equal to the value of $C_x \times$ the range multiplier.

With C_x disconnected and after a measurement cycle, R_3 is adjusted to null the output offset on range 1, and R_2 is adjusted on range 4 to compensate for timing delay caused by the gate. Calibration is achieved by repeatedly measuring a known capacitor and adjusting R_1 for a correct reading. Extra ranges can be added by using different counters, but switching inaccuracies will also be increased.

M. Slater
Druids Heath
Birmingham

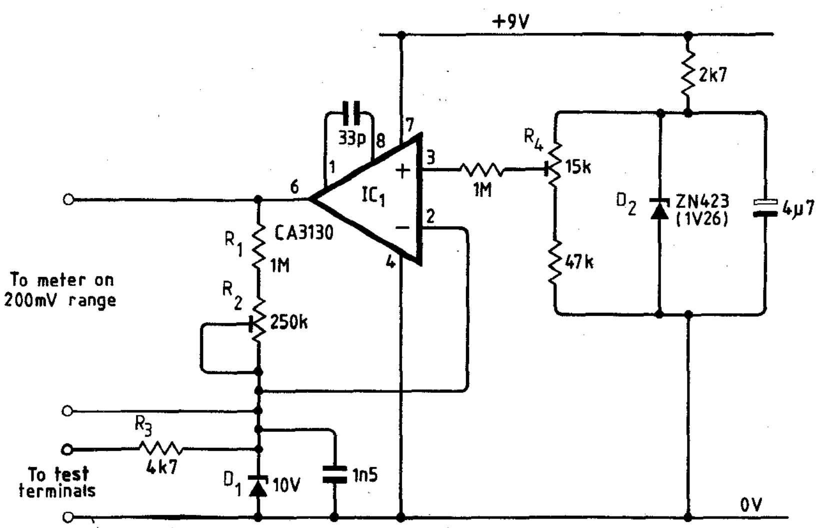
Random sweep for phaser



Most musicians are familiar with the effect of phasing or flanging which gives a "spacey" sound to a guitar or keyboard. This seems to work well when playing with other instruments, but in solo passages the regular sweep backwards and forwards can become noticeable. This design provides a random sweep for a c.c.d. phaser or reverberation unit, and can be useful for double-tracking or adapted for use with f.e.t. controlled phasers.

A reverse biased germanium diode provides a noise source and is used in preference to a Zener diode because the low-frequency content is greater. A 741 amplifies the noise, C_2 gives infinite feedback at d.c., C_1 and C_3 filter the high frequencies and R_1, R_2 provide speed and depth control. The v.c.o. is formed by a c.m.o.s. astable multivibrator and produces the required anti-phase clock signals for the bucket bridge. Resistor R_3 and the Zener diode prevent any low-frequency noise on the supply lines from reaching the amplifier input. Supply voltages from 8 to 18V can be used provided D_1 is altered accordingly and R_4 is adjusted until the op-amp output is midway between the supply rails with R_5 shorted.

C. Malloy
Darlington
Co. Durham



Conductance meter

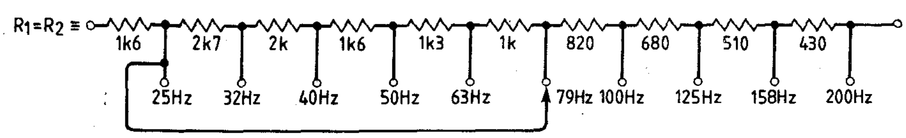
Equivalent resistances up to 10,000MΩ can be measured with this simple circuit. Pin 2 of the op-amp is kept at 1V by a current supplied from pin 6, and the voltage dropped by this current flowing through R_1 and R_2 is fed to a d.v.m. The meter should be set to 200mV and have an input impedance of over 5MΩ. D_2 and the associated components provide a stable 1V reference, while R_3 and D_1 protect the circuit with inputs of up to 100V.

To adjust the circuit, connect the d.v.m. to the test terminals and set R_4 for a reading of 1V. Connect the meter to the op-amp terminals and, with an accurate 10MΩ resistor connected to the input, adjust R_2 for a reading of 100. The meter gives a direct readout of conductance (inverse resistance) in nanosiemens (nS). Care should be taken in construction to reduce leakage currents around pin 2.

The circuit can measure leakage in capacitors and between p.c.b. tracks and, if a diode is connected to the input, leakage current at a reverse bias of 1V is displayed in nA. When a capacitor of over 100pF is connected to the input, the small charging current will be indicated on the meter. Conductance readings can be converted to resistance by using the formula 1000/(nS reading) = MΩ.

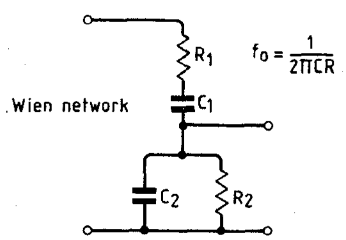
J. Pigott
Clonskeagh
Dublin

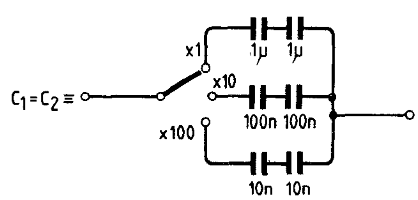
Spot-frequency oscillator



Switched values of frequency, equally spaced on a logarithmic scale, greatly reduce the time taken to measure frequency response. This network provides ⅓-octave intervals for a Wien bridge oscillator. In addition, resistance matching between the arms of the oscillator is much better than with a dual potentiometer.

S. Landin
Moss side
Manchester





World of Amateur Radio

New bands?

Wartime servicemen, American and British, had an expression "snafu" (bowdlerised as "situation normal-all fouled up") that has been described by lexicographers as "one of the few really good coinages of the war". Unfortunately "snafu" appears to be the only term that can adequately describe the current confusion over the use (or non-use) by British amateurs of the new 18 and 24 MHz h.f. bands. These bands were awarded to amateurs at WARC79 subject to the completion of the satisfactory transfer of all assignments operating in these bands and recorded in the Master Register. Last September the Home Office let it be known that British amateurs would be permitted to use these bands from January 1, 1982 on a secondary non-interference basis: but two months later, it said "No", pointing out that it had no right internationally to authorize such operation, although the question was still "being examined" to see if permission could be granted on this basis before the completion of the transfers (which could take up to 1989!). As these notes are being written (early December) the matter is unresolved.

The Home Office, however, has been quick off the mark in making less welcome changes: 200kHz has been taken out of the special UK 70 MHz band which now becomes 70.025 to 70.500 MHz. Similarly the 1.3 GHz band is now restricted to 1240 to 1325 MHz instead of 1215 to 1325 MHz (satellites 1260 to 1270 MHz), a loss of no less than 25 MHz! The January 1 profit and loss account was thus: profit 50 kHz at 10.1 MHz, loss 25.2 MHz higher up the spectrum – a result hardly in accordance with the intentions of WARC79. The voluntary 70 MHz band plan becomes: 70.025 to 70.075 MHz beacons; 70.075 to 70.150 MHz CW; 70.150 to 70.260 MHz s.s.b./cw (70.200 MHz s.s.b. calling frequency); 70.260 to 70.400 MHz all modes (70.260 MHz national mobile calling frequency, 70.300 MHz r.t.t.y. calling, 70.350 to 70.400 MHz Raynet); 70.400 to 70.500 MHz f.m. simplex (70.450 MHz f.m. calling frequency).

TVI to come

Radio-frequency interference problems, like accidents, don't just happen but are often created. In the USA, some 144 MHz operators are finding that local cable networks are now distributing television programmes on channels within the 144 MHz band. This creates a two-way problem: strong local transmissions break into the often rather "leaky" cables and interfere with viewers' television; in the reverse direction, sufficient tv signal may be radiated from the cable to mar the reception of distant amateur signals. So far this does not seem to be a problem in the UK but Shaun Shannel, G3ZSU has pointed out that 144 MHz is one of the authorized frequencies for the British 7-channel cable systems that are based on the use of a 159.625 MHz pilot carrier. As the number of cable channels increases over the next year or two here is an r.f.i. problem in the making.

Fake QSLs

Recently I reported an ARRL investigation into the authenticity of some QSLs cards emanating from several much-publicized "dx-peditions". Further evidence of practices that could bring the whole amateur operating awards system into disrepute can be found in "Amateur Radio", the journal of the Wireless Institute of Australia. Ken McLachlan, VK3AH in his October "How's DX" column under the heading: "Blank QSLs – Fair-play sport?" writes: "Amazing things turn up in our mail box but the contents of an unsolicited letter from an amateur dx-peditioner and QSL manager well known in Europe, really set me thinking. It contained a number of QSL cards, duly signed, but the pertinent details were left blank, also a little note accompanied them saying that if I didn't want them I may know someone who did. I know half of VK (Australia) would . . . This is a very serious situation, coupled with some amateurs using 144 MHz links with a friend in a better location, using a friend's callsign to get him a report, or getting a friend to operate your station . . . It is not ethical or within the rules of fair play. To what extremes will some amateurs go in a hope of achieving honour roll status?"

Such practices are, I believe, still uncommon: most QSL managers and dx-pedition organizers are scrupulously correct in the distribution of the eagerly sought after "confirmations" – but even a few . .

Expanding space

Following the successful launch of the British UOSAT – OSCAR 9, membership of AMSAT-UK has passed the 1000 mark and some 2000 enquiries were received by the group in the weeks following the launch seeking technical and other information, many from schools and technical colleges. Ron Broadbent, G3AAJ, has expressed disappointment however that the media have been referring to this amateur radio experimental satellite as the "British Schools Satellite" although no part of the cost of the project was contributed by schools. All on-board systems have been checked-out although the satellite is taking time to stop spinning.

A rumour that beacon transmissions heard on 29.331 MHz denoted that the expected new Russian satellites, RS3 and RS4, were up and working proved false: the signals were from a satellite unit under test in the Moscow area. The new Soviet satellites will each carry 145/29 MHz transponders (up 145.860 to 145.900 and 145.910 to 145.950 MHz: down 29.360 to 29.400 and 29.410 to 29.450 MHz). Amateur enthusiasts regard further active transponders in medium height orbits as their prime requirement. AMSAT-UK members are "not particularly enthusiastic" about further purely research experiments that do permit two-way working.

Here and there

The British Amateur Radio Teleprinter Group, in order to promote more interest in r.t.t.y operation on the v.h.f./u.h.f. bands is introducing awards for amateurs and listeners showing they have worked or heard: (1) 100 different stations on 144 MHz; (2) 50 on 432 MHz; and (3) 10 on 1296 MHz. Endorsement stickers for each additional 25 (1296 MHz 10) stations to a maximum of 200. Rules from (s.a.e.) Ted Double, G8CDW, BARTG Contest and Awards Manager, 89 Linden Gardens, Enfield, Middx, EN1 4 DX.

In the U.K. for the opening of c.b. was Al Gross who can claim to be the single individual most responsible for the start of c.b. in America. Al Gross based his ideas on his work on the OSS's wartime Joan-Eleanor v.h.f. (260 MHz) transceiver used in 1944 to work agents from Mosquito aircraft – and seemingly a later version of SOE's "S-phone" (450 MHz) originally developed at St. Albans by Bert Lane. In 1945 Al Gross was assigned the experimental call W8XAF to develop 465 MHz c.b.

In brief

Cyril Parsons, GW8NP and 1976 president of the R.S.G.B. has died. He had recently been actively involved in encouraging disabled people in the Cardiff area to take up amateur radio . . . The R.S.G.B. is now maintaining a computer file on stolen equipment . . . Garry O'Keefe-Wilson, G4MIA is the new chairman of the Wirral Amateur Radio Society . . . Fewer transAtlantic signals have been heard on 50 MHz this winter although some 28/50 MHz cross-band working has proved possible . . . The East Suffolk Wireless Revival mobile rally will be held at the usual Ipswich venue on May 30. The Northern Mobile Rally is on May 23 at a new venue: the Great Yorkshire Showground, Harrogate .

PAT HAWKER, G3VA

Letters to the Editor

Failure of distress signals at sea

During the past few years you have published about a dozen letters on the above subject. They seem to have brought to light a number of different problems.

To solve a particular problem you have to face it in its totality, before going into the details. So, first, we have to describe the circumstances:

- There is a good conductive medium – the sea with things in it like fishes, pollution, and ships.
- Upon this is a dielectric layer, consisting of air saturated with salt water spray or droplets.
- Over this layer is the so-called "ether".

The boundaries of the dielectric layer are not as well defined as we would like them. At the bottom is the boundary of the sea surface. The dielectric property of the layer decreases with increasing approach to the sea surface. The thickness of the layer depends on weather conditions and will vary from a few centimetres in fair weather to a few decades of metres in gale conditions.

Because the ship belongs to the good conductive medium, the aerial system will not outreach this dielectric layer in bad weather conditions. This is especially true for lifeboats, with their even smaller aerial systems. Because the antenna wire and feed-through insulator under these circumstances are coated with a salt water film, these parts can almost be seen as a part of the good conductive medium.

Secondly, there is the physical principle by which the distress communication takes place. This happens by means of 600m long electromagnetic waves at a frequency of 500 kHz. At this frequency e.m. wave propagation takes place by means of the surface wave. It means that the upper half of the electric part of the e.m. field is mirrored by the top layer of a good conductive medium, in this case the sea surface. So you can say they are walking over the water.

Because the losses are small this gives a reliable means of communication. The above-mentioned dielectric layer has only a small influence on this type of wave propagation at these frequencies because this layer is much smaller than the wavelength and the E field stands almost perpendicular to this dielectric layer.

When using short wire or whip aerials there are three barriers against helping these e.m. waves onto their feet. This type of aerial is much smaller than the wavelength so the radiation resistance is low and the aerial behaves like a capacitor, which ought to be tuned by a large inductor with its own resistive losses. So the efficiency from this matching system is low. As explained above, in gale conditions the aerial is short-circuited by a so-called "salt water capacitor". Finally this electrical type of antenna is stretched out in the dielectric layer which in its turn absorbs a part of the remaining energy.

Let us now consider the small magnetic loop antenna. The small loop antenna also has a low radiation resistance but the aerial behaves like an inductor, which should be tuned by a large capacitor with its (relative to the tuning inductor) smaller resistive losses. So the efficiency of the matching system will be higher. In gale conditions the loop antenna is in parallel with the capacitor produced by a film of salt water. In fact this capacitor detunes the aerial system a little, but this can be cancelled out by the matching capacitor unit. Finally this magnetic antenna is placed in the dielectric layer which has only a negligible influence on it.

At about a quarter wavelength away from this magnetic loop antenna, the electric part of the e.m. wave is formed. This happens at positions far enough above the sea surface and the dielectric layer. In this indirect way the 500kHz surface waves are put on their feet, so they can walk over the sea surface carrying with them the distress signals.

In my opinion the magnetic loop antenna wins on several points against the electrical short wire or whip antenna, but the price of the loop and its matching unit will be higher – but what can be the price of human safety? At the moment I am working on a second type of loop aerial, but it seems impossible to simulate gale conditions in my laboratory because it is no film studio.

R. R. Venekamp
Eindhoven
Holland

Amateur licences in Germany

I read with interest Peter Saul's letter in the November issue on the under-utilisation of the 28MHz amateur band by class A licence holders. First let me correct his statement that no test is required in West Germany for 10-metre operation. In Germany we have class A, B and C licences. The C licence holders enjoy roughly the same privileges as B licensees in the UK, with some differences in power output and, from 1st January 1982, v.h.f. and u.h.f. allocations.

Licensees wishing to operate below 144 MHz must have passed a Morse test and, briefly, the requirements are as follows:

Licence Class	Morse requirement	Amateur bands
A	60 letters per minute (Tx/Rx)	All amateur bands
B	30 letters per minute (Tx/Rx)	3.5- 3.8 MHz 7.00- 7.1 MHz 14.00-14.350 MHz 21.00-21.45 MHz 28.00-29.700 MHz plus v.h.f./u.h.f. bands
C	None	v.h.f./u.h.f. bands only

As an ex c.b.er (we've had c.b. in Germany since 1975), I agree that the 27-28 MHz bands are indeed a very interesting part of the spectrum; the 12-channel a.m. c.b. band is frequently swamped by S9 signals seemingly eminating from the USA. However, if Peter was prepared to study for his ticket in the first place, surely a little extra swotting shouldn't put him off if he really wants to go h.f. After all, this could also be considered as fulfilment of the self-training clause of the licensing conditions.

I would agree that £12.00 just for a Morse test does seem rather steep but I'm somewhat out of touch with price levels in the UK. The fee here is DM40 – which includes both the technical exam and the Morse test (a repeat of either costs DM20). I'm working towards the Morse test next year.

No, let's keep the licensing conditions as they are. If anything needs changing it's the Home Office's indifferent attitude to pirates and *illegal* c.b. operators.

V. A. Sancto, DDFM (G6BWH)
Bad Aibling
W. Germany

Recharging dry cells

Following your recent items on recharging dry cells (August 1981 issue, pages 46 and 70), I have been investigating the Rostlund patent application at the SRL (Patent Office Library). It is much easier to find under its publication number, WO79/01061, and is owned by BE LE INVENT A/B, Stockholm.

The specification (in English) gives a wealth of details of suggested circuits and component values most unusual in a patent but prefers to use only slightly discharged cells.

It is well worthy of study by those interested but, for my part, I am quite happy with nickel-cadmium cells and chargers especially now that they are available at such reasonable prices from firms such as Argos, Comet and Jessop of Leicester.

F. E. Smith
London W5

Intentional logic symbols

J. E. Kennaugh's letter (October 1981) concerning Intentional Logic Diagrams (Cassera, November 1980, pp61-62) has missed the point of the intentional symbols. Diagrams should speak for themselves with very little need to revert to textual descriptions. Logic diagrams in particular should indicate the logic function that is being performed and if a gate is performing an OR function on low asserted inputs then it should be drawn as a low asserted OR gate and not as a NAND gate.

Not only is the intentional diagram much clearer as to the function of the gates (Fig 1(a) is

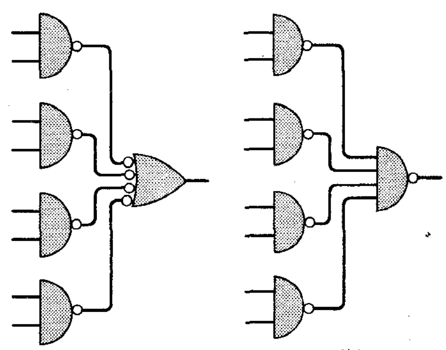
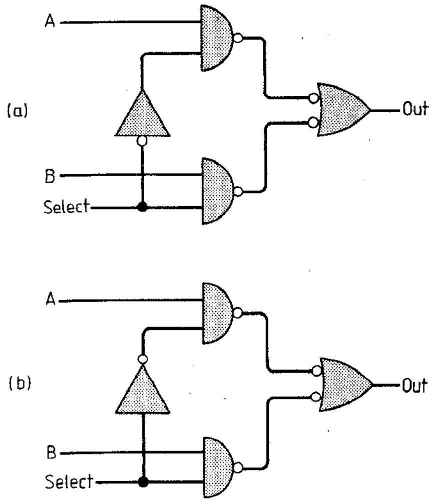

(a) (b)

Fig.1

clearly a sum of products whereas Fig 1 (b) is not clearly so), but what is of equal importance is that the assertion level of both the inputs and outputs is clear. If the assertion level of a signal is low, it should be derived from a circled output to indicate this. Similarly a low asserted line should connect to a circled input of a gate. A general rule is that circled outputs should connect to circled inputs and non-circled outputs should connect to non-circled inputs. Sometimes this rule needs to be broken (some purists state that the connection should not be shown but that the signal should be named and the inverted name with a bar over should be shown on the input, but this is possibly taking things too far and adding confusion). A typical example of the rule being broken is for a 2 to 1 multiplexer:



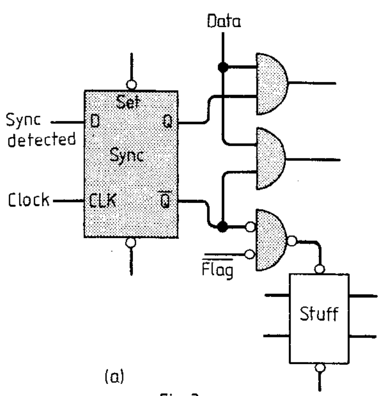
(a) is preferred to (b)

Fig. 2

From Fig. 2(a) it can clearly be seen that a low on the select line selects the A input, whereas from Fig. 2(b) a little more thought is needed as a high on the select line de-selects the A input.

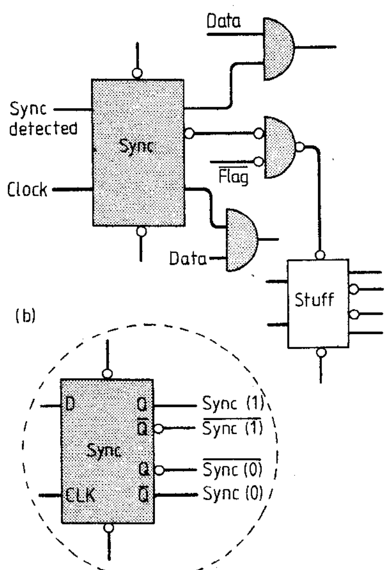
This intentional logic diagram notation can be extended even further than the original article by Tony Cassera (November 1980) to take account of flip-flops. Flip-flops have two states (Set, Reset) or (asserted, not asserted) and two outputs Q & $\overline{Q}$ (although for a D-type flip-flop both Q and $\overline{Q}$ are high when both Set and Reset lines are asserted). However, a typical circuit drawn with just two outputs does not fit well into the intentional logic system and, as a result, it is not always clear from the diagram what logic function is being performed.

In Fig. 3(a) it is not clear that the flip-flop Stuff should only be asserted when the Sync



(a)

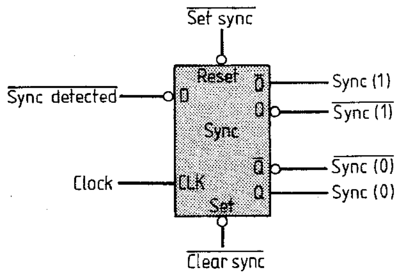
Fig. 3



flip-flop is set (presumably as Sync-detected has been seen) and $\overline{\text{Flag}}$ is asserted. In the re-drawn flip-flop, an asserted high and asserted low output can be clearly seen as well as an unasserted high and low output. This means that signals can relate to the flip-flop asserted condition and there is no need to perform mental contortions in deciding whether the flip-flop needs to be asserted or not before the second flip-flop sets.

The D-type flip-flop still only has two output pins (Q and $\overline{Q}$) but showing each output in two different positions can clarify the functions of the flip-flop within the circuit as a whole.

This idea of intentional logic can be further extended to produce "re-asserted D-type flip-flops." It may be desirable in a logic system to have a low signal to assert the D-type flip-flop. In this case the D input is shown with a circle and a 0 on the D input will assert the flip-flop and a 1 unasserts it. In this case the Q and $\overline{Q}$ outputs reverse roles and so do the Set and Reset inputs.



To those who have never used intentional logic diagrams they seem very alien and peculiar but once they have been accepted and tried anything else becomes second-class. Intentional logic diagrams have been well established for many years and at least one major computer company (Digital Equipment Corporation) uses such a system and with many years of experience (and probably the biggest user of logic diagrams) this system has proved very successful. With many more functions being performed on a chip, logic diagrams are using fewer and fewer gate and flip-flop symbols and more and more meaningless boxes. The more help that the few remaining gates and flip-flops can be the better.

J. E. Kennaugh's method of logic symbols relies on reading the truth table of each gate which may indicate the transfer function of the particular gate but says nothing about the function of that gate within the circuit as a whole. The logic diagram should "speak for itself".

Christopher Hudson
Computer Systems Laboratory
Queen Mary College
London E1

James Clerk Maxwell

It seems very remarkable, having regard to the now almost classical experiments in atomic physics of the 1920s and 30s by Rutherford, Cockcroft, Walton and others, which led to the invention of the atomic pile, atomic bomb, atomic reactors, fusion research, etc., that Mr Wellard cannot appreciate the meaning and validity of $E = mc^2$ (Letters, October, 1981) which he describes as a "meaningless, misleading" equation.

Firstly, the equation has been very thoroughly experimentally verified. Secondly, it is, as Mr Wellard shows, quite sound dimensionally. Thirdly, it surely relates to the equivalence of rest mass and energy, and not to the dynamic energy of an accelerating or moving mass, as Mr Wellard seems to want. He seems confused on this point. Also, from the equation we can simply derive, using a little calculus and Newtonian mechanics, a further equation showing that mass increases with velocity, becoming infinite at the velocity of light, c. This latter equation was also derived by Einstein and has again been verified by, e.g., experiments with very high speed electron beams, and particle accelerators.

By all means, Mr Wellard, let us encourage the spread of the philosophical spirit, but we must surely begin by agreeing on what determines the meaning and validity of our equations. It seems quite unfair to start by claiming that Einstein's theory is "safe from experimental verification".

Peter G. M. Dawe
Oxford

The big c.b. con

We are not sure at whom Mr Wheeler's accusing finger is pointing. Natcolcibar and the Citizens' Band Association are the two national associations which the Home Office, at any rate, has recognised as representative enough to be called into consultation. May we point out on their behalf that at no time has any alleged suitability of 27MHz a.m. for dx working been either explicitly or implicitly any part of their reasons for objecting to the Home Office proposals for f.m. on unique frequencies. Indeed, both of us advocated, until it became obviously hopeless to do so, a v.h.f. allocation.

We regard dx-ing and the exchange of QSL cards as being as peripheral to the uses of c.b. as, say, train-spotting to public transport or philately to postal service operation. But, once it became clear that British c.b. would be in the 27MHz band at all – about the most unsuitable possible part of the spectrum for mobile radio but that which, for historic reasons, is by now in almost universal use for c.b. – we did and do say that it is absurd to start from scratch a system operating on unique and completely incompatible standards. The European CB Federation's request, over a year ago, to the EEC to promote a common specification throughout Europe and a common user's licence, a request which we strongly support and which is being actively followed up by the Commission, is based not on a desire for international communication across distance, but on the consideration that c.b. is often at its most useful when one is travelling away from home, and with international road travel already common and becoming more so it is a total nonsense to have to leave one's c.b. rig at home.

Surely the real con trick is perpetrated by those who pretend that f.m. (to the Home Office MPT 1320 spec.) gives superior results. Certainly f.m. with an advantageous deviation index can give good results – at a cost in bandwidth which has not been allowed to c.b. Narrow-band f.m. can also be satisfactory where dedicated channels and designated service areas guarantee a minimum signal strength and freedom from co-channel interference. But in a random service where there is contention for use of channels, a.m. really is indicated as the primary mode – which is no doubt why it is used by aircraft in the v.h.f. band, and perhaps why the FCC will not authorise f.m. even as an option for 27MHz c.b.

The Home Office never pretended they were specifying f.m. in order to give a better c.b. service, of course: they said we must have it because it causes less interference (which, in itself, is at best a half truth!). Surely the claims that it gives a superior performance – which are already proving to have been quite false – came after the event, from the necessity to sell what you have been given no option about producing?

James Bryant
Citizens' Band Association
Ian Leslie
Natcolcibar and
European Citizens' Band Federation
London N10

Poor deal for amateur radio

I was not at all surprised personally to read in your journal and others that class B amateur radio licence holders are not to be allowed to use the 4 metre band. In fact I consider it a minor miracle that the RSGB gets any concession at all from these faceless and all powerful government employees who, it seems, are only to be swayed by outright, blatant "radio anarchy".

For many years amateur radio has been a service enjoyed by very many, including myself. Understandably it has to be regulated, and controlled . . . but why so heavily handed and authoritarian? One is almost drawn to a wry smile at the pathetic bleats now being heard regarding the possible future withdrawal of 4 metres. It's too late RSGB, they have pushed you around for years, although it does not seem quite that way via the "Old Boy Network" and "Old School Tie" club.

For very many years dedicated supporters of the RAEN have been ready and willing to offer emergency help, but have been largely ignored, mainly I feel because *we* just cannot turn up and fill the "communication gap", as others do, who, now able to drop their incredibly imported American verbal disguises, have revealed the extent of their organisations.

As a free-lance writer, I have found it a terribly frustrating time, wondering mostly about the lack of comment from the RSGB on these matters. Perhaps I am being unfair, maybe someone from RSGB did try to correct blatant ignorance, but was met by "What's the RSGB?" and "What's amateur radio?". Perhaps if you mentioned "Boy Scouts" they would have realised. It was maddening to see the enormous lineage given to these people, whilst they were breaking the law . . . and promising to carry on using a.m. rather than the new f.m. service.

In my mind there is no doubt that in the eyes of the invisible "government employee" an organisation that does as it is told, has well behaved menbers who do their level best to conform to the rules of society in general (even to being forced off the air through no fault of their own) does not deserve respect, but can in effect be ignored.

The British radio amateur is not even allowed to let his wife's voice be heard returning a greeting over the air, and by becoming an amateur literally gives unrestricted access to officials on demand . . . to examine the station.

And along came "citizens' band", long a cause of trouble in America and virtual anarchy alongside to help it achieve its goal. Unfortunately, perhaps because there have been no "novice or beginners' " licences available in this country the "Open Channel" seems to have developed into some sort of substitute for amateur radio. Only once in recent years has any one person in central government stated his intentions to make amateur radio more accessible to newcomers etc . . . he was sacked in less than two months.

In closing let me state quite clearly that I am not against citizens' band radio (I hold a licence myself), but I use it as it was intended, as a service and not as an end in itself. I feel that the "powers-that-be" brought all this down on themselves by refusing to acknowledge the need for such a service, allocated on a British channel well away from possible verbal pollution originating from North America. I have enraged local 27MHz f.m. users by speaking plain English, with the absolute minimum of formality . . . with the result that they refuse to talk to us, thus leaving the channel clear to use over our smallholding in peace.

Come on RSGB, and amateur radio in general, stand up for your rights! If anarchy can do it, and force its wishes on us all, let us all unite and show what true democratic rational thought can do . . . despite civil servants. I would rejoin the RSGB immediately if I thought this would happen soon.

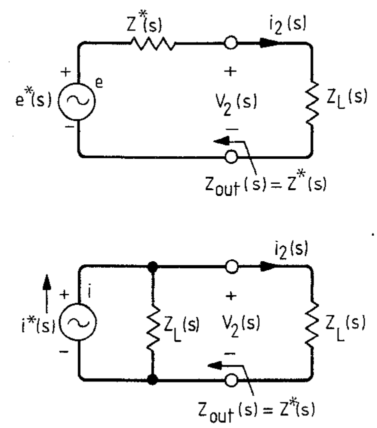
Robert B. Mannion G3XFD
Alresford, Hampshire

The image equivalent-generator theorem

While it is true that Thévenin's theorem handles dependent sources, it does so in a roundabout, time-consuming way, forcing us to use the Applied Source method, in which we attach a source to the output port. This means that we reverse the direction of signal transfer in the network. This method of determining generator impedance is often cumbersome. The same criticism applies to Norton's theorem. In contrast, the well-known Thévenin-Norton Function-Source Theorem elegantly yields the generator impedance as $Z^*(s) = e^*(s)/i^*(s)$, where $e^*(s)$ is the open-port voltage and $i^*(s)$ the closed-port current.

In using Thévenin's theorem, we are actually deriving a transfer function, but one that does not include the load impedance. However elegant, this is the particular step that robs us of the information about the generator impedance and forces upon us the Applied Source method. The same is true for Norton's theorem. One cannot help wondering if there does not exist a simple method of retaining the load, and thus side-stepping the sometimes awkward Applied Source method. A solution to the problem has been found, and yields the following theorem:

> The series-form and parallel-form generator equivalents of a linear network have a generator impedance $Z^*(s) = Z_{OUT}(s)$, obtained when the output port is terminated in $-Z_{OUT}(s)$, having a current source $i^*(s)$, which is the load current for closed output port, and a voltage source $e^*(s)$, which may be obtained either as the open-port voltage, or is calculated from $e^*(s) = Z^*(s)\, i^*(s)$.



The proof of this theorem lies in Tellegen's Theorem, stating amongst other things that for lossless condition, the impedances on both sides of a port cancel. The load sees its image, of opposite sign, and this new theorem has therefore been called the Image Equivalent-Generator Theorem. It allows direct construction of both the series-form and the parallel-form generators with all dependent sources automatically taken care of. It is a time-saving theorem. Fig. 1 has been included to exemplify the meaning of the series-form and the parallel-form equivalent generator. Because this new theorem introduces a totally different method for the determination of the generator impedance, it provides an alternative to the Thévenin-Norton Function-Source Theorem, and is thus useful as a checking tool on obtained answers.[2]

Harry E. Stockman
Arlington, Mass.
USA

References

1. Stockman H. E. "Tellegen's Theorem – some applications", *Wireless World*, Feb. 1981, pp. 77-79.
2. Stockman H. E. "The Theorem Book", Sercolab, Box 78, Arlington, MA 02174.

Cordless telephones

I was amazed to see *Wireless World* publishing a display advertisement for cordless telephones (September 1981 issue, page 106), which are neither licensable for UK use nor British Telecom approved. The advert failed to mention either of these facts. These devices transmit in the frequency band 49.5-49.8 MHz, which is within television Band 1, Channel 2. According to BBC Engineering Information, Channel 2 is used by three main transmitters (Swingate Dover, North Hessary Tor and Winter Hill) and will continue to be used until 1985 or 1986.

Apart from the legal aspects and the interference with television broadcasts, the current generation of cordless telephones can cause problems if two people within the claimed 200-metre range have telephones operating on identical frequencies. As well as overhearing each other's calls, one might make an outgoing call from the handset and activate the neighbour's base station, thus making the call at the neighbour's expense.

The problem of security is reported to have been solved (*Electronics Times*, 3rd September 1981) by one manufacturer whose product is to be submitted to British Telecom for evaluation. If the Home Office allocates a frequency for cordless telephones in the near future it cannot be 49MHz, so equipment currently on sale will never be legal. The import and sale of 49MHz cordless telephones and transceivers should be banned, as was done for 27MHz equipment a

decade ago. If nothing is done, increasing unauthorised use of 49MHz will make the frequency useless for any other purpose after 405-line tv transmissions are ended.
D. M. Lauder
Barnet
Herts.

Microchips and megadeaths

I suppose one could go on endlessly about the subject of your November 1980 editorial, and I do not wish to do that. I would, however, like to reply to Mr Hind's letter in your August 1981 issue.

The two atomic bombs which were dropped on Japan put an immediate end to the hostilities which could have dragged on for years, killing many more than the bombs did. Furthermore the Japanese were not just killing and torturing soldiers but women and children as well. It is notable that since the bombs were dropped there has been no European war, except the one-sided slaughter in Hungary and Czechoslovakia.

The world does not change. It does not matter whether it is bows and arrows, flintlocks, supersonic aircraft or guided missiles, unless you have what the other man has then you will live under his rule, or die under it. The small countries that live in "freedom" do so very precariously and by virtue of the fact that we have two balanced super-powers, and I suspect that Mr Hind knows this fact only too well. I am just as capable as Mr Hind of conceiving what a nuclear war would be like and do not regard his question "do I hold my children responsible for treatment of prisoners by the British army?" as valid. However, I would proudly say that such prisoners have always been treated very humanely and in strict accordance with the Geneva convention. In any case does Mr Hind really believe that war is any longer restricted to the military or, indeed, that it ever has been? Just look at recent events all round the world.

If, as Mr Hind says, our freedom is illusory because engineers do not prevent the government from having nuclear weapons, it would be equally illusory, in fact non-existent, if they did.

In conclusion I would like to say that I think that the preceding letter in your August issue by Mr Belcourt puts the case extremely well.
L. G. Martin
Abbots Leigh
Bristol

"Spreading"

Mr Yates' letter on "Spreading" (October 1981) led me to make some tests, using a receiver with cascaded filters (about 100dB ultimate rejection). I found that:

1. Stations with signals peaking over S9+20 almost always trigger the S meter on the "suppressed" sideband (not surprising when you consider that 60dB of sideband suppression is not easy to achieve in a transmitter).
2. Stations producing S meter readings that hardly move down from the peak reading during speech almost always have splatter both above and below the necessary bandwidth, and often over more than the 8 or 10kHz mentioned by Mr Yates.
3. Most of the stations producing S meter readings with a variation of several S points between the peaks and the valleys of speech are clean above and below the necessary bandwidth, even though some of them are very strong (more than 500 microvolts across 50 ohms).
4. There were a lot of stations "spreading". An hour or so with a spectrum analyzer will confirm this.

Digging a bit deeper, "spreading" comes mostly from over-driven "linear" amplifiers, over-reliance on a.l.c., or misadjustment of the *output* control of the voice processor. I don't know how many wide receivers there are, but there are plenty of wide transmissions. One thing was very noticeable in the tests mentioned above – many very strong stations were very clean, and many lower-powered ones were "spreading".

The bandpass filter mentioned by Mr Yates does not restrict the transmitted sideband to about 3kHz – it only restricts the signal fed *to* the final to that bandwidth. What happens in an overdriven and wrongly biased final is anybody's guess.

But surely, even in Australia, where the proportion of phone operation is higher than most places, it would not be true that amateurs are "almost exclusively" s.s.b. I think Mr Yates meant that those *who operate phone* are almost exclusively s.s.b. when they do so. We mustn't mislead the general reader.
Bob Eldridge, VE7BS
Pemberton
B.C., Canada

The death of electric current

Dermond J. O'Reilly, whose letter was published in the December 1981 issue under the title "The death of electric current", must have missed my article under that title in the December 1980 issue. I wrote, "Electric charge does not exist according to Theory C," and yet a year later Mr O'Reilly writes, "Will [Catt] be announcing the death of electric charge next?"

In his third paragraph Mr O'Reilly attacks what I believe to be my accurate statement of the conventional theory. Surely he should be defending, not attacking, "our great heritage of scientific understanding"?

In paras. 4 and 5, O'Reilly makes the same mistake as Dawe made in the November 1981 issue, page 55. I wrote about the additional *charge* on a wire after the passage of the step, and did not mention the *current*. (See *WW* August 1981, page 40, para. 3) ". . . extra electrons must appear [in/on the wire]", *not* (extra) current must flow.

As to para. 6, if i and dD/dt are one and the same thing, then does it flow in direction BB′ (i) or in direction BC (dD/dt)? One current cannot flow in two directions at the same time.

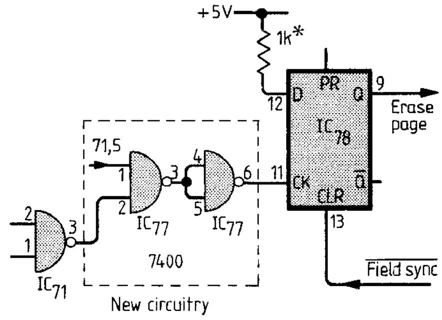
Para. 7, I wonder whether

$$\frac{E}{H}=\sqrt{\frac{\mu}{\epsilon}},\ H=\frac{B}{\mu}$$

were nonsense in Professor Bell's article, *Wireless World* August 1979, page 44? Or are the defenders of classical electrodynamics allowed to write such stuff, but it becomes nonsense when written by a dissident?
Ivor Catt
St Albans
Herts

WW teletext decoder

The erase page circuitry of the Wireless World teletext decoder will not function correctly when several magazines are interleaved, as now happens in the new 'high speed' Oracle service. This is because an erase bit detected in the header row of the selected page will be cancelled by one occurring in a header of a different magazine transmitted subsequently in the same field interval. However, if the modification for automatic clear (May 1979, p.86, Fig. 7) has been incorporated, it is a simple matter to wire the spare gates of the extra i.c.77 as shown to restore correct operation.



*1kΩ resistor added for purists who do not like o.c. inputs

Modification to ensure correct operation of the Wireless World teletext decoder when magazines are interleaved within the same field interval.

Firstly the tracks to (78,12) and (78,11) are broken. (77.1) is then connected to (71,5), (77,2) to (71,3) and (77,6) to (78,11). Finally pins 3, 4 and 5 of i.c.77 are wired together. This ensures that once an erase bit has been detected co-incidentally with a correct header row, (78,9) will remain at logic 1 until the next field sync pulse arrives, when the memories will have been completely cleared.
Alan Pemberton, G8ZHG
Sheffield
Yorks

The dream of objectivity

In response to Mr Dawe's letter in the December 1981 issue, I would like to quote Ronald Knox, who wrote:

There was a young man who said, 'God
Must think it exceedingly odd
If he finds that this tree
Continues to be
When there's no-one about in the Quad'.

REPLY

Dear Sir:
Your astonishment's odd:
I am always about in the Quad
And that's why the tree
Will continue to be,
Since observed by
Yours faithfully,
God.

I trust that Mr Dawe will not mind my observing that rainbows did not exist before the creation of man (Genesis 9, verse 13).
M. J. Walker
Department of Physics
University of Nottingham

Foot-controlled radio

I have just been reading your issue for May 1981 (I'm not a slave of time!) and am reminded by Mixer's comments under the heading "Traffic diversions" of a report in a *Wireless World* issue of about thirty years ago.

Somebody had invented, or at least marketed, a foot-controlled car radio by which one could change stations without taking one's hands off the wheel, in very busy traffic situations. I wish I had made a note of the actual wording. Mixer would have enjoyed it.
Ronald Gill
Allestree
Derby

Data recording on audio cassette

The solution adopted for the Open University's Radiotext project

by **P. Smith and P. I. Zorkoczy,** The Open University

For the past two years, work has been carried out within the Faculty of Technology of the Open University to produce a low-cost method of using a v.h.f. radio broadcast network to deliver computer software, text and graphic material for educational purposes.

The circuit to be described allows data at 2400 baud to be recorded reliably on any audio cassette recorder. The solution adopted may be pertinent to other applications, such as the storage of microcomputer software at this fairly high data rate.

It is intended that these transmissions will take place outside the normal hours of service of the broadcast network and, on reception, be automatically recorded onto audio cassette. This mode of operation not only provides a use for transmitter resources that would otherwise be idle, but enables the unmodified broadcast network to be used with radio receivers of conventional design.

The recorded material may subsequently be displayed on a conventional television or printed on a low-cost printer at a time convenient to the student. Recording allows the material to be studied at any desired rate with repetition if necessary. Figure 1 is an overall system diagram.

Reliable recording of the broadcast data is therefore an essential ingredient of Radiotext. A number of proposed methods[1,2] for high-speed data recording were examined but none of them satisfied the requirements of this application.

The audio cassette recorder

To appreciate the performance of the audio cassette recorder it is useful briefly to examine the record-playback process.

The waveform to be recorded is applied as a current to the tape head windings, together with a bias current for linearity. An external magnetic field, through which the magnetic tape passes, is developed across a narrow gap in the tape head. The resulting tape surface magnetization is approximately proportional to the signal current.

On playback, the recorded tape is passed over the tape head, causing the surface magnetic field to pass through the head core. A voltage is induced into the tape head windings which is proportional to the rate of change of this magnetic field. So the playback waveform is proportional to the rate of change of the recorded waveform. For a recorded sine wave $A\sin\omega t$ (A is amplitude, ω is angular frequency) the playback voltage is proportional to $A\omega\sin(\omega t+\pi/2)$. Thus, for the recording of a variable-frequency sine wave with constant recording current, the playback voltage will increase linearly with frequency. The phase response is not the one normally associated with this frequency response: the output voltage has a constant +90° phase shift which is independent of frequency, and is due to the $\pi/2$ term mentioned earlier.

The playback amplitude variation with frequency found in practice is shown in Fig. 2. In addition to the 6 dB/octave rise in the output with frequency, losses take their toll at high and low frequencies. At low frequencies the head-core magnetic flux no longer remains proportional to the surface magnetic flux of the tape, and the output voltage falls accordingly. At high frequencies, various losses contribute, the most significant being the gap effect. As the wavelength of the signal on the tape approaches the length of the playback head gap the output voltage falls – eventually to zero.

To obtain an overall flat amplitude/fre-

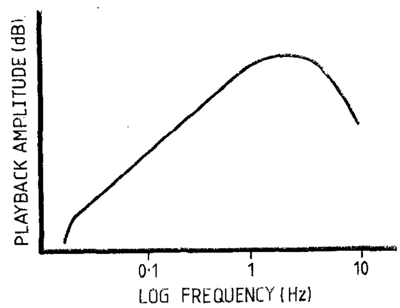


Fig. 2. *Amplitude/frequency response of playback head from constant-current recording.*

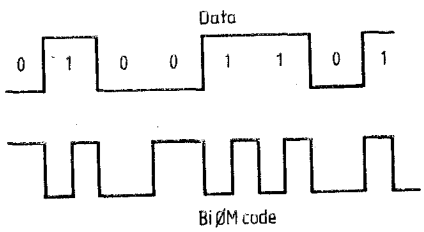


Fig. 3. *Biphase M coding.*

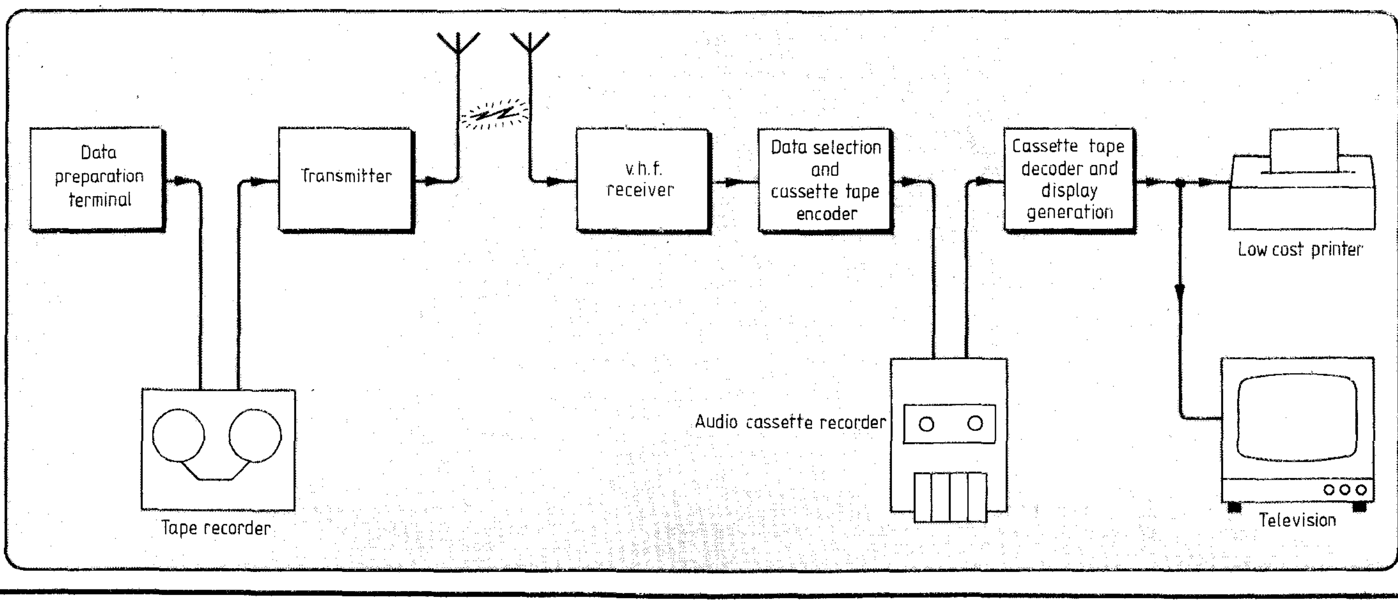


Fig. 1. *Radiotext system produced by Open University*

quency response, an amplitude equalization filter is normally included in the playback amplifier. It provides a 6 dB/octave fall in amplitude with frequency, up to the point where high-frequency losses take effect. The filter characteristic is then designed to give increasing amplitude with frequency to combat these losses. The frequency at which the equalization filter characteristic is changed is standardized at approximately 1350 Hz (equalization time constant 120μs) for ferric tapes and approximately 2250 Hz (equalization time constant 70μs) for chrome tapes.

The bandwidth of a low-cost audio cassette recorder is typically 150 Hz to 6 kHz, but can be reduced either temporarily or more permanently by two effects caused by deficiencies in the tape transport mechanics or in the tape itself. If contact between the tape surface and head is lost, a resulting output loss occurs. Such losses are commonly termed "drop-outs". The spacing loss formula

$$\text{loss in dB} \simeq 55\, d/\lambda$$

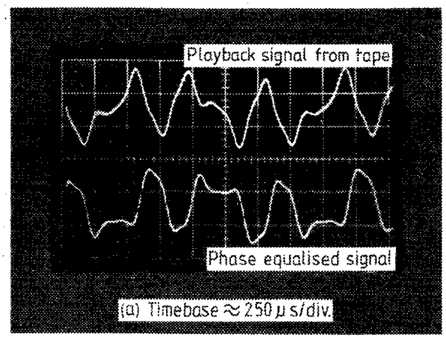
where d is separation (mm) and λ is wavelength of signal on tape (mm) shows the loss to be severe for even small separations and to be proportional to frequency. For example, at 5 kHz the wavelength of the recorded signal on tape is approximately 0.0095 mm. A head-to-tape separation of this distance will cause a 55 dB loss at this frequency. Surface imperfections in even good quality cassette tape still cause an almost continuous rapid variation in amplitude of the playback signal.

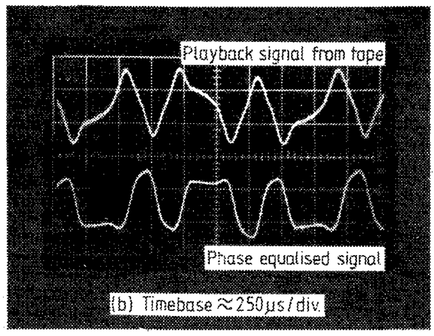
As the tape passes over the head, the longitudinal axis of the tape must be exactly perpendicular to the head gap, so that the gap lies across the tape. Any misalignment of the head has the effect of increasing the gap length, reducing the amplitude of high frequency signals resulting in loss of available signal bandwidth.

Channel code

The audio cassette recorder is essentially a band-pass channel which suffers amplitude instability, particularly at the higher frequencies. The channel code should therefore be d.c.-free and have a frequency spectrum, for the chosen data rate of 2400 baud, which lies at the lower end of the bandwidth available. A code which satisfies these requirements is Biphase M or Manchester coding[3] as shown in Fig. 3. A transition occurs in the code at every bit edge with an extra transition midway through the bit period to distinguish data 1 from data 0. Unlike the other well used biphase code Biphase L, Biphase M is unaffected by the signal inversion which occurs in some makes of audio cassette recorder and the lack of need for synchronization ensures that recovery is rapid should a signal drop-out occur.

For this application Biφ M is preferred to more efficient codes, because of the simplicity of implementation and its robustness in the presence of time jitter caused by short-term speed variations of the cassette tape.





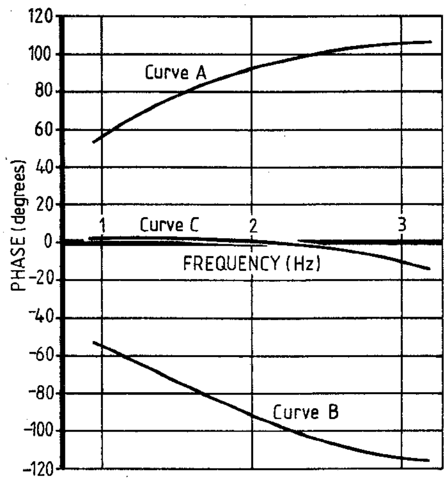


Fig. 4. *Phase response of typical audio cassette recorder is curve A; curve B is phase characteristic of all-pass filter and C is combined phase response of recorder and filter. Photos show effect on typical signals from two recorders with characteristics close to likely extremes.*

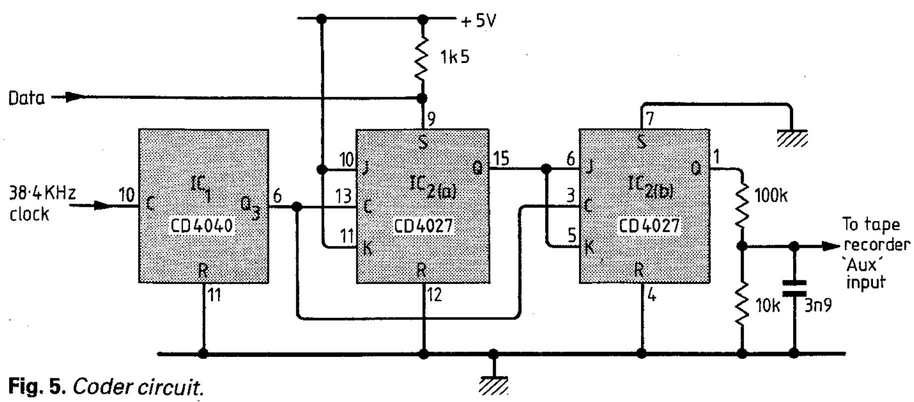


Fig. 5. *Coder circuit.*

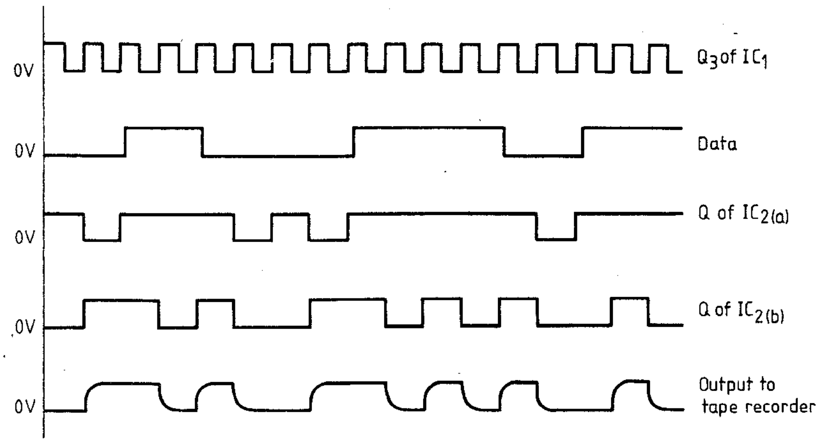


Fig. 6. *Waveforms in circuit of Fig. 5.*

Phase equalization

Because the human ear is fairly insensitive to the relative phases of the frequency components within a signal, no attempt is made by the manufacturers of low-cost audio cassette recorders to linearize the phase response. In order to preserve the waveform of the Biφ M coded signal, however, such phase linearity is essential and has to be provided externally.

The measurement of phase response is difficult to make because of the discontinuous nature of the record-playback process. The overall recorder response was estimated by measurement of record and playback amplifier phase responses and by measurement of the relative phase of signals at different frequencies recorded together. Additionally, the phase response of the recorder was modelled, the expected response to various waveforms calculated

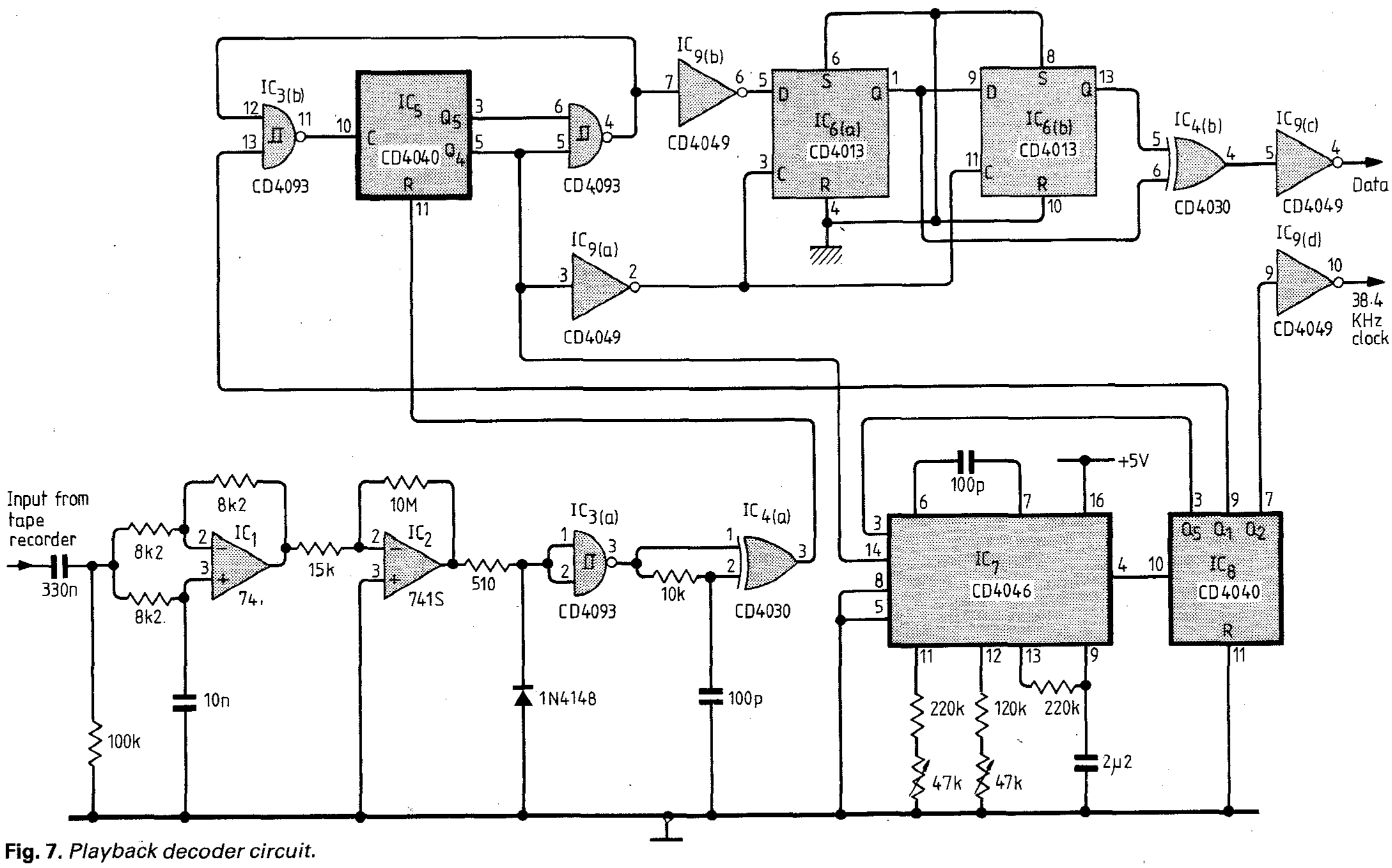


Fig. 7. *Playback decoder circuit.*

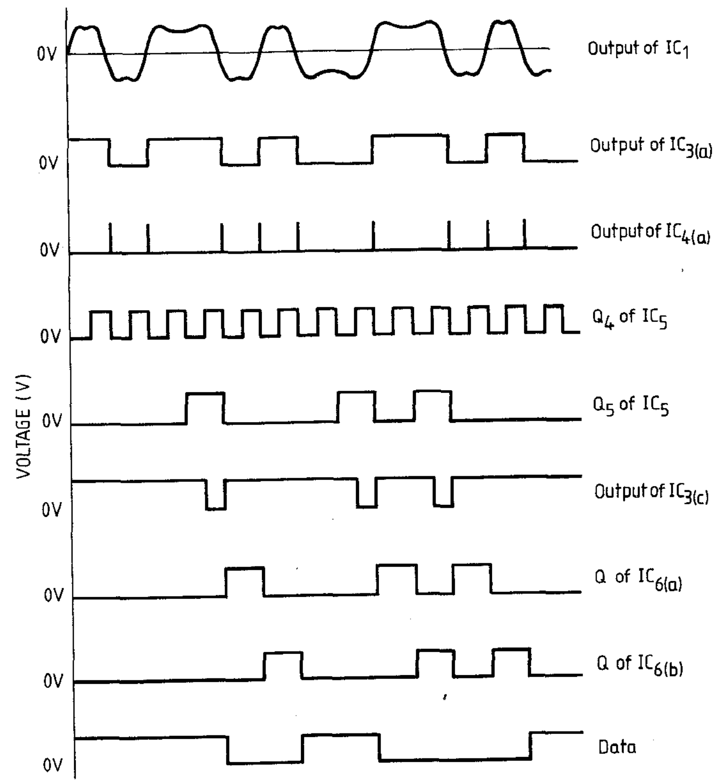


Fig. 8. *Waveforms in decoder.*

and comparison made between the calculated response and that found in practice.

A typical phase response, ignoring the time displacement between record and playback, is shown in Fig. 4 curve A. The response is essentially that of the +90° shift independent of frequency produced at the playback head with a lag caused by the low-pass characteristic of the amplitude equalization filter at lower frequencies and the lead caused by the high-pass characteristic of this filter at higher frequencies.

Transversal (tapped delay line) filters can be used for phase equalisation[4]. However, in the case of the audio cassette recorder, an active all-pass filter provides a simple, yet effective method of achieving the required phase linearity. The phase response of the filter used is shown in Fig. 4 curve B, and the overall response of the recorder and filter shown in Fig. 4 curve C. The photographs show the effect of the phase equalization on the output of two audio cassette recorders chosen because their characteristics are close to the extremes of performance likely to be encountered.

Coder

The circuit used to provide the Biϕ M coding is shown in Fig. 5. The data in serial format, together with a 16-times data-rate clock are provided, via an asynchronous communications interface adaptor. IC_1 divides the clock frequency to provide the 4.8 kHz clock required by IC_2. Flip flops $IC_{2(a)}$ and $IC_{2(b)}$ produce the coded data. The output of $IC_{2(b)}$ is reduced to a suitable level for application to the auxiliary input of the audio cassette recorder. The output is filtered (using a low-pass filter of approximate cut-off frequency 4.5 kHz) to avoid ringing which occurs with some record amplifiers. The waveforms applicable to various points within the coder are shown on Fig. 6.

Decoder

The decoder used is shown in Fig. 7. It relies for its operation on timing the interval between transitions of the playback signal. The timing is achieved by use of counter IC_5, which is provided with a clock frequency derived from the incoming signal and so accurate decoding is independent of tape speed.

IC_1 provides phase equalization of the playback signal, which is then squared by IC_2 and $IC_{3(a)}$. It is then passed to exclu-

Continued on page 62

Symmetrical-output dividers

Wide frequency-range dividers for odd and even-number division

by Gerard Girolami and **Philippe Bamberger**

These articles describe a technique for designing a symmetrical-output divider, operating over a wide frequency range with a choice of logic-selectable division ratios. Here, simple fixed and variable-ratio dividers are discussed to illustrate how logic i.cs are used to eliminate frequency-dependent components such as capacitors and monostables. Expansion of the programmable binary-input divider described will be discussed in a subsequent article, as will a similar circuit but with b.c.d. inputs.

Dividing a frequency by an odd number and obtaining a symmetrical output is not too great a problem, provided that the division ratio is constant. When a programmable divider is required, the problem is somewhat greater, especially if the circuit must be capable of operating over a wide range of frequencies. Circuit elements such as monostables, capacitors and resistors are useful when working with a fixed frequency or in a very narrow band, but for a wide-band, variable-ratio divider, another method of obtaining a symmetrical output must be found. Another drawback of most conventional dividers of this type is that they are difficult to cascade when high division ratios are required. This method uses only logic elements and provides a solution to the aforementioned problems.

Fixed-ratio dividers

Figure 1 shows the timing diagram for a symmetrical output divide-by-three circuit to illustrate two points;

– the output signal must change state on either the rising or falling edge of the input signal.

– to obtain a change of state in the middle of the output cycle the input signal must be symmetrical.

It was decided that the desired output could be obtained by modifying the clock signal of the counter.

A divide-by-15 circuit with its clock input modified as shown in Fig. 2 is the first example. Here a synchronous divider, the 74163, is fed with a clock signal modified by an exclusive-OR gate. Figure 3 shows the timing diagram for the modified clock input, H′, the input, H_0, and the output of the divider, from an initial value of five. It is obvious that at counts eight and zero the modified clock pulses are half a period shorter than the rest. If the A input of the exclusive-OR gate is connected to Q_C output of the divider, four shorter periods will occurr at zero, four, eight and twelve. As a

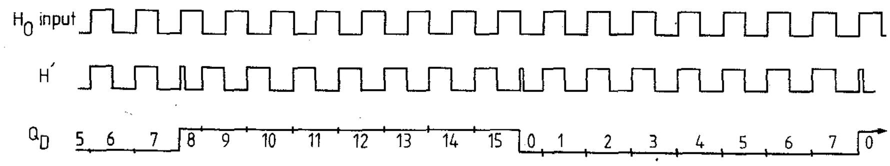
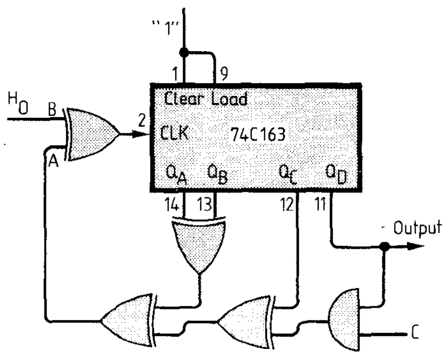


Fig. 3. Timing diagram of the divide-by-15 circuit with the modifications shown in Fig. 2.



$C = 0 \quad A = Q_A \oplus Q_B \oplus Q_C$: divide-by-10

$C = 1 \quad A = Q_A \oplus Q_B \oplus Q_C \oplus Q_D$: divide-by-11

***Fig. 4.** A logic selectable divide-by-10 (C-low) or divide-by-11 (C-high) circuit.*

Table 1: Requirements for division ratios from 1 to 16. Here, $H' = H_0 \oplus C$ applies where C is the combination. H' and H_0 are made clear in Fig. 2. When dividing by 2, 4 or 8, two combinations are possible.

Ratio	Combination	Output
16	0	Q_D
15	$(Q_C\oplus)Q_D$	Q_D
14	$(Q_B\oplus)Q_C$	Q_D
13	$Q_B\oplus(Q_C\oplus)Q_D$	Q_D
12	$(Q_A\oplus)\ Q_B$	Q_D
11	$Q_A\oplus Q_B\oplus(Q_C\oplus)Q_D$	Q_D
10	$Q_A\oplus(Q_B\oplus)Q_C$	Q_D
9	$Q_A\oplus(Q_C\oplus)Q_D$	Q_D
8	Q_A	Q_D
8	0	Q_C
7	$(Q_B\oplus)Q_C$	Q_C
6	$(Q_A\oplus)Q_B$	Q_C
5	$Q_A\oplus(Q_B\oplus)Q_C$	Q_C
4	Q_A	Q_C
4	0	Q_B
3	$(Q_A\oplus)Q_B$	Q_B
2	Q_A	Q_B
2	0	Q_A
1	Q_A	Q_A

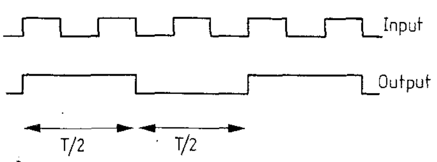


***Fig. 1.** Timing diagram for a symmetrical-output divide-by-three circuit. To obtain a symmetrical output, the input must be symmetrical.*

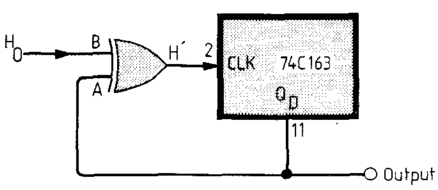


***Fig. 2.** How the divider's clock input is modified using an exclusive-OR gate to give a divide-by-15 circuit with symmetrical output.*

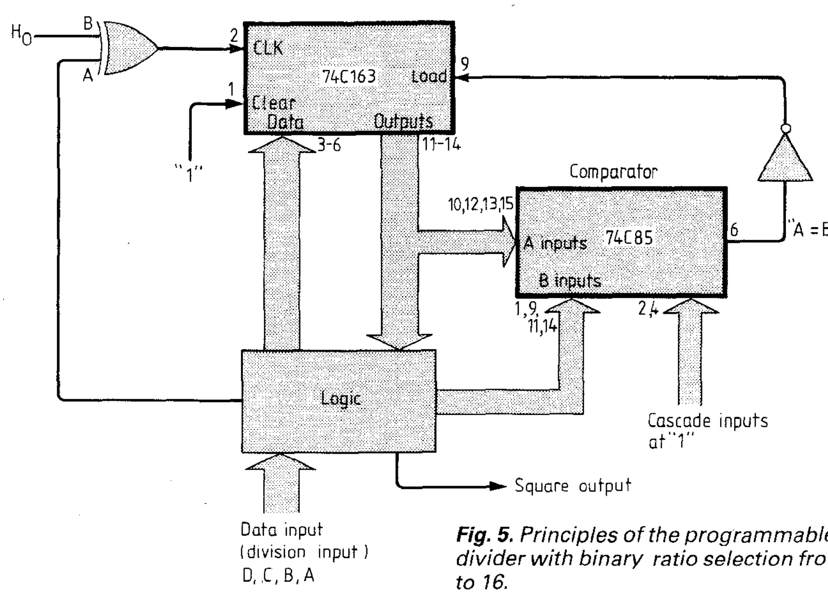


***Fig. 5.** Principles of the programmable divider with binary ratio selection from 1 to 16.*

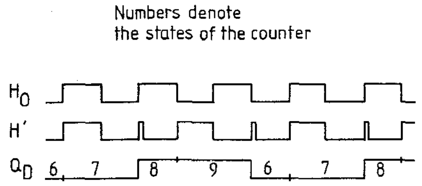


Fig. 6. Timing diagram of the programmable divider for divide-by-three to illustrate dividing by an odd number.

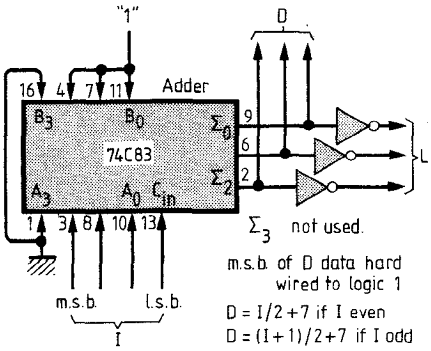


Fig. 7. Applying equations 2 and 3. The most significant bit of the data lines is always logic 1 and should be hard wired. Sum output 3 of the 4-bit full adder is not used here.

result, a divide-by-14 circuit is obtained. In fact, all the ratios shown in Table 1 can be obtained using different combinations of divider outputs at the A input of the OR gate.

The terms in parentheses in the table, although not strictly necessary here, do not modify the result and can be used in designing a variable-modulus counter such as that shown in Fig. 4, where the circuit divides by ten or eleven. The above method is useful when the dividing ratio is constant or when the number of ratios required is limited. But if the modulus has to be changed frequently it may be more practical to realize a programmable divider as described in the following section.

Programmable divider

The following describes a programmable divider with binary inputs from 1 to 16. A 74163 is still used but its cycle is modified to force it to oscillate around counts seven and eight. Consequently a square output is obtained at the Q_D output. This can be done with a circuit built using the principles shown in Fig. 5, where the load input of the 74163 is connected to the inverted "A=B" output of a 4-bit magnitude comparator. The comparator's inputs are the outputs of the counter and certain logical functions of the command bits A, B, C and D.

Suppose that a divide-by-four circuit is

Table 2: With the divide-by-4 circuit described, output symmetry at the 7 to 8 count transition is maintained by loading 6, counting 6, 7, 8 and 9, and then loading 6 again.

	Q_D	Q_C	Q_B	Q_A	Count
On 9,	0	1	1	0	6
reload 6	0	1	1	1	7
	1	0	0	0	8
	1	0	0	1	9

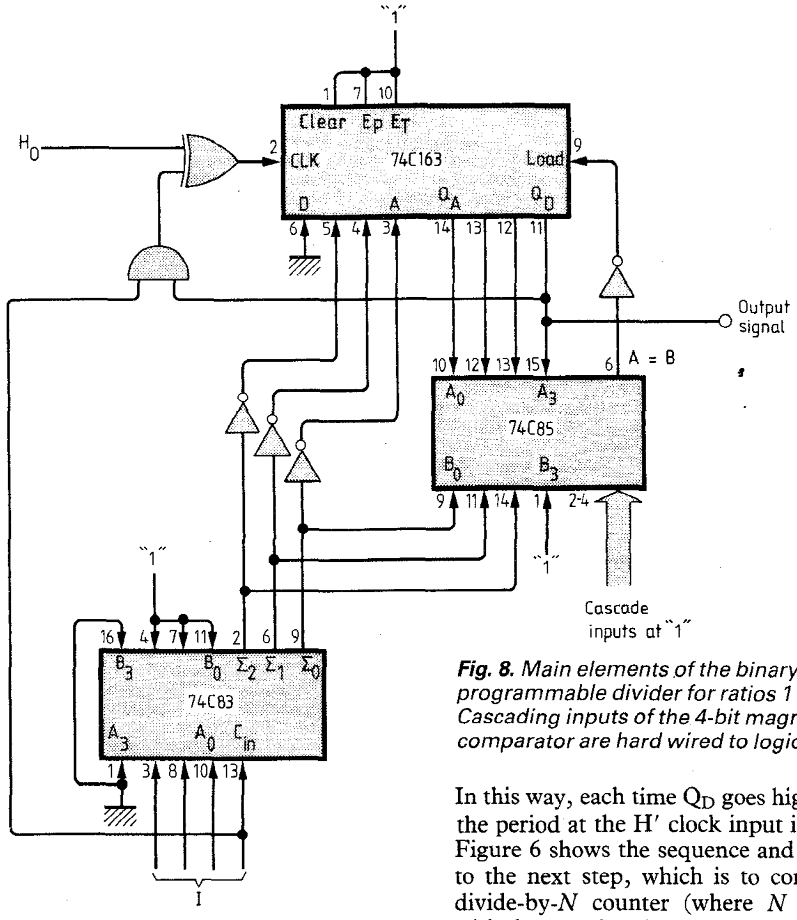


Fig. 8. Main elements of the binary programmable divider for ratios 1 to 16. Cascading inputs of the 4-bit magnitude comparator are hard wired to logic 1.

Table 3: Input, load and detect values for the 1 to 16 programmable divider.

Ratio	Load	Value	Detect	Value	Input
1	0111	7	1000	8	0001
2	0111	7	1000	8	0010
3	0110	6	1001	9	0011
4	0110	6	1001	9	0100
5	0101	5	1010	10	0101
6	0101	5	1010	10	0110
7	0100	4	1011	11	0111
8	0100	4	1011	11	1000
9	0011	3	1100	12	1001
10	0011	3	1100	12	1010
11	0010	2	1101	13	1011
12	0010	2	1101	13	1100
13	0001	1	1110	14	1101
14	0001	1	1110	14	1110
15	0000	0	1111	15	1110
16	0000	0	1111	15	10000

required (four is an even number but nevertheless a good starting point). If one pulse is saved to load the counter, three periods remain for further use. The only way to obtain symmetry around the seven-/eight transition is to load the six, count seven, eight, nine and then load the six again. This gives a symmetrical divide-by-four circuit, the sequence of which is shown in Table 2.

Suppose that a divide-by-three circuit is now required. Some periods will have to be shortened again as they were for the fixed ratio divider. This can easily be done by connecting the A input of the exclusive-OR gate to the Q_D output of the counter.

In this way, each time Q_D goes high or low the period at the H' clock input is halved. Figure 6 shows the sequence and leads on to the next step, which is to construct a divide-by-N counter (where N is even) with shortened cycles.

Table 3 shows what data must be loaded into the counter and what the comparator must detect in order to obtain the given ratios. Even though five digits are required to write 16, it is still possible to divide by 16, as a combination with four digits at "0" is spare, i.e. the four zeros are not used for any of the previous division ratios. The following additional information can also be obtained from Table 1. Firstly, the load and detect data are always complementary so,

$$L+D=2^4-1=15 \quad (1)$$

Secondly, if the input is even,

$$D-I/2=7 \quad (2)$$

and if the input is odd,

$$D-(I+1)/2=7. \quad (3)$$

Figure 7 shows how this can be applied, and the complete divider is shown in Fig. 8.

Again using Table 3 it is possible to derive logic relationships between the command inputs (I), and the load data, and one can verify that, where $\oplus$= exclusive OR,

$L_0=I_0\oplus I_1$
$L_1=(I_0+I_1)\oplus I_2$
$L_2=(I_0+I_1+I_2)\oplus I_3$
$L_3=0$

and, of course,

$D=\overline{L}$

This method is the same as used in the first example.

To be continued

Economical Z80 development system

Software interface links Nascom and Softy

by G. Winstanley, B.Sc., **and S. R. W. Grainger,** B.Sc.

The successful design of microprocessor-based equipment requires a flexible development system for preparing, debugging and modifying software. This design uses a software interface to combine a Nascom microcomputer with a Softy e.p.r.o.m. programmer/emulator, and offers important facilities such as object-code to tape, memory and full assembled listing to printer, memory-mapped output to a tv, and a range of editing functions.

This interface enables Z80 machine-code programmes to be developed on an expanded Nascom microcomputer, via the resident assembler, with the subsequent object-code transferred to a device capable of r.o.m. emulation and eventual programming. The hardware comprises a Nascom microcomputer connected to a standard Softy which is used as an e.p.r.o.m. programmer and as an emulation device with 1K of memory. Some software has been developed for the Softy which contains a INS8060 microprocessor to enable data transfer.

The Nascom assembler (Zeap II) is capable of placing assembled object-code directly in the locations specified by the program origin statement, and is ideal for programmes developed for the Nascom, with error traps included to prevent overwriting of valuable areas of memory. However, if the system is to develop r.o.m.-based programmes for use with a separate unit, usually with an origin at address 0HH, two basic methods of assembly and transferral are possible. The first involves obtaining and transferring each byte of data at the moment of assembly, which has the advantage of immediate transferral with no intermediate steps, but has the disadvantage that any alterations have to be made to the assembler program itself. The method chosen for this development system relies on a feature of the assembler which allows object-code to be placed in a different area of r.a.m. An area of memory, 1000 Hex to 13FFH, is allocated for dumping assembled object-code. Only 1K bytes of memory are allowed because the Softy is limited by its user r.a.m., but this is adequate for most assembler programmes. The Softy is capable of emulating 2K-byte e.p.r.o.ms, but in such a case, 1K bytes must be resident in e.p.r.o.m. When the data has been loaded into the correct areas of memory at the end of assembly, transferral from Nascom to Softy can take place.

Provision has been made in the Softy firmware for programmes to be run from an e.p.r.o.m. placed in the programming socket. Because Softy is an intelligent device, the microprocessor can be used to accept data and place this in the user r.a.m. Although the two ports possessed by the Softy r.a.m. i/o are used for keyboard scanning functions, they are brought out to an edge connector so one port can be used for 8-bit parallel data transferral, and the second can be used for handshaking purposes. Data dumped into the 1K block of Softy r.a.m. can be programmed into an e.p.r.o.m. using the burn routine, or used for emulation. The complete operation is controlled by the Nascom system, with the Softy providing handshake pulses. At the end of data transferral the Softy c.p.u. is reset by the Nascom and is ready for independent operation. Because the interface involves asynchronous data flow, the system relies on handshaking for successful transferral. Necessary control lines include Data ready, Byte transferred, and All data processed. Nascom 1 has an uncommitted programmable i/o device with two ports, and in this system one is committed to the output of each 8-bit data byte, while two control lines are taken from spare bits of otherwise committed i/o ports. This uses the ports efficiently and allows port B of the p.i.o. to be used for other purposes. A block diagram of the interface is shown in Fig. 1. Port 0 of the Nascom is assigned to scan the keyboard and sense, with bit 7 input and bit 2 output uncommitted.

Although the Softy ports are committed to scanning the keypad, they can be accessed in parallel with the keyboard function. Port B is monitor programmed and wired as an 8-bit input, and therefore the logical choice for the 8-bit parallel input. A few port A lines are used for internal system operation, but allow some bits to be used for handshaking purposes.

As mentioned earlier, programmes can be run from an e.p.r.o.m. in the program socket, which allows the full 1K of user r.a.m. to be accessed for storing external data. Thus, a transfer program can be programmed by the Softy into an e.p.r.o.m. and the existing monitor used to initiate this program. With the data handshaking system between Nascom and Softy it is necessary for the Nascom to reset Softy at the end of the data transfer. Ideally the Softy interrupt should be used to transfer back to the monitor, but this is used for video control functions. Because Softy and Nascom are based on different microprocessors with dissimilar clock frequencies and instruction cycle times, reliable operation depends on handshaking to overcome these problems and ensure efficient data transfer. For this reason a twin software system has been developed in which the Softy is initialised and waits for a Go command from the computer. The data downloading software is illustrated in Fig. 2. With Softy waiting in a continuous loop for the start command, data transferral can only be accomplished at a time determined by the user via the Nascom and its transfer program in e.p.r.o.m. After initialisation of the memory pointer and byte counter, the first byte of data is latched onto the parallel output port (A of p.i.o.). This is performed before the Data available command so that program timing differences

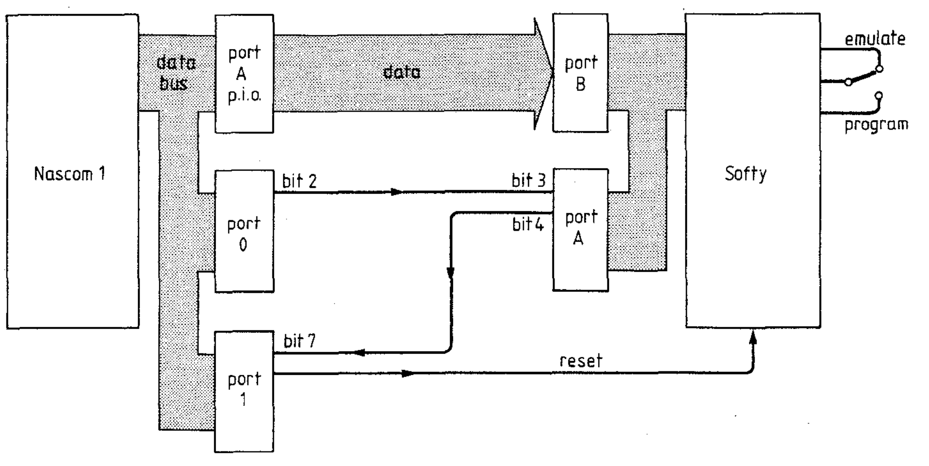


Fig. 1. Nascom-Softy interface.

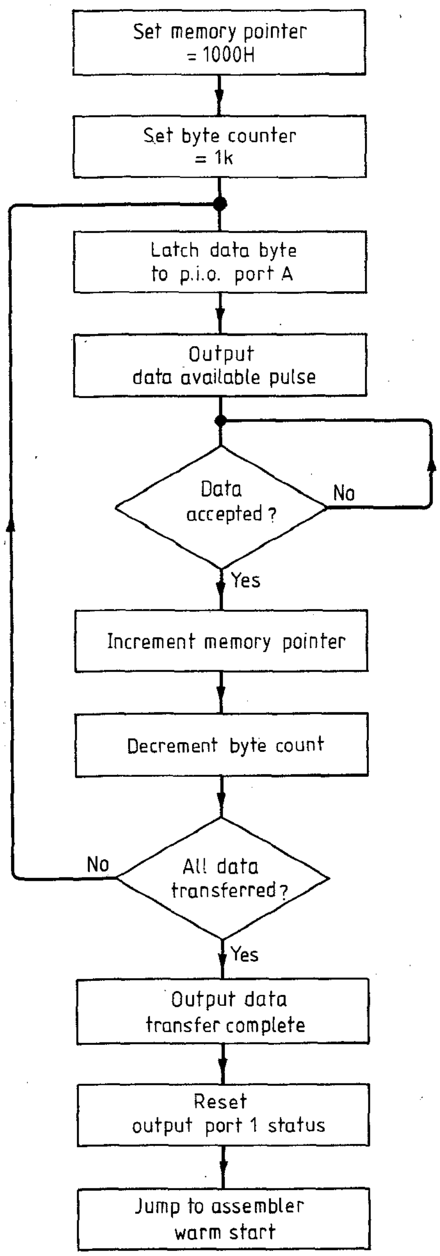


Fig. 2. Flowchart for the Nascom software.

Table 1. Nascom transfer program.

```
                ; ** NASCOM DATA TRANSFER PROGRAMME **
                ;
24C8 DØØ3       WRMS  EQU  ØDØØ3H     ;ZEAP WARM START
24C8 1ØØØ       LOCA  EQU  1ØØØH      ;START LOCATION
24C8 Ø4ØØ       BTYE  EQU  Ø4ØØH      ;BYTE COUNT
                ;
ØBØØ                  ORG  ØBØØH
                ;
ØBØØ 3EØF       STRR  LD   A,ØFH
ØBØ2 D3Ø6             OUT  (Ø6H),A    ;CONTROL PORT PIO
ØBØ4 21ØØ1Ø           LD   HL LOCA    ;START LOCATION
ØBØ7 Ø1ØØØ4           LD   BC BTYE    ;BYTE COUNT
ØBØA 16Ø5       DELL  LD   D,Ø5H      ;TIME CONSTANT
ØBØC 7E               LD   A,(HL)     ;GET 1ST BYTE
ØBØD D3Ø4             OUT  (Ø4H),A    ;Ø4=DATA PORT PIO
ØBØF 3EØ4             LD   A,Ø4H      ;BIT 2
ØB11 D3ØØ             OUT  (ØØH),A    ;PORT Ø
ØB13 15         DCRR  DEC  D          ;DELAY LOOP
ØB14 2ØFD             JR   NZ DCRR
ØB16 E6FB             AND  ØFBH       ;RESET PULSE
ØB18 D3ØØ             OUT  (ØØH),A
ØB1A DBØØ       TSTT  IN   A,(ØØH)
ØB1C CB7F             BIT  7,A        ;DATA ACCEPTED ?
ØB1E 28FA             JR   Z TSTT     ;NO !,WAIT
ØB2Ø 23               INC  HL         ;FOR NEXT BYTE
ØB21 ØB               DEC  BC         ;BYTE COUNT
ØB22 79               LD   A,C
ØB23 A7               AND  A          ;CHECK FOR Ø
ØB24 2ØØ4             JR   NZ RSTT    ;NOT FINISHED
ØB26 78               LD   A,B
ØB27 A7               AND  A          ;CHECK FOR Ø
ØB28 28Ø8             JR   Z ENDI     ;ALL DONE.END
ØB2A DBØØ       RSTT  IN   A,(ØØH)
ØB2C CB7F             BIT  7,A        ;TRAILING EDGE ?
ØB2E 28DA             JR   Z DELL     ;RESET DELAY
ØB3Ø 18F8             JR   RSTT
ØB32 3E2Ø       ENDI  LD   A,2ØH      ;BIT 5
ØB34 D3ØØ             OUT  (ØØH),A
ØB36 3EFF             LD   A,ØFFH     ;ALL BITS HIGH
ØB38 D3Ø4             OUT  (Ø4H),A    ;ON PIO
ØB3A C3Ø3DØ           JP   WRMS
```

between the two systems can be overcome, i.e. data is available for transfer at the start and during the handshaking pulse. After data has been latched, the transfer command is given. A pulse of approximately 35μs was chosen because Softy has a maximum response time of about 30μs governed by Load and Store instructions. At the trailing edge of this pulse the Nascom waits in a loop for a signal indicating correct single-byte transferral and, in this case, the computer is instructed to recognise and wait for the leading and trailing edge of a pulse. This software allows for the slower operating speed of Softy and ensures that data is not latched and a handshaking output not given during the time taken by Softy to deliver its own handshaking pulse.

After successfully completing each data-byte move operation, the Nascom program services its memory location and byte-counter registers ready for the reset data-byte transferral, and checks the status of the latter for final block transfer completion. The result of this check directs the program to another identical operation or to the final part of the program. This is important because data is transferred from the Nascom p.i.o. to port B on the Softy and a physical connection therefore exists between the two. Problems may arise because port B is also used for keyboard entry and if any bits of the port are maintained at 0, the Softy keyboard is disabled. Therefore, any device connected to the Softy in this way must be either in a high impedance state or at 1. For this reason the final part of the Nascom program includes a routine to set all bits of port A p.i.o. to 1 and a pulse is delivered to the Softy, similar to the data-available signal, as a reset command.

When used in conjunction with the Nascom Zeap assembler, the final instruction of the transfer program is very important. An unconditional jump to the Zeap start location would require subsequent keyboard commands to reload buffers, set object-code-to-r.a.m. option, and a r.a.m. location origin command. However, a warm start does exist (location D003 H) and an unconditional jump to this location on completion of data transfer ensures correct buffer and assembler option status. In fact, transfer-program completion is indicated by the unusual return to assembler readout, and at this point full keyboard control is available for assembler program modification and eventual re-assembly. The Nascom transfer program listing is given in Table 1 and a flow diagram for the Softy software is shown in Fig. 3. After initialisation of the ports and memory pointer, the program waits for the trailing edge of a data transfer pulse. When this pulse is received, data is transferred from port B to the user r.a.m. starting at location 0C00H. After the data is stored, a handshake pulse is returned to the Nascom and the program returns to the data-transfer wait loop for the Nascom generated reset pulse. The speed of data transfer is therefore optimised.

A complete listing of the Softy program is given in Table 2. This is stored in a 2708 e.p.r.o.m. and is run from the programming socket. The program relinquishes control of the Softy after a reset pulse from the Nascom at the end of the transfer or by a manual pushbutton. It would have been desirable to use the 8060 interrupt line for this purpose, but it is already used in the internal operation of Softy.

Although the programs have been developed for use in conjunction with the assembler mentioned, programs developed in long-hand can be downloaded in the same fashion and tested by running or

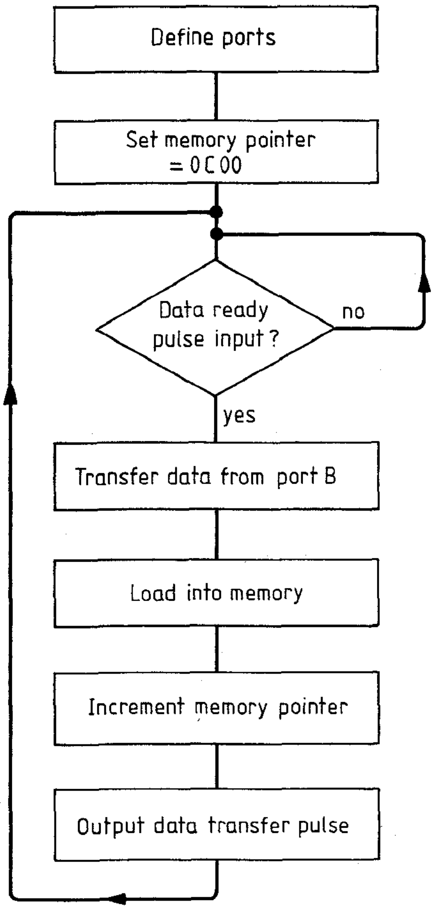


Fig. 3. Flowchart for the Softy software.

Table 2. Softy software, run from a 2708 in the programming socket.

```
0000    08       NOP
0001    C4 07    LDI
0003    36       XP2(H)
0004    C4 80    LDI
0006    32       XP2(L)    ;P2=0780
0007    C4 00    LDI
0009    CA 23    ST        ;PORT B OUTPUT
000B    C4 F0    LDI
000D    CA 22    ST        ;PORT A FUNCTION
000F    C4 0C    LDI
0011    37       P3(H)
0012    C4 00    LDI
0014    33       XP3(L)    ;P3=RAM ORIGIN
0015    C2 03    LD        :LOOP
0017    94 FC    JP        ;WAIT FOR PULSE
0019    C2 03    LD
001B    9C FC    JNZ       ;LOOK FOR EDGE
001D    C2 21    LD        ;INPUT DATA
001F    CF 01    ST@+1     ;STORE,INC. P3
0021    CA 14    ST
0023    CA 04    ST        ;OUTPUT O.K.
0025    90 EE    JMP       ;JUMP TO LOOP
OK
```

single-stepping at the transfer location if designed to be relocatable. This operation is useful for debugging subroutines prior to final downloading in an external system. Both memory pointer and byte counter could be made variable for programs greater or smaller than 1K, and monitor subroutines could prompt the user for an input of the two variables. At present this has not been included because, with the options available on the assembler, object-code can simply be dumped in the same safe area of memory on every occasion. However, such an alteration may be useful in cases where software or firmware must be copied from elsewhere in the addressable memory field.

A breakpoint function can be added if the system under development has numerical readout facilities. The prototype has a 3-channel, 4-digit display, and a subroutine has been added to display the main register set and contents of vital memory locations, which are accessed by single-stepping a dedicated hex keyboard. This facility is useful if routines need to be tested on the prototype without alternative debugging facilities. The small subroutine indicates the origin of the data and the register or memory contents. Also, break points can be included anywhere in the assembly language program.

The transfer routine in Fig. 2 is completely relocatable and can therefore be operated in any convenient area of memory as firmware. It can also be used in any Z80 based microcomputer, provided the port locations are re-assigned accordingly. Data is presently downloaded via a Z80 p.i.o., and the early part of Table 1 is responsible for its initialisation. A system using a conventional input port for this function would omit these instructions.

The system, which has been used to develop, document and debug a complex instrument, can be expanded to include automatic initialisation and bidirectional data flow which is useful for data verification. However, such modifications will need extra handshaking lines. The Nascom monitor subroutines can be used for prompts and indication of correct operation. The Softy routine is completely functional and there is little to be gained by modifying it, but the host computer transfer routines could become an integral conditional output of the resident assembler.

Literature Received

Leaflet is available from Astralux Dynamics on the 400 series of **dry-reed relays**, which come in a very wide range of styles, contact configurations and coil characteristics. The relays are PO, DEF 05-21 and BS 9000 approved. Copies of the leaflet from Astralux Dynamics Ltd, Red Barn Road, Brightlingsea, Colchester CO7 0SW. WW401

Active and passive components, hardware and measuring instruments from sixteen different makers are described, with their prices, in a 152 page catalogue from Abacus Electronics PLC, Kennet House, Pembroke Road, Newbury, Berks. RG13 1BX. WW402

Enclosures — a term which covers everything from 10cm square plastic boxes to large control desks — are made by Sarel Electric, of Cosgrove Way, Luton, Beds., who can supply a catalogue. WW403

Catalogue of scientific and technical **books** published by Adam Hilger and the Institute of Physics can be had from Booksales Department, The Institute of Physics, Techno House, Redcliffe Way, Bristol BS1 6NX. WW404

152 page catalogue from Watkins-Johnson lists the complete range of **solid-state amplifiers**, with selection charts, a glossary and some applications information. Watkins-Johnson International, Dedworth Road, Oakley Green, Windsor, Berks. SL4 4LH. WW405

Illustrated price list of **measuring instruments** for 1981/2 is obtainable on request from Bach-Simpson (UK) Ltd, Trenant Estate, Wadebridge, Cornwall PL27 6HD. WW406

Measuring instruments of various kinds made by the Austrian firm of NORMA are described in a new catalogue, which can be obtained from the UK agent, Cropico Ltd, Hampton Road, Croydon CR9 2RU. WW407

New catalogue of **microwave instruments** and components from Marconi Instruments, including the X-band signal generator 6812 and an adaptor to enable bus control of non-bus programmable instruments, is now available from MI Microwave Products Division, PO Box 10, Gunnels Wood Road, Stevenage, Herts. SG1 2AU. WW408

Colour brochure from Studer illustrates the 169/269/369 range of **mixing consoles** for use in small studios or o.b. vans, being designed to a compact format. The brochure describes the units available for the consoles, giving block diagrams and full specifications. F. W. O. Bauch Ltd, 49 Theobald Street, Borehamwood, Herts. WD6 4RZ. WW409

Various types of moving-coil, moving-iron and electronic panel **meters**, measuring instruments and special-purpose meters are made by Anders, who describe them fully in a new catalogue. Anders Electronics Ltd, 48-56 Bayham Place, London NW1 0EU. WW410

Designing with microprocessors

12 – Hardware for direct memory access systems

by **D. Zissos** assisted by **Glen Stone**
Department of Computer Science, University of Calgary, Canada

Direct memory access (d.m.a.) systems allow data to be transferred directly between a peripheral and the main memory in microprocessor-based systems. An outline of this technique was given in the September 1981 issue and the authors now go on to look at the basic hardware components of d.m.a. systems, describing their function and operation.

Direct memory access systems, as test-and-skip and interrupt systems, can be implemented using either programmable chips or dedicated logic. Although our design procedures accommodate both, we shall concentrate on systems using dedicated logic. The reason for this is that such systems are more easily understood and simpler to implement. Using programmable chips is simply the next step.

In the previous article, in the September 1981 issue, we explained the d.m.a. concept and described the basic d.m.a. configuration and its step-by-step operation. For ease of reference we reproduce the (simplified) block diagram of the d.m.a. configuration in Fig. 1. Briefly, its operation is as follows. When the d.m.a. controller has been initialized by the programmer, it turns signal E on, which enables the peripheral interface. When enabled, the peripheral interface requests the microprocessor to go on hold, whenever it recognizes that the peripheral is ready to communicate with the memory. When the microprocessor goes on hold, it pulls line HLDA high. Signal HLDA goes low when the microprocessor comes out of the HOLD state. That is, in the case of cycle-steal systems, the HLDA line is pulsed during each cycle steal. These pulses are used to decrement the word count ($n := n - 1$). When $n = 0$, indicating that the last byte has been transferred, the d.m.a. controller de-activates the peripheral interface and generates the end-of-transfer signal, ϵ.

A more detailed block diagram showing the main hardware components of a d.m.a. system is shown in Fig. 2. They are

1. An address decoder,
2. A d.m.a. controller,
3. Cycle steal logic, and
4. An interface, as shown in Fig. 2.

A detailed description of each of the four components is given next.

The address decoder is a standard i.c. chip, which in conjunction with signal OUT, allows the programmer to send to the d.m.a. controller the starting address, the block length, the direction of transfer and the 'go' command. As we have already explained, the OUT and address signals are generated during the execution of i/o instructions. From this point of view, the d.m.a. controller appears to the microprocessor as a peripheral that can be accessed with i/o instructions.

The d.m.a. controller consists of two counters connected in cascade, two flip-flops and a few gates, as shown in Fig. 3. The initializing information, comprising the initial address, the block length, the direction of transfer and the 'go' command, is loaded in the following manner.

The programmer moves into the accumulator the initial memory address and executes an i/o out instructions with address A_p. This generates an i/o pulse on the OUT terminal in Fig. 3, which is routed by address signal A_p to the parallel-load line of the two counters. This transfers the contents of the accumulator (starting address) into the first counter. At the same time, because the two counters are connected in cascade, the contents of the first counter are pushed into the second counter. The programmer then moves into the accumulator the block length and executes the same i/o instruction. This causes the initial address (stored in the first counter) to be pushed into the second counter, and the value of the block length (held in the accumulator) to be loaded into the first counter. Next the programmer executes another i/o instruction with address A_q if data is to be read from memory, and with address A_r if data

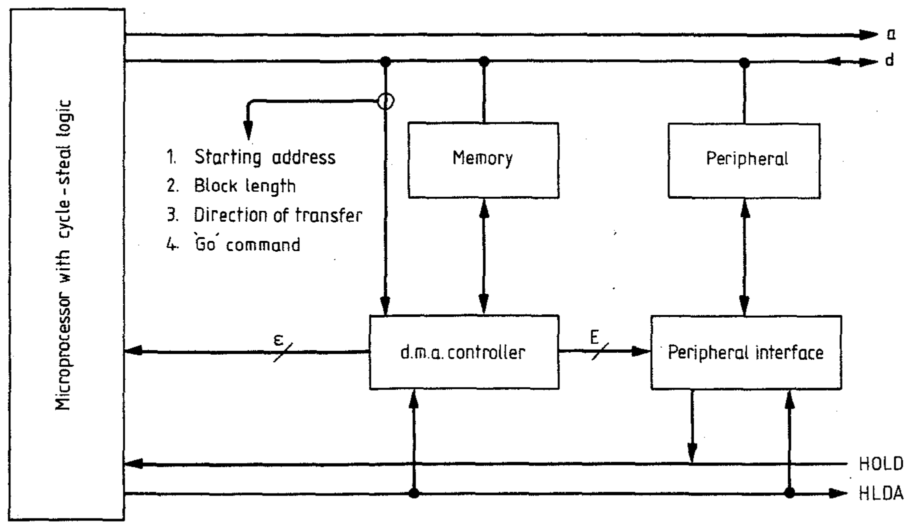


Fig. 1. *Simplified form of a d.m.a. system.*

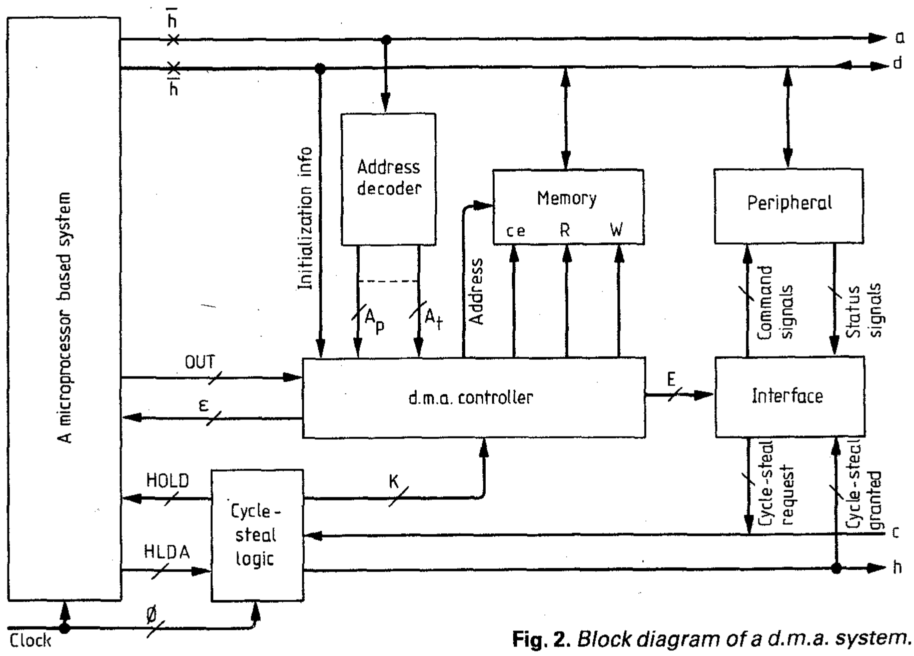


Fig. 2. *Block diagram of a d.m.a. system.*

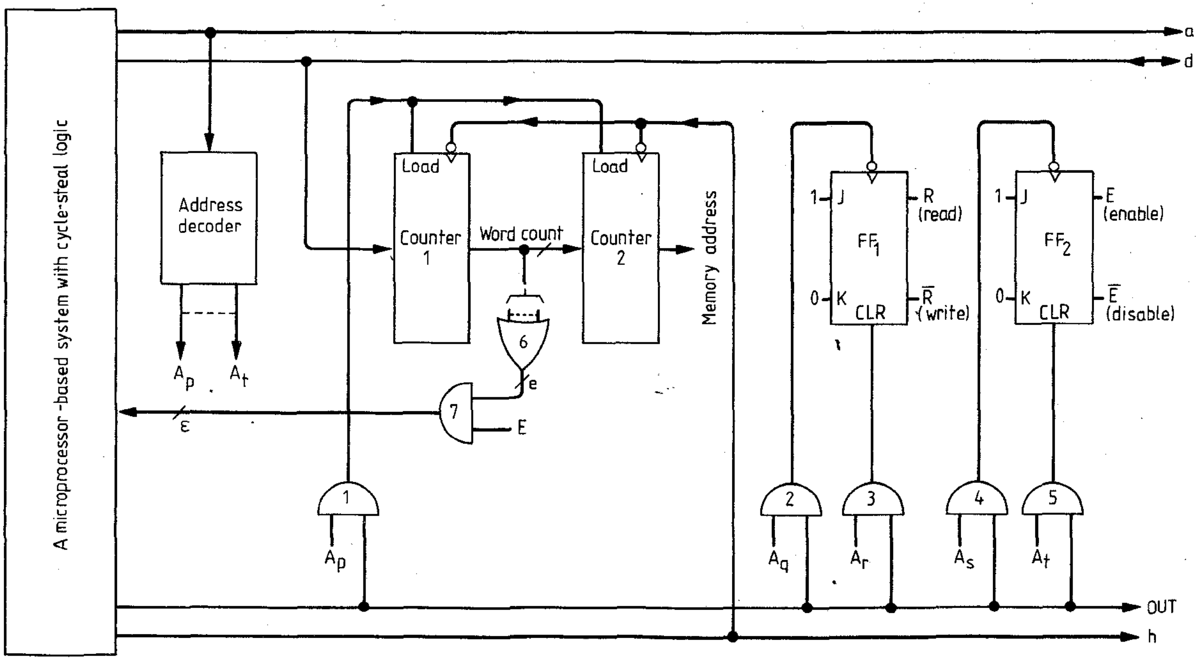


Fig. 3. *Logic of a d.m.a. controller.*

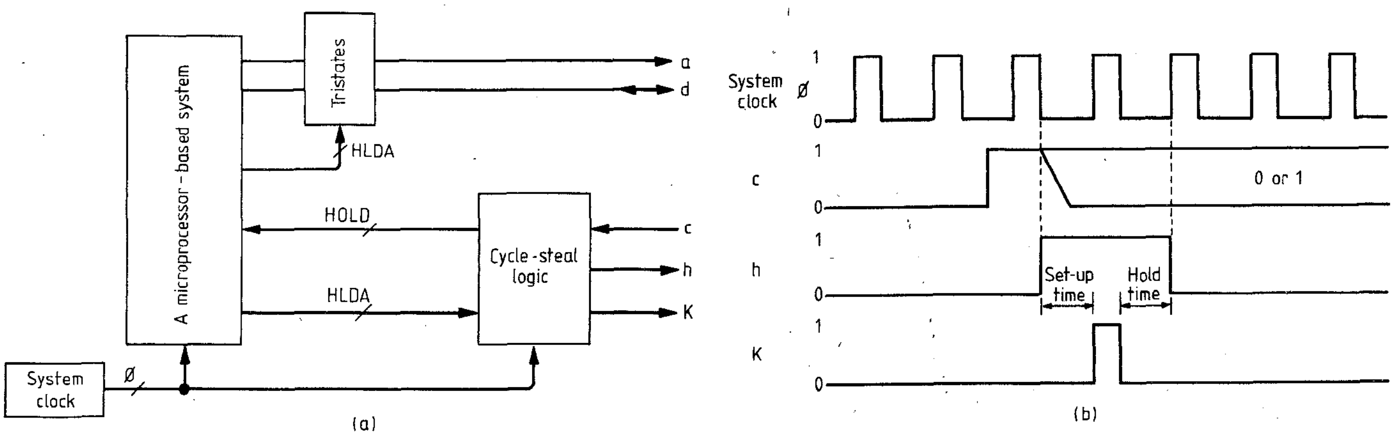
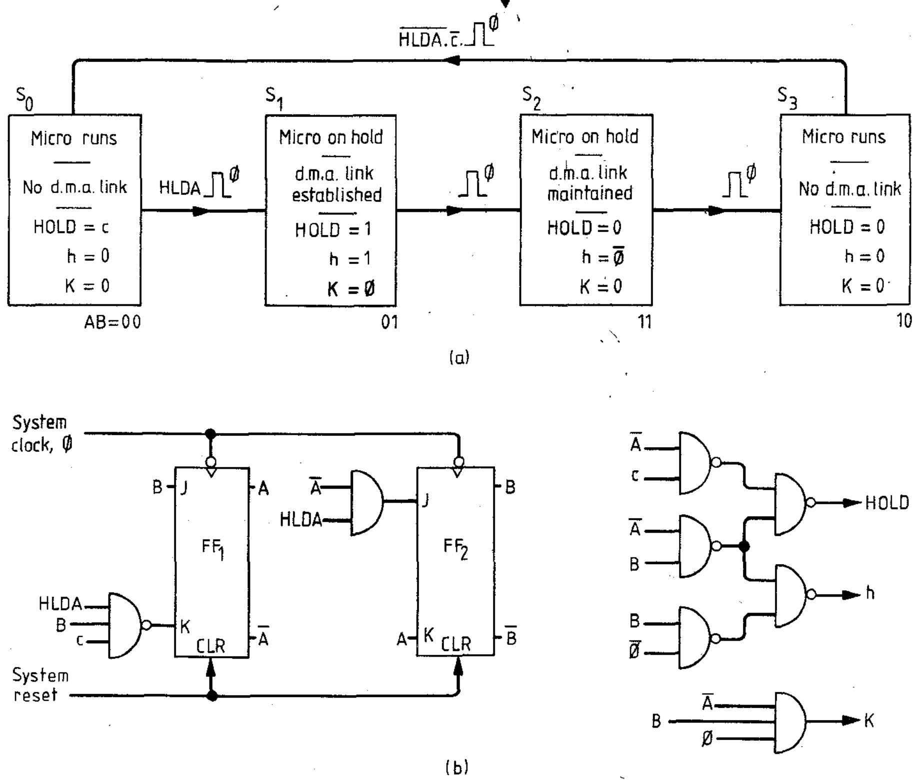


Fig. 4(a). *Block diagram of the cycle-steal logic. (b) Relative timing of cycle-steal signals.*

Fig. 5(a). *State diagram of the cycle-steal logic in Fig. 4(a). (b) Circuit implementation of the cycle-steal logic.*



is to be written into memory. Fig. 3 shows that in the first case FF1, the read/write flip-flop, is set, whereas in the second case it is reset. The 'go' command, which also takes the form of an i/o instruction, with address A_s in our case, sets FF2, the enable/disable flip-flop. Its output E, when equal to 1, activates the peripheral interface, and when equal to 0 deactivates it. The flip-flop is reset with an i/o instruction and address A_t. At this point the reader should recall that all interfaces in a system must be provided with an enable/disable flip-flop, to allow the user to isolate individual system components by resetting the flip-flop for such purposes as maintenance, trouble shooting, dynamic responses and so on.

End-of-transfer signal ε is generated by ANDing enable signal E with the output of the NOR gate, e, which goes high (e := 1) when the word count becomes zero; that is immediately the last piece of information has been transferred in or out of memory. Signal E is software-cleared by executing an i/o instruction with address A_t, which resets FF2.

Cycle-steal logic. As we have already explained, each time the main memory in a microprocessor-based system is to be accessed, the HOLD signal in Fig. 2 must be pulled and maintained high until direct access to the memory is no longer required. In the case of cycle stealing, direct access to the memory is required for one memory cycle, which is the time needed for an item of information to be read from it or written into it. For this purpose we need a logic circuit that will generate a HOLD signal, when access is memory is required, and terminate it when the microprocessor has been held off for one memory cycle.

In our case, cycle-stealing will be initiated by pulling line c in Fig. 2 high. When the microprocessor chip goes on hold, our cycle-steal logic generates two signals, h and k. Signal h indicates to the rest of the system that the microprocessor has gone on hold for one memory cycle, and signal k is a pulse to be used by the d.m.a. controller during the memory cycle for reading or writing a byte into the memory chip. The block diagram of the cycle-steal logic is shown in Fig. 4(a) and the timing of its signals in Fig. 4(b). The relative timing of cycle-stealing signals has been defined arbitrarily, although not unrealistically, and can be easily modified to meet specific restrictions, such as setup and hold times.

The design and implementation of cycle-steal logic is straightforward, as we shall illustrate by means of the following problem.

Problem

Design and implement the cycle-steal logic, whose block diagram is shown in Fig. 4(a). The timing of the cycle-steal signals in relation to the system clock is shown in Fig. 4(b).

Solution

Step 1: external (i/o) characteristics. As defined.
Step 2: internal characteristics. A suitable internal state diagram is shown in Fig. 5(a). Its operation is self explanatory.
Step 3: state reduction. Omitted for clarity of design. The 2^n rule, explained on page 11 of 'Problems and Solutions in Logic Design' by D. Zissos (Oxford University Press, 2nd edition 1979) in our case is met.
Step 4: circuit implementation. By direct reference to our state diagram, we obtain

$S_A = S1 = A \cdot B$ therefore $J_A = B$

$R_A = S3 \cdot \overline{HLDA} \cdot \bar{c} = A \cdot \bar{B} \cdot \overline{HLDA} \cdot \bar{c}$, therefore $K_A = \bar{B} \cdot \overline{HLDA} \cdot \bar{c}$

CLR_A = System reset

$S_B = S0 \cdot HLDA = \bar{A} \cdot \bar{B} \cdot HLDA$ therefore $J_B = \bar{A} \cdot HLDA$

$R_B = S2 = A \cdot B$ therefore $K_B = A$

CLR_b = System reset

$HOLD = S0 \cdot C + S1 = \bar{A} \cdot \bar{B} \cdot c + \bar{A} \cdot B = \bar{A} \cdot c + \bar{A} \cdot B$

$h = S1 + S2 \cdot \bar{\phi} = \bar{A} \cdot B + A \cdot B \cdot \bar{\phi} = \bar{A} \cdot B + B \cdot \bar{\phi}$

$K = S1 \cdot \phi = \bar{A} \cdot B \cdot \phi$

The equivalent circuit is shown in Fig. 5(b).

The next article in the series will deal with d.m.a. interfaces

Data recording on cassette

Continued from page 52

sive-OR gate $IC_{4(a)}$ with a delayed version of itself, to produce a narrown pulse. This pulse is used to reset counter IC_5.

The clock requency for IC_5 is provided by phase-locked-loop IC_7 and divider IC_8. The count between resets should be 16 for a data 1 and 32 for data 0. $IC_{3(c)}$ output goes low after count 24, the threshold count midway between the two extremes. Flip flops $IC_{6(a)}$ and $IC_{6(b)}$, together with exclusive-OR gate $IC_{4(b)}$, provide a symmetrical data stream. Variable resistors at pins 11 and 12 of IC_7 are used to give a phase-locked-loop range of approximately 120 kHz to 180 kHz. The 4.8 kHz output at Q5 of IC_8 is locked to the 4.8 kHz output at Q4 of IC_5. The output frequency of the phase-locked loop, given accurate tape speed, will be 153.6 kHz, which is twice the frequency it needs to be for 2400 baud operation. It has been made deliberately so to give the option of recording at 4800 baud on suitable recorders with the minimum amount of circuit change.

A 16 times data rate clock is provided at Q2 of IC_8. Decoder waveforms are shown in Fig. 8.

Testing

The data-recording technique described was tested with a varied selection of audio cassette recorders and tapes over a period of four months. Since then it has been used continuously during the development of Radiotext. During testing, the normal precautions that would be taken in any magnetic recording procedure were observed. In particular the recorders were regularly cleaned and care was taken to avoid damage to the surface of the tape. No attempt was made to correct for head misalignment in any of the recorders.

Testing involved the recording of various pseudo-random sequences with continuous automatic error checking on playback. Additionally use was made of a microcomputer memory verify routine to compare sample blocks of data recorded onto tape with those decoded on playback.

With good quality tape this method of recording performed particularly well with wide margin for error indicated by the "eye diagram" of the decoder signal. From a total of 250 test recordings, only two error bursts were detected, both with samples of a very low cost tape and both due to clearly visible tape defects. The decoder proved insensitive to volume control settings, to the use of automatic record level control and to variations in tape speed.

Use of the circuit described has shown that reliable data recording can be achieved using audio cassette. The requirements of the Radiotext project are satisfied in that it should be possible to record at a data rate of 2400 baud with any low-cost cassette recorder. With minor circuit modification a data rate of 4800 baud can be provided. Performance at this data rate is reliable in all but the most basic recorders.

Acknowledgement
The radiotext project is supported by the Faculty of Technology of the Open University. The authors wish to thank members of staff of the Electronics Discipline for helpful discussions throughout the duration of the project.

References

1. Smith, R. A., A Low Cost Storage System For Use On Audio Cassettes, *Electronic Technology*, Vol. 13, March 1979, pp. 65-69.
2. Koanantakool, T., 4800 Baud Cassette Interface, *New Electronics*, Vol. 12, No. 21, October 30 1979, pp. 36.
3. Mallinson, J. C. and Miller, J. W., On Optimal Codes for Digital Magnetic Recording, Proceedings of the Conference on Video and Data Recording, Birmingham, England, 20-22 July 1976, (I.E.R.E. 1976), pp. 161-9.
4. Preis, D., Phase Equalisation of Analogue Magnetic Recorders by Transversal Filtering, *Electronics Letters*, Vol. 13, No. 5, March 3 1977, pp. 127-8.

Digital, multi-track tape recorder

Digital circuitry for playback

by A. J. Ewins, B.Tech., Research Department, London Transport

The overall design philosophy of the recorder and a detailed description of the digital recording circuitry were presented in previous parts of the article, which continues with a detailed description of the digital playback circuitry.

A block diagram of the playback interface (the peak detector), Miller decoder and associated differentiator and resettable oscillator were shown in Fig. 7 (Nov. 1981). Due to the frequency response characteristics of the cassette tape-recorder, the recorded Miller-coded data stream looks like a series of positive and negative peaks, which are associated with the original positive and negative transitions of the Miller-coded date stream: a typical playback waveform is illustrated in Fig. 24.

The usual way of detecting such peaks is to differentiate the waveform and then detect the resulting zero-crossing points. A differentiator and zero-crossing detector were originally designed and constructed, but were found to be unsatisfactory, due to the fact that the distorted playback signal contained the occasional subsidiary peak not associated with the original recorded positive and negative transitions. An alternative peak detector was sought, and eventually one designed by Brian Evans[1] was modified and found to be highly satisfactory. Figure 25 shows the resulting circuit in detail. To understand the operation of the circuit readers are invited to read Evans' article as it is not proposed to enter a detailed description here. However, a number of the waveforms present in the circuit of Fig. 25 are shown in Fig. 24. The re-shaped, Miller-encoded data is taken from the divide-by-2 counter to the Miller decoder circuit, shown in Fig. 26, with the relevant timing diagram in Fig. 27.

The information contained in the Miller-encoded data is carried solely in the timing intervals between transitions, the direction of a transition being irrelevant. At the input of the decoder circuit, therefore, is a differentiator circuit consisting of a buffer, a capacitor and a 2-input Ex-Or gate. A short-duration, positive pulse is present at the Ex-Or output for every signal transition of the Miller-encoded data, and is used to reset an oscillator running at four times the required tape-clock frequency. It is essential that the recovered tape-clock, RTC, reflects the wow and flutter content of the timing information contained in the Miller encoded data: continually resetting the oscillator allows this to happen, retarding it or advancing it in keeping with the timing errors. A maximum timing error of up to plus or minus one quarter of a RTC cycle may occur before any error is produced in the decoding process. The output from the resettable oscillator is divided by four by a pair of divide-by-2 counters, 'A' and 'B', the output from 'B' being the recovered tape-clock, RTC and $\overline{RTC}$, correctly phased by the output from the phase detector.

The phase detector detects the timing interval in the Miller-encoded data stream produced by a 1, 0, 1 sequence in the NRZ data: it is reset by the output from the differentiator and counts the number of cycles clocked by the resettable oscillator. A count of 8 is the error-free timing period between transitions in the Miller-encoded data of a 1, 0, 1 sequence: a count of 6 would occur for the next smallest timing period (produced by a 1, 0, 0 or 0, 0, 1 sequence). Thus, if a count of 7 is reached (at which point the phase-detector counter disables itself), the timing interval must be that produced by the 1, 0, 1 data sequence. (An output from the phase-detector is also produced by the artificially long interval created by the 1, 0, 0, 1 sequence of the sync. word. This does not, however, produce any error in the phase detection process.) The logic 1 present at the output

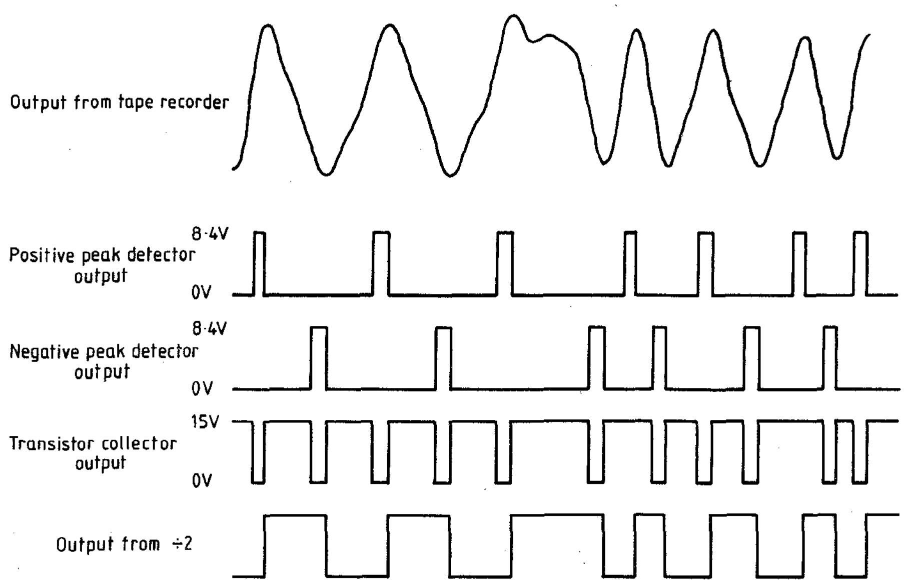


Fig. 24. Operation of the peak detector, designed to avoid the effect of spurious transitions.

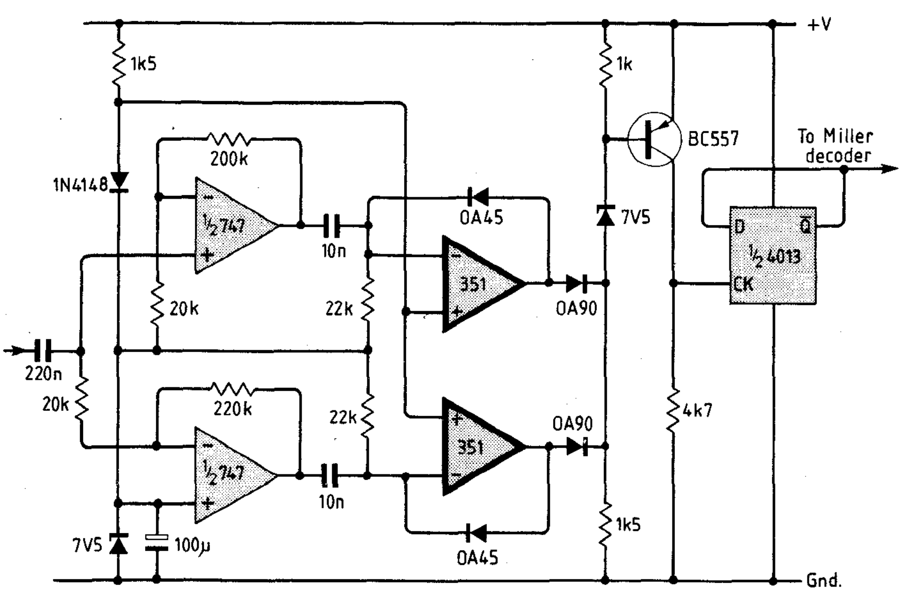


Fig. 25. Circuit of Evans peak detector, which interfaces tape recorder to Miller decoder.

from the phase-detector on receipt of the next pulse from the differentiator is used to set the B flip-flop to its correct phase, via the logic sequence of flip-flops D and E.

In a similar manner, the sync. detector detects the artificially long interval created by the 1, 0, 0, 1 sequence in the sync. word. A count of 12 is the error-free timing period between transitions in the Miller-encoded data of this sequence. Thus, when the sync. detector counter reaches a count of 9 (at which point it too disables itself), the interval detected must be that produced by the 1, 0, 0, 1 sequence of the sync. word. The resulting output pulse from the sync. detector is used to verify the occurrence of the sync. word and is passed to the control circuitry of Fig. 30.

Operation of the rest of the circuit of Fig. 26 is best understood by referring to the logic sequences illustrated in Fig. 27. The Q output from the C flip-flop has the same frequency as RTC, but leads it in phase by 90°, i.e. a quarter of a RTC cycle. This output is Anded with the output from the differentiator, via the 2-input diode And gate, to set the F flip-flop to a logical 1 every time the output from the differentiator is associated with the coding of a 1 bit cell. In the absence of a set pulse the F flip-flop output is clocked to 0, indicating a 0 in the decoded data stream. Flip-flop F output is thus that of the decoded NRZ data stream.

The logic sequence of pulses illustrated in Fig. 27 is drawn assuming no timing errors in the Miller-encoded data. Fig. 28 illustrates the effect of early and late timing pulses on the decoding process. It shows why the maximum timing error that can be tolerated is just less than a quarter of a RTC cycle.

The resettable oscillator of the circuit of Fig. 26 is shown in detail in Fig. 29, and consists of a crystal-controlled oscillator, running free at a frequency of 3.2768 MHz, followed by a divide-by-36 counter. Its output is a frequency of precisely four times the desired RTC; i.e. 91022.2 Hz. The divide-by-36 counter is made up of two divide-by-9 counters, which operate alternately, followed by a divide-by-4 counter: the reason for using two divide-by-9 counters is because the duration of the short positive pulse produced by the differentiator (used to reset the oscillator) is considerably longer than one cycle of the 3.2768MHz oscillator. It is, however, shorter than nine such cycles. The output pulses from the differentiator thus reset the divide-by-4 counter but alternately start and stop, via the divide-by-2 flip-flop, the two divide-by-9 counters. No timing errors due to the finite duration of the differentiator pulses are therefore produced. However, since no attempt is made to reset the actual 3.2768MHz oscillator, a maximum timing error of one cycle of the oscillator is possible. This error is 1/36 of a quarter of an RTC cycle and thus reduces by a very small amount the tolerance of the system to accommodate wow and flutter. All the components of the peak detector circuit and the Miller decoder and clock recovery circuit are constructed on one circuit board*.

Storage buffers and control

The next block diagram of the digital playback electronics to be described in detail is that of Fig. 8 in the November issue. Figures 30, 31, 32 and 33 show the circuit, which consists of four temporary storage buffers, an 8-bit shift-register and associated control circuitry, which remove the sync. word from the data stream and remove the wow and flutter. Inputs to this circuitry are the recovered tape-clock, RTC and $\overline{RTC}$, the sync. pulse, the data-clock, DC and $\overline{DC}$; from the recording stages of the digital electronics, and the decoded serial data stream containing wow and flutter. Outputs are produced that control the subsequent demultiplexing of the serial data and its reconversion from digital to analogue data. The control circuitry shown in Fig. 30 produces the correct sequencing of the filling and emptying of the four storage buffers in Fig. 33. Decoded NRZ serial data passes first of all through the 8-bit shift register with 8-bit parallel outputs. When the 8-bit sync. word is present in the shift-register the 1s comparator and the 0s comparator both produce logic 1s at their outputs. Because it is possible for the 8-bit sync. word to be present elsewhere in the data stream, the sync. pulse from the decoder circuitry is used to verify the position of the true sync. word. However, the sync. pulse is not

*Suggested strip-board layouts will be made available when the series finishes.

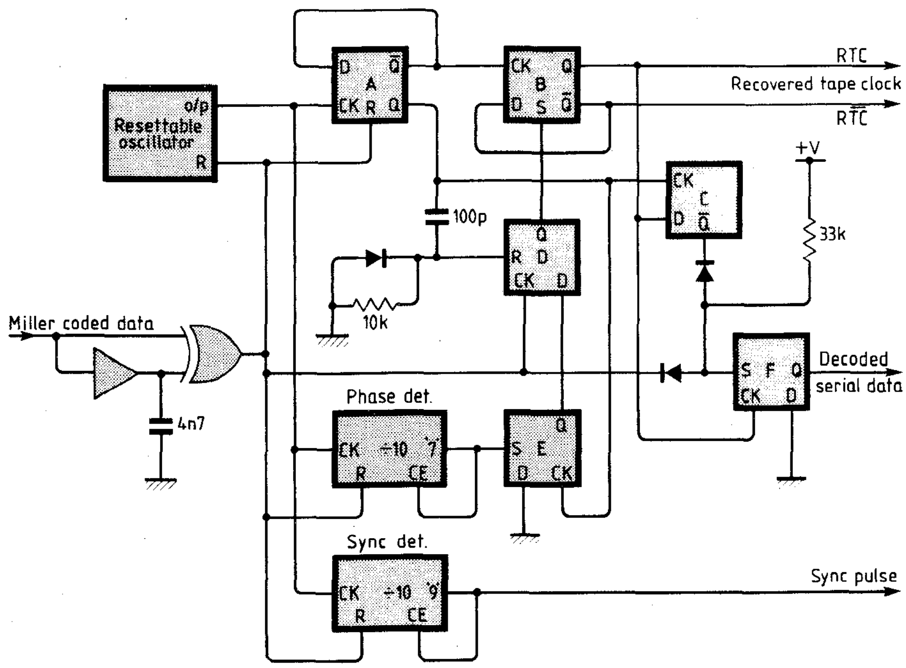


Fig. 26. Miller decoder and circuit for recovery of 'tape clock'.

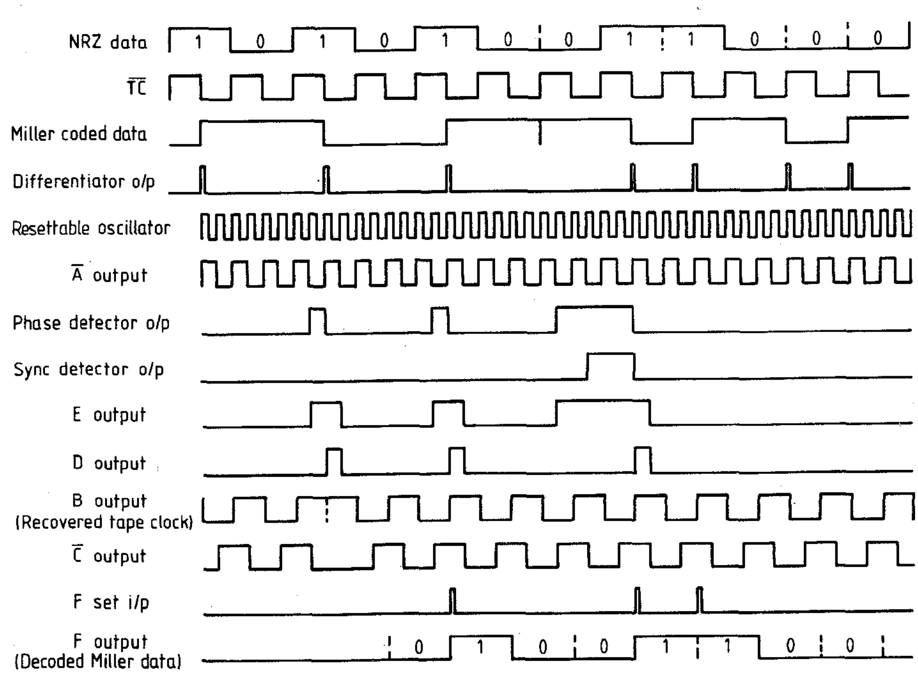


Fig. 27. Operational sequence of Miller decoder and clock recovery circuit.

produced at precisely the right time, and for this reason the two D-type flip-flops, D3 and D4, are included, to delay the sync. pulse to coincide with the instant that the sync. word is detected. The Q output from the D3 flip-flop is the correctly delayed sync. pulse. Thus, at the instant that the sync. word is detected, three logic 1s are simultaneously produced which are Anded, together with RTC, by a four-input And gate. The resulting output pulse from the And gate resets the divide-by-80 counter and thus synchronizes the sequence of filling and emptying the storage buffers. Figure 34 shows the logic sequence of pulses produced by the 0 and 1 comparators, the flip-flops D3 and D4 and the output from the four-input And gate, during the detection of a sync. word.

The filling and emptying sequence of the four 72-stage temporary storage buffers is initially set by a logic 1 pulse on the SET input to the control circuitry. Via a D-type flip-flop, D1, this resets a divide-by-4 counter, A, to its initial state. It also sets another divide-by-4 counter, B, via D-type flip-flop D2 to its initial state. Both divide-by-4 counters, A and B, produce four sequential output pulses which operate gates controlling the flow of data into and out of the four storage buffers and also the clocks used to shift the data. At the instant of a SET pulse then, a logic 1 is produced at the 1 output of the A divide-by-4 counter, A1. A logic 1 is also produced at the 1 output of the B divide-by-4 counter B1. This output is Anded with the $\overline{Q}$ output of D2 and, since this is initially logic 0, no control pulse is output from the And gate, 5. Thus B*1 must be at the logic 0 level. Due to the presence of a logic 1 at the Q output of D2, the divide-by-72 counter is also inhibited from counting. With A1 at the logic 1 level and B*1 at the logic 0 level the serial data will be clocked into the first storage buffer under the control of $\overline{RTC}$. These conditions will remain so long as the SET input stays at the logic 1 level. Once the SET input returns to logic 0 the control circuitry becomes receptive to the RESET pulses produced by the detection of a sync. word.

Upon receipt of the first RESET pulse after the release of the SET input, D1 is clocked and its outputs reverse, releasing the A divide-by-4 counter from its reset state and also D2. The A1 output remains at the logic 1 level, however, and the conditions of the divide-by-72 counter and the B divide-by-4 counter remain the same. A frame of serial data, led by the 8-bit sync. word, is thus clocked into the first storage buffer under the control of $\overline{RTC}$. At the end of eighty $\overline{RTC}$ pulses the 8-bit sync. word will have passed right through the 72-stage buffer and it will contain only one frame of data, i.e. 6 × 12-bit data words. On the eightieth $\overline{RTC}$ clock pulse, the divide-by-80 counter clocks the A divide-by-4 counter to produce a logic 1 at its A2 output. A further RESET pulse, produced by the detection of the next sync. word serves only to synchronise the divide-by-80 counter, coinciding precisely with the eightieth RTC pulse.

With A2 now at logic 1 and all B*n

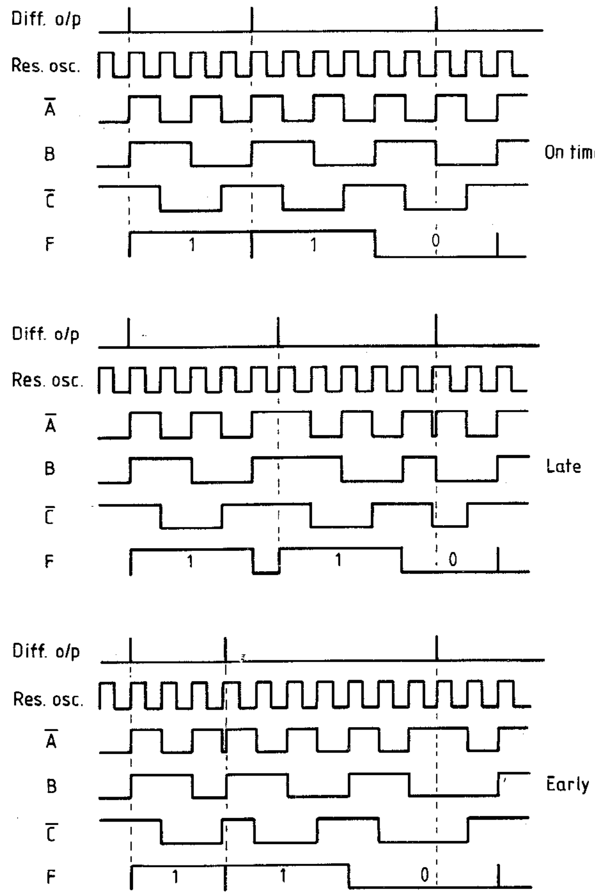


Fig. 28. Effect of early and late timing pulses on Miller decoding process.

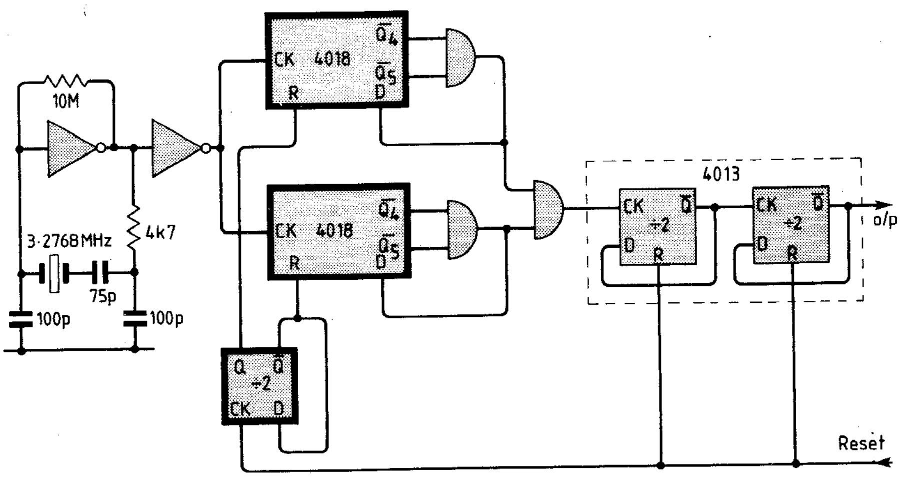


Fig. 29. Resettable oscillator of Fig. 26.

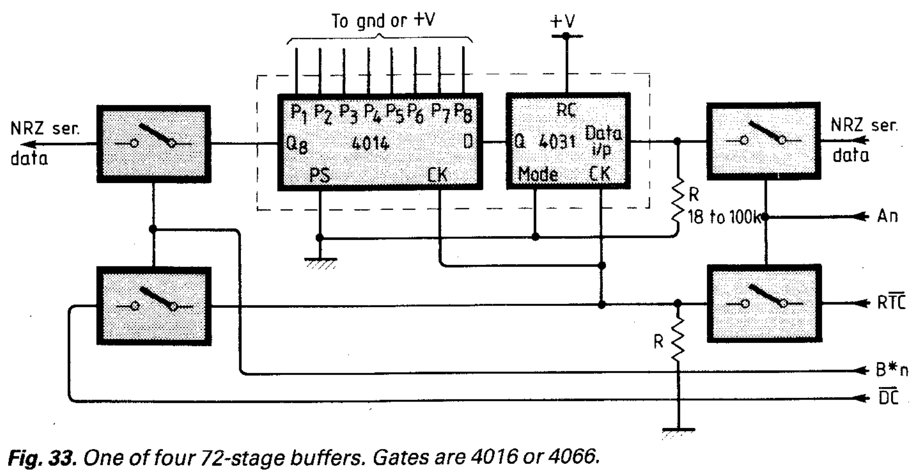


Fig. 33. One of four 72-stage buffers. Gates are 4016 or 4066.

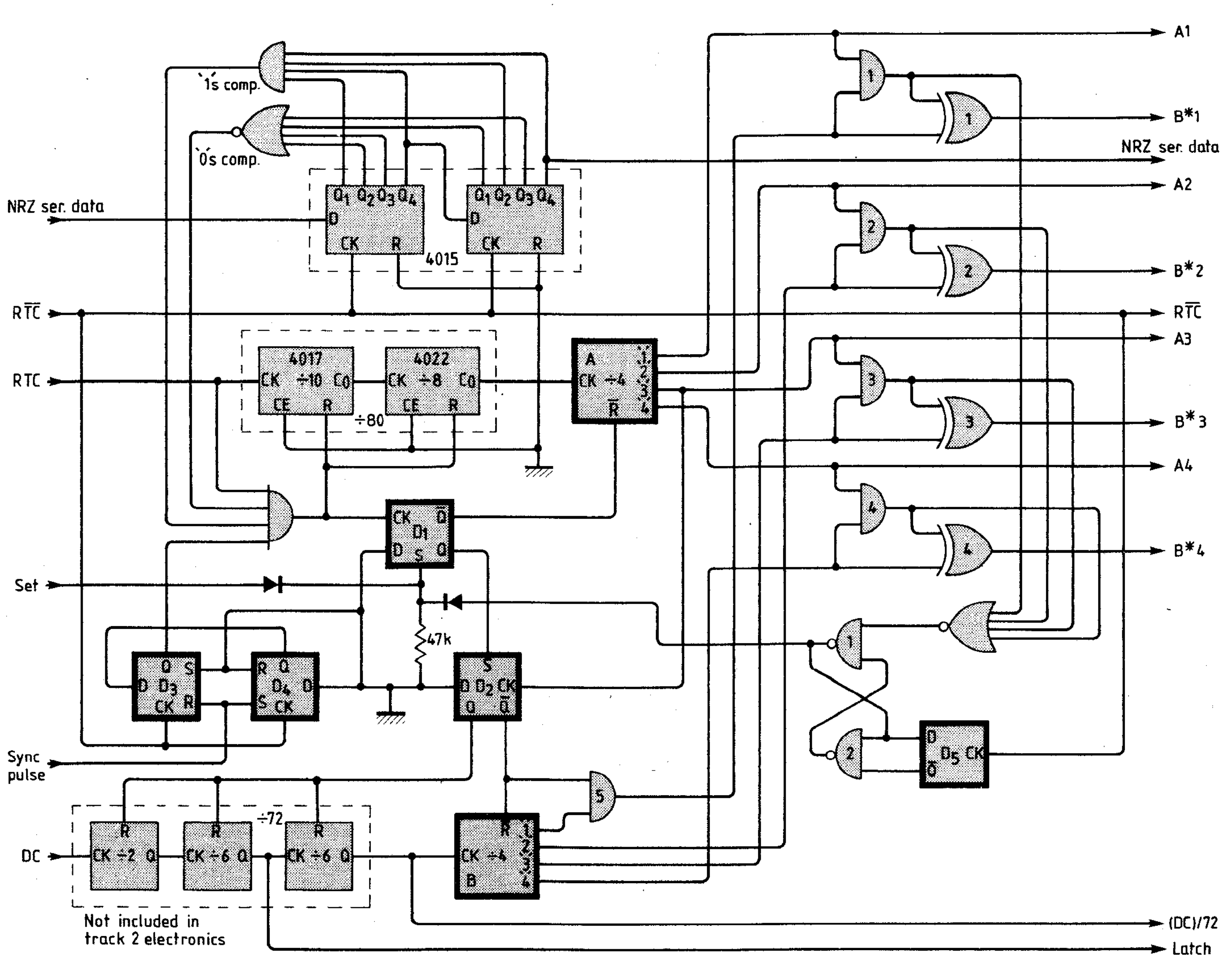


Fig. 30. Control circuitry, shown in block form in Fig. 8 of November article.

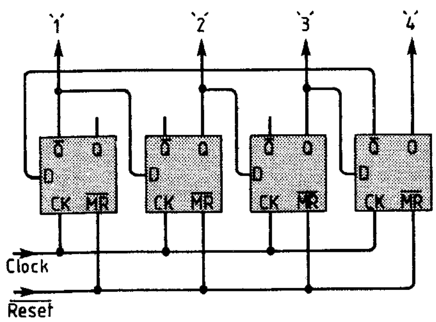


Fig. 31. Divide-by-four circuit, used in control circuitry of Fig. 30, using HEF40175.

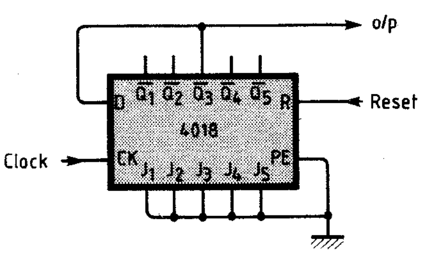


Fig. 32. 4018 divide-by-6 circuit.

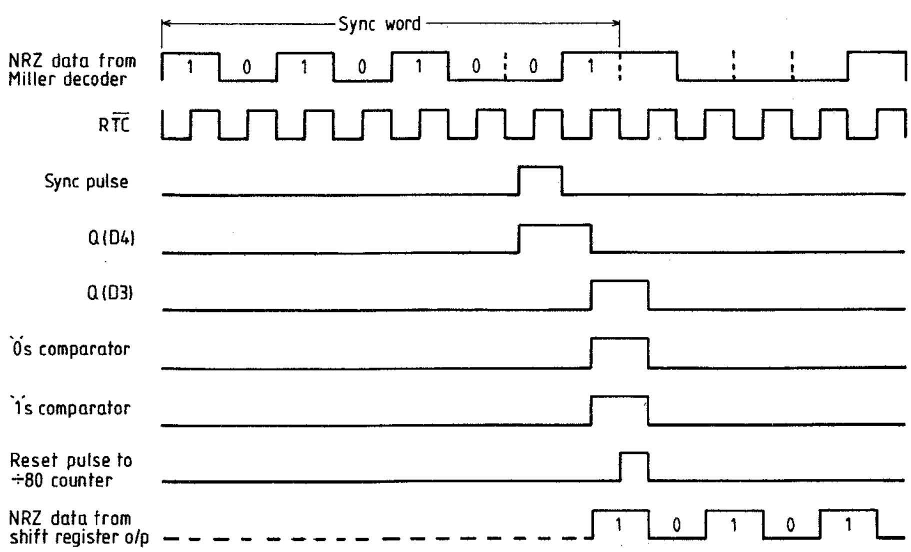


Fig. 34. Sequence of pulses during sync. word detection.

outputs still at logic 0, temporary storage buffer number 2 is now filled with serial data under the control of RTC. At the end of eighty RTC pulses this buffer is also full of data, the 8-bit sync. word having passed right through. The eightieth $\overline{RTC}$ pulse in this sequence clocks the A divide-by-4 counter to produce a logic 1 on its A3 output. Filling of the third storage buffer under the control of $\overline{RTC}$ thus begins. However, as A3 goes to the logic 1 level it clocks D2 so that its outputs change state. This releases the divide-by-72 counter, the B divide-by-4 counter and produces a logic 1 and at the output of And5. B*1 thus becomes logic 1, allowing the data held in the first storage buffer to be emptied under the control of $\overline{DC}$. Thus as buffer number 3 in being filled with data under the control of $\overline{RTC}$, buffer number 1 is being emptied under the control of $\overline{DC}$. The preceding sequence of events creates a time difference of 160 $\overline{RTC}$ pulses (or 144 DC pulses) between the filling and emptying of the storage buffers. This time difference is more than enough to 'mop-up' the wow and flutter content in the incoming data. The long term stability of $\overline{RTC}$, and the speed of the incoming data, are synchronized to $\overline{DC}$ by means of a phase-locked loop in the speed control of the tape-recorder.

With this synchronization between $\overline{RTC}$ and $\overline{DC}$, and the filling and emptying processes of the storage buffers initiated two buffers apart, it is most unlikely for an attempt to be made to empty a buffer whilst it is still being filled. However, to ensure that this does not happen the combination of two-input And gates, numbers 1 to 4, and the two-input Ex-Or gates, numbers 1 to 4, were included. With the combination of Ands and Ex-Ors as shown it is impossible for B*n to be at the logic 1 level at the same time as An; the output An being given precedence over B*n. It is, however, possible for An and Bn to be at the logic 1 level simultaneously and should this occur a logic 1 is output from the n And gate. Via the four-input Nor gate, such an output produces a logic 1 at the output of the two-input Nand gate, 1. This logic 1 output is diode-Ored with the SET input to produce the equivalent of a SET pulse on the input to D1. This automatically resets the filling and emptying process of the four storage buffers two buffers apart. It does, of course, produce an 'error' in the output data. This process will not normally occur in practice, however, unless the lock of the phase-locked loop of the speed control circuit is disturbed. Having produced a logic 1 at the output of Nand 1, the 'toggle' action of Nands 1 and 2 are automatically reset by the first $\overline{RTC}$ pulse to be received by D-type flip-flop D5.

The output (DC)/72 from the divide-by-72 counter is used as a reset pulse in the final block circuit of the demultiplexer and d-to-a converter, to indicate the position of the beginning of a data frame. The LATCH pulse as produced is used in the demultiplexer to identify the presence of a complete data word for demultiplexing.

A and B divide-by-four counters, used in the control circuit of Fig. 30, are constructed from quad, D-type, flip-flop i.cs, type HEF40175, as shown in detail in Fig. 31. The divide-by-6 counters, used in the divide-by-72 counter, use i.c. types 4018, interconnected as shown in Fig. 32. All the electronics of the control circuitry shown in Fig. 30 are constructed on board 5.

Figure 33 is the circuit diagram of one of the four temporary, 72-stage, storage buffers. All four storage buffers are constructed on one circuit board, board 6. The 72-stage buffers are made from an 8-bit shift-register, i.c. type 4014, and a 64-bit shift-register, i.c. type 4031, in the same way as for the 72-stage buffers of the digital 'recording' electronics. The four gates of each buffer are contained in quad switch i.cs, type 4016 or 4066. NRZ serial data is input to the gates of all four buffers in parallel: similarly, the NRZ serial data output from the gates of all four buffers are connected in parallel. Because there will be times when the data input to a buffer would be left floating it is grounded via a resistor of value between 18kΩ and 100kΩ. For the same reason the clock inputs are grounded via a similar valued resistor.

Reference

B. T. Evans. Digital data recording without f.s.k. *Wireless World,* April 1979.

To be continued

Opto-electronic contact breaker –

In the April 1981 issue, a replacement for the conventional contact breaker was described. The following notes have been inspired by a number of enquiries which have been received, and are intended to any other would be constructors.

There have been several enquiries about the source of the specified opto-electronic components (TIL 31 and TIL 81). These are available from most Texas Instruments distributors: the author obtained parts for the prototype from Quarndon Electronics of Derby. According to the TI cross-reference guide, the following alternative parts are close equivalents.

TI	AEG Telefunken	Siemens
TIL 81	BPW 14	BPY 62
TIL 31	CQY 35	CQY 77

These devices have lens ends to accurately define the beam. Cheaper epoxy types are not acceptable.

Several readers have queried the choice of the 5401 chip, and asked whether the 7401 is suitable. The 54 series device is specified to operate from −55°C to +125°C, whereas the 74 series device is only rated from 0°C to 70°C, and cannot be recommended in this application. SN5401J is in a ceramic package and is preferable to the SN5401N, which is encapsulated in plastic.

Some constructors have attempted to retain the existing points as well as fitting the opto-electronic breaker. This cannot be recommended for the following reasons:–

- Conventional points introduce timing scatter by the intermittent nature of the load which they apply to the distributor shaft. The retention of points degrades the scatter performance of the optoelectronic system to that of points.
- The optoelectronic breaker is designed for indefinite life, so the back-up of points is superfluous, and they would be mechanically worn out quite soon, negating the maintenance free concept of the unit.

Many add-on electronic ignition systems, both ready-made and in kit form, suggest that the points capacitor is retained. The main reason is that reversion to conventional ignition is made easier by the retention of the capacitor. When using the optoelectronic breaker, it is not possible to revert to conventional ignition, so the points capacitor should be removed, since it serves only to degrade the risetime of the output signal.

The prototype was designed for high reliability, in that all components are generously de-rated. To minimize dissipation in the regulator zener diode, R_1 is specified as 1k. With this value, there is a possibility of insufficient base current to cater for a low-gain MJE 340 in the series regulator when the supply is down to 7 volts. This is only likely with a large engine at very low temperatures, when the load in the battery due to the starter motor is considerable. The value of R_1 can be reduced to 330Ω, but this requires the substitution of a BZX85C5V6 1.3 W zener in order to have adequate derated dissipation. The author has not seen a problem with the original published values, but this information is supplied to pre-empt any problem in extreme circumstances.

Finally, a word about reliability. The prototype has been in use for six years now without a failure. The vehicle to which it is fitted is regularly taken abroad, and was once driven virtually non-stop from Reading to Geneva with a team of three drivers. Provided that the specified components are used, this degree of reliability can be expected from other examples of the design.

News of the Month

Digital sound mixer

A significant advance in sound reproduction was made when digital equipment was introduced into the recording/reproducing chain. Further advances are still being made as the chain lengthens. The Compact Disc, developed by Philips, seems to have won the race of being established as the world standard digital audio disc, although some other companies are still running as rival contenders. Digital discs complete the chain from microphone to loudspeaker so that sound waves, converted into digits at the input end, suffer no degradation (assuming an adequate digital system) before being reconverted into sound waves at the output end. The middle part of the chain is no longer processing analogues of the sound patterns, but is dealing with digits. Computers rather than amplifiers predominate.

Neve Electronic International, in close collaboration with the BBC, have produced a Digital Signal Processor (DSP). A prototype is undergoing assessment tests at Broadcasting House and when the first production version of the processor is delivered to the BBC in Autumn, it is believed that it will be the world's first comprehensive all-digital sound mixing desk to enter operational service in broadcasting.

The 48-channel mixing desk can perform all the normal processes such as fading, mixing, filtering and compression. In addition it can provide time delays in every channel and control comprehensive signal routing.

The channel processor design is based upon COPAS, the computer for processing audio signals, developed by the BBC's research department. Conventional microprocessors and minicomputers are too slow for audio signals and a 'bit-slice' technique has been employed in COPAS to overcome the problem. COPAS also uses other techniques to maximize its operating speed. Multiplication is done outside the microprocessor in a single-chip multiplier that operates 16 times faster than the multiplying function of the microprocessor itself. Another important technique is known as 'pipe-lining'. This makes it possible to put the next microinstruction into the 'pipe-line' while the first is being executed, and this almost halves the cycle time. Together, these techniques produce a machine in which 16 different 'activities' can be programmed into each 56-bit micro-instruction, which can be executed in 140 nanoseconds.

The production version of the DSP will be installed in a BBC digitally equipped radio outside broadcast vehicle to be used on a variety of programme applications. The vehicle will also contain two fixed-head digital tape machines and will have provision for a multitrack machine.

Neve still have faith in the continuation of the demand for analogue studio equipment and at the same time that they announced the digital console, they also launched three new series of analogue consoles.

The 51 series of broadcasting production consoles, which are constructed from a number of modules, are the result of the ingenious application of appropriate op-amps and interconnection techniques to give a family of versatile flexible building blocks. The range includes the 5104, a four-bus, two-output mixer from 12 to 48 inputs with a choice of facilities. The 5116 adds the facility of 24 track metering an monitoring with comprehensive over-dub facilities. It offers comprehensive multi-track capabilities more easily adaptable to the layout of radio and television studios.

The 4322 on-air consoles have been designed specifically for 'live' broadcasts, continuity and disc jockeys etc. as may be found for instance in local radio. It is intended to cater for a wide diversity of operational requirements with a minimum of exposed controls. This has been achieved with a number of pre-set links and switches which may be configured to adapt the console to individual users.

Finally the 8128 multi-track music recording console is a range of mixer desks for the recording industry. It has evolved from the commercially successful 8108 range and incorporates Neve's Formant Spectrum Equalisation, the name they use to describe their equalizer charactistic. They pay particular attention to subjective listening tests and claim that their consoles are the most 'musical' ones available.

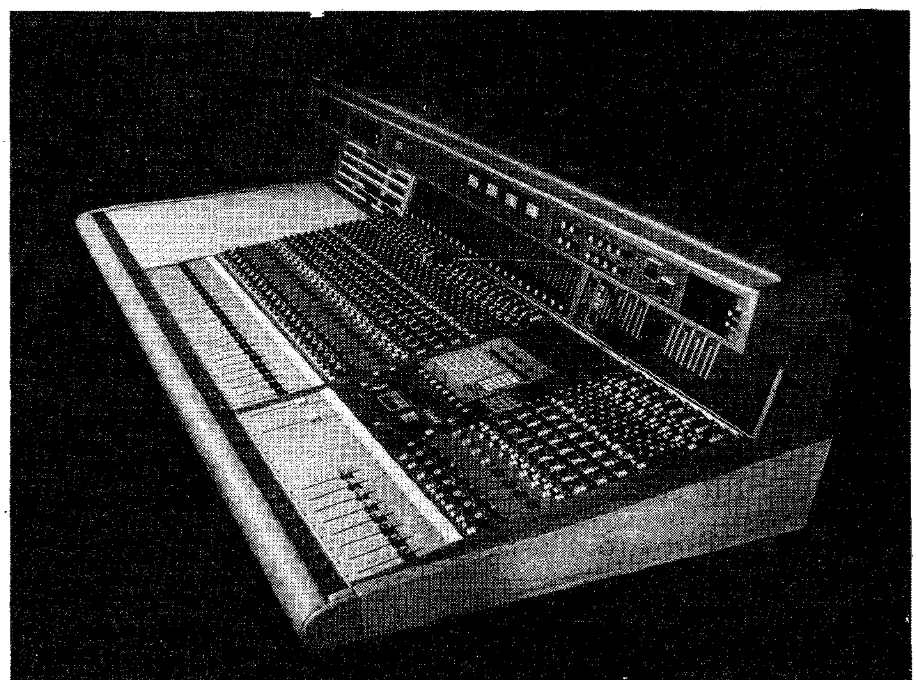
Digital telecommunications by X-Stream

Following on from the development of System X telephone exchanges, British Telecom are to launch digital communications services. The new services are to be marketed under the general title 'X-Stream' and have been made possible through the spread of digital transmission through improved cables, microwave and optical fibre networks. The services are known as Megastream, Kilostream, Switchstream and Satstream. The first to be offered is Megastream which transmits at 2 or 8 Mbit/s. A basic form of Megastream is already working on a special London overlay network which started in September 1981 as a private circuit for Chase Manhatten Bank. The overlay uses 2Mbit/s links capable of carrying 30 telephone conversations simultaneously. About 30 orders for similar services are already being processed by BT.

By the end of 1982 the Kilostream service will be in operation, offering digital services at 2.4, 4.8, 9.6, 48 and 64 Kbit/s on a special private circuit network. The network will interlink the London overlay – covering 45 exchanges in the capital – with 30 towns and cities in the rest of Britain. It will be extended to cover nearly 200 business centres by 1985.

Switchstream will combine digital transmission links with the System X digital telephone exchanges to create an integrated services digital network. A pilot scheme with capacity for about 250 businesses will be based on the large local System X exchange installed in Baynard House, in the City of London.

Telecom will be starting a fourth digital service at about the same time that Switchstream commences, towards the end of 1983. This will use small-dish terminals beamed to the European communications satellite. Called Satstream, this service is intended for private business communications across Europe.



The Neve 8182 multi-track music console for 24, 32 or 48 track recording or mix-down.

Education and industry

The Department of Industry has announced that they are ready to receive entries for the Young Engineer for Britain 1982 competition. Boys and girls between the ages of 14 and 19 are eligible, whether they're at school, college, university, or working in industry. Entrants are required to produce a project which could be of a mechanical, electronic or chemical nature. It should aim to improve industrial production or an existing process, have commercial potential, save waste or conserve energy, or meet a social or domestic need. The project can be as simple or as complex as the competitor chooses. Entries are divided into age classes: 14-15; 16-17 and 18-19 years with a separate class for entries from industry. Entrants can be individuals or groups of up to four. The closing date for entries is March 31. You may remember our report in December 1981 of the three young men who won a prize for their computer design.

Patrick Jenkin, the Secretary of State for Industry, has announced an award scheme for stimulating improvement in the competitive performance of British industry and commerce through successful participation between higher education and industry or commerce. Attempting to remove the barriers between academics and businessmen, the scheme is intended to recognize the contribution that academic knowledge can make towards commercial applications. Awards will be granted by a panel of judges chaired by Sir Henry Chilver, Vice-Chancellor of the Cranfield Institute of Technology, who will assess the benefits accruing to companies participating in the scheme, such as improvement to a product, process or service, or in the development of the company and its personnel. They will also take into account benefits to the educational establishment, to education as a whole and to the community.

Two cash grants of £25,000 each will be awarded to the most successful of the education teams for the purpose of helping to develop the joint venture. Entries are to be submitted, with supporting material by April 30th. The scheme is to be known as the EPIC Award (Education in Partnership with Industry and Commerce).

Details of both competitions can be obtained from the Industry/Education Unit, Department of Industry, Room 354, Ashdown House, 123 Victoria Street, London SW1E 6RB Telephone: 01-212 0458.

Active silence

The British Technology Group have published details of the installation of the first large-scale active silencing system at a British Gas compressor station at Duxford. The technique was originally proposed by Dr Malcolm Swinbanks, ten years ago, while he was working at Cambridge University. Aware that passive silencers for low-frequency noise were not practicable, he analysed the theoretical behaviour of active sound absorbers in ducts carrying airflow. The potential application to industrial gas turbines became apparent.

The NRDC (now part of BTG) supported research in the active noise control field through a series of projects including one with Dr Swinbanks and Topexpress Ltd to carry out a feasibility study on the gas turbine silencer at Duxford.

The basic principle of active silencing is that an additional source of sound is deliberately introduced to provide an anti-phase replica of the unwanted disturbance, when then opposes and reduces the severity of that disturbance. Modern electronics and control techniques have made possible the suppression of random, aperiodic noise by the rapid, continuous generation of its inverted replica. At Duxford, the 34 ft high, 10 ft diameter exhaust stack emitted a rumble in the lowest audible octave, i.e. from 22.5 Hz to 45Hz, and this could be heard or felt at distances up to a kilometre. The active silencer consists of four microphones mounted inside the stack which pick up the exhaust noise; programmable digital filtering apparatus for processing the noise signal, and powerful audio equipment with 12 amplifiers capable of a total of 11kW output which drive 72 loudspeakers arranged around the top of the stack. The loudspeakers are housed within the exter-

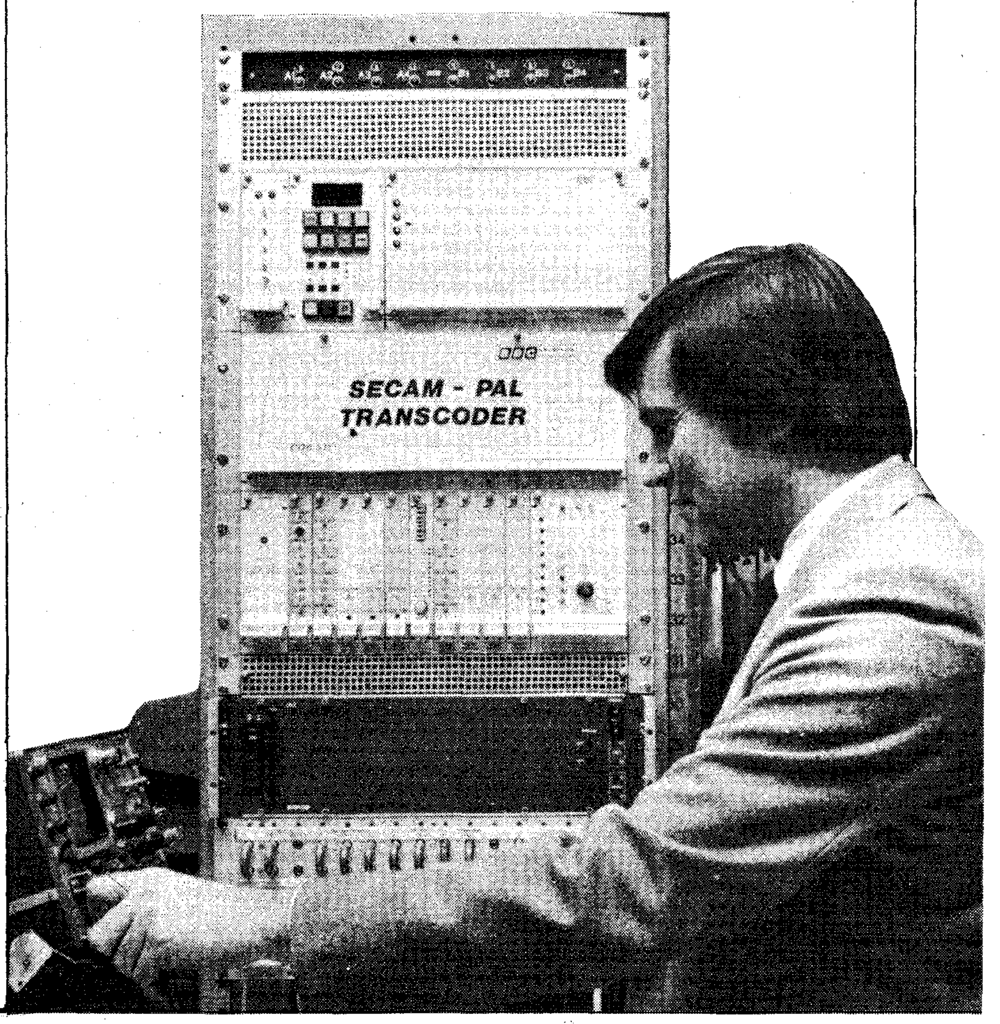
Digital tv standards converter

Following closely on the establishment of international standards for digital television signals the BBC have announced the development of a digital SECAM to PAL transcoder using the CCIR recommended digital sampling standard of 13.5MHz for luminance and 6.75MHz for each of the colour-difference signals.

The incoming SECAM signal is decoded into the luminance (Y) and the two colour-difference signals (U and V). These are converted to digital signals and placed in a two-field store with a clocking signal rate of 864 clock pulses per line. This operation is locked to the input line frequency. The contents of the store are then read out at the same number of clock pulses per line but with the clocking frequency locked to the PAL output frequency. This has the effect of widening or narrowing the time interval between samples in order to produce a correct output line frequency. The signal components are re-converted to analogue signals and coded into a system I PAL signal.

If the SECAM input is replaced by a high-quality PAL decoder, then the transcoder may be used as a synchronizer. As the input clock has been designed to follow the rapid changes in line length associated with some helical scan video recorders, the transcoder can operate as a time-base corrector.

The transcoder was developed by the BBC's engineering design department. David Bradshaw, head of the design section says; "The new equipment has been designed to transcode a signal from SECAM to PAL without the difficulties associated with earlier techniques for carrying out this process."



Roger Robinson, a design engineer at the BBC, testing the SECAM to PAL digital transcoder.

nal cladding of the stack and none of the equipment is visible from the ground.

Trials and modifications of the system have taken nearly a year but the low-frequency rumble has now been effectively eliminated.

The results achieved at Duxford indicate that active silencing techniques may be applied with success to low-frequency noise emitted by powerful gas turbine engines. It may be incorporated into original designs or added to existing installations. The technique could lead to smaller stacks, saving building costs and improving the turbine efficiency. Jet-engine test beds and airport ground-running facilities are suitable candidates for the system.

David Bradly with his turntable and parallel tracking pick-up arm which he constructed from the design in "Wireless World." It was made as a project for his 'O' level 'Elements of Engineering Design' and counts for 20% of the marks for that exam. David built the power supply separately to reduce hum. He confesses that he had to copy the gearing set out in the original article as he found it hard to design a closed gear system. He is modifying the gear system to the second version (Wireless World, July 1981). David will continue with Engineering Design in 'A' level along with physics and maths.

Shuttle success

Despite the truncated second voyage of the American space shuttle, Columbia, the scientific experiments on board all produced successful results, according to a report from NASA. The ground crew were able to operate the experiments for more than the minimum of required observation time for most of the experiments. The shuttle imaging radar obtained eight hours of operation with excellent data from all over the world. Looking at the earth from an angle, unlike the Seasat radar satellite which looks straight down, the experiment provided a lot of relief information. The ability to penetrate cloud cover and vegetation provided valuable data for geological exploration. The radar results were recorded on board the shuttle by filming the images produced on an oscilloscope.

Similarly useful results were obtained from:
– the multi-spectral infrared radiometer, which will also aid petralogical and mineralogical exploration.
– the ocean colour experiment. Colour analysis of the oceans indicate the position of chlorophyll concentrations. This locates plankton and the fishes that feed on them, and would help commercial fisheries as well as oceanographers.
– the measurement of air pollution. Sensors detected the concentration of carbon monoxide in the atmosphere.

Less successful was the spectral feature identification and landmark experiment which used a microprocessor to determine from a spectral 'signature' whether a scene contained vegetation, water, barren land or clouds/snow. Some of the data may have been lost due to the marginal performance of a trigger mechanism which sighted the Sun and triggered the experiment when the sun was at an approximate 'morning' or 'afternoon' position. The crew were asked to film and record lightning storms as Apollo and Skylab crews had reported that such storms seemed to rhythmic or cyclic in character, but it is thought that insufficient information had been collected to investigate this. A plant growth experiment had insufficient time to allow the plants to grow.

The general preliminary conclusion is that in spite of the shortened duration, the mission was highly successful. The shuttle has carried its first scientific payload, including the ESA Spacelab pallet.

Intelligent electronics for A310 Airbus

All-digital electronics units, which can automatically ensure the safe operation of an airliner's wing-mount slats and flaps during take-off and landing, have been delivered by Marconi Avionics on schedule to Aerospatiale, Toulouse France, for installation on the new A310 Airbus. Marconi is under contract to supply the computers, claimed to be the world's first all-digital units, to Liebherr Aero-Technik, the West German company responsible for the A310's slat and flap control system.

Marconi Avionics Flight Controls Division of Rochester, joined forces with Liebherr in a successful bid to supply this system (see WW, April 1980, p44). Although the computers incorporate an entirely new technique in "fail safe" systems design, involving separately programmed pairs of microprocessors of different types, the delivery of the first control channel was made only seven months after the go-ahead and the first aircraft set of flight hardware was delivered just fourteen months after that. A further five units of the control electronics to 'A' model standard, have also been delivered for systems testing at Liebherr, Lindenberg; VFW, Bremen, and Aerospatiale, Toulouse.

The microprocessors are used to control the positioning of slat and flap control surfaces in response to pilot commands. Inbuilt protection features prevent the retraction of leading edge slats below the safe limiting value of wing incidence and inhibit wrong operation of the powerful flying controls. In addition, any condition which might cause asymmetric operation of slats or flaps (i.e. operation on one wing only) is prevented, and any failure which might cause a control surface "runaway" is automatically isolated. The unit has been described as "intelligent", because it can assess the validity of slat and flap commands detect failures in the electronic, electrohydraulic or mechanical parts of the control system and communicate information to the flight crew.

A high degree of integrity required has been achieved through the use of dual control systems, in separate units, each of which contains duplex electronics for slats and for flaps. By careful design, only six tapes of electronic sub-assembly are used, each appearing four times in the complete aircraft installation, conferring important logistic advantages.

Dual dissimilar microprocessors and software ensures that a fault in the program of either of them will be detected by the other, a safety measure which avoids the disadvantages of having an analogue control lane as an extra safety monitor. The microprocessors used are the Intel 8085 and the Motorola 6800, both of which are approved for airborne applications.

A.m. c.b. will never be legal – official

The Home Office has issued the following advice to traders and prospective c.b. buyers: "Don't be misled by unfounded rumours claiming that the use of illicit 27MHz a.m. sets will be legalized; the Government has no intention of making any changes to the new legal 27MHz f.m. c.b. service". The warning follows a large number of inquiries to the Home Office concerning rumours of a.m. legalization, and reports of a.m. sets carrying labels stating that the apparatus should not be used "until April 1982" or similar wording. Any such stickers or lables which imply pending changes in the U.K. c.b. service are quite simply hoaxes.

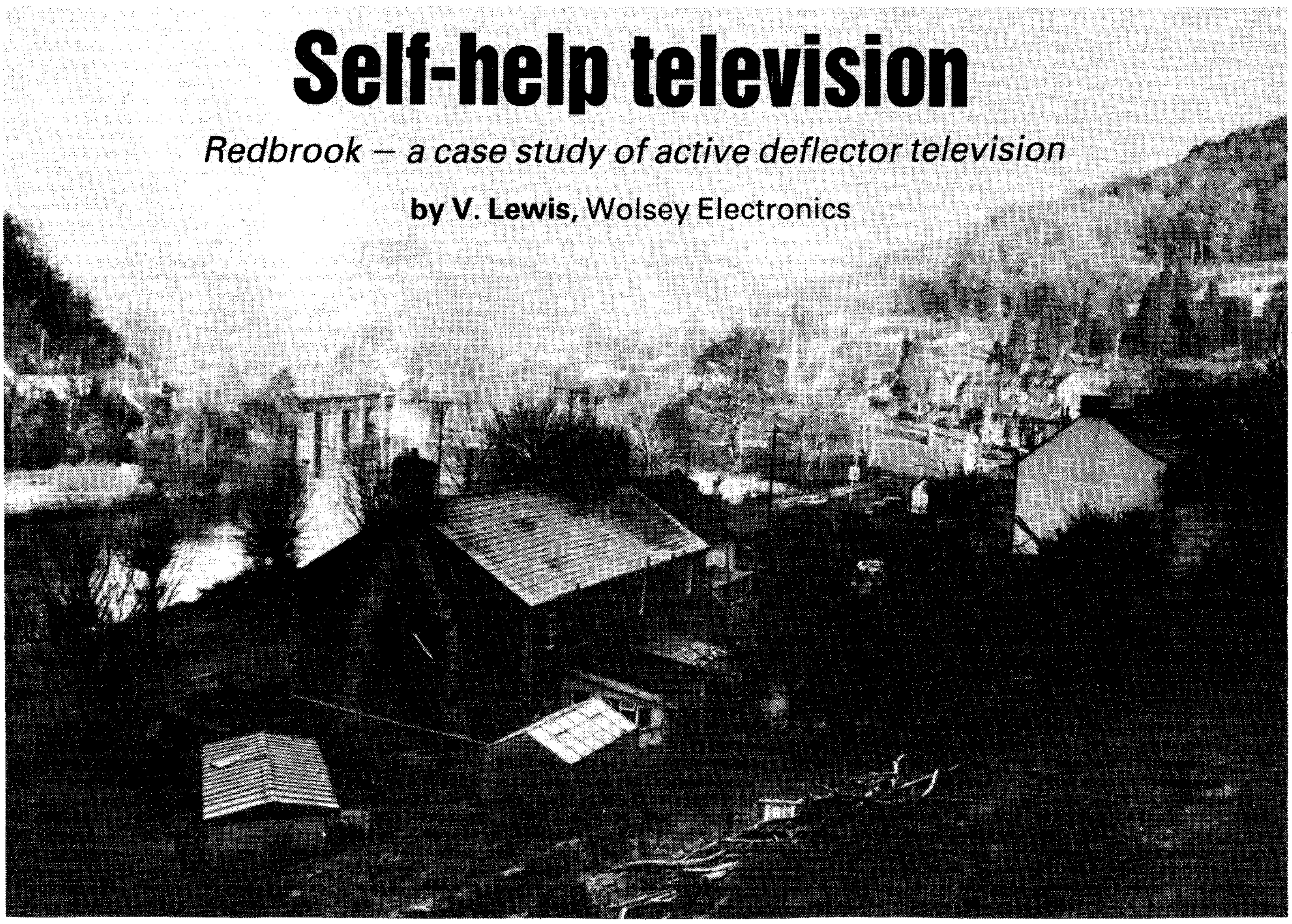
Human aspects of computer systems

A short course on ergonomics for managers, designers and users of computer systems is being organised by Brian Pearce at the Loughborough University of Technology, from 21st to 26th March. It is to be repeated in September. The course covers such topics as visual display ergonomics, the environment, dialogue design, the user's task, user support and training, and the human aspects of the system design process.

Self-help television

Redbrook – a case study of active deflector television

by V. Lewis, Wolsey Electronics



This case history outlines one of the largest "active deflector" systems installed in the UK. Being an official forerunner of this type and size of self-help community approach, much engineering time and money was involved in the venture. Results of colour television and teletext operation far exceeded expectations and delighted the community, who had struggled for years to receive movement on their television screens undisguised by snow and multireflections of seasonal variation.

Redbrook is a village near Monmouth, Gwent, spread over two well-screened valleys with a total population of about 350. Unserved by the national television network, there are areas of the village which have received very poor 625-line signals from Mendip, Ridge Hill, and 405-line signals from Wenvoe. The community is entitled to an official uhf television relay station under phase 3 of the UK 625-line television broadcast plan from about 1984 to 1986. But after consideration at village meetings it was decided to improve the reception of uhf television signals during the projected four to six year interval. Both cable and active deflector schemes were considered and the lower transmitting site capital cost of an active deflector system chosen. It was my opinion that the cable system was preferable technically and the overall total cost comparable, taken over the period envisaged.

* Paper to the Society of Cable Television Engineers, April 1980.

Transmitting Authorities uhf television coverage plan

Phase 1	> 1000 populations
Phase 2	500 to 1000 populations by 1984
Phase 3	200 to 500 populations from 1984 to 1986 (flexible)

In consultation with the secretary of the Redbrook Community Committee, a plan of action was drawn up.

1 Determine receiving site from survey by M. W. Turner in conjunction with IBA engineers. Choose suitable receiving site based on stable reception of three channels with line-of-sight reception and no visual adjacent channel or co-channel interference in normal conditions. (The projected Monmouth television relay station has allocations that may present a problem in the future, as recorded on the Home Office licence application form.)

2 Determine transmitting site similarly. Map village reception areas onto Ordnance Survey map with the necessary horizintal and vertical reception angles together with distances. Consider possible interference to existing viewers who may not participate in the scheme. Also consider possible interference to planned reception areas from existing television signals, taking cross polarization protection and directivity protection ratios into account.

Direct reception of Mendip signals in the smaller screened area to be covered at Redbrook could produce aerial terminal voltages of about 36dBμV which would interfere with the active deflector signals. Application to change channels by means of block convertor equipment was refused (the transmitting authority planning computer showed no possible interference-free channel allocations), and with a channelized convertor being too expensive for this small area of coverage it was decided that the minimum received aerial terminal voltages from the active deflector transmitter must be around 60dBμV. Combined with a degree of directivity and cross polarization protection, this level provided a satisfactory result.

3 Plan minimum transmitting power for the two areas calculated from signal levels necessary from transmitting to receiving aerials.

4 Check planning permission, sites location permission and sites access permission. Agree possible power cable types, routes and extent of practical community involvement.

5 Plan field trials of both areas with sample test readings of signal levels to form overall signal level contour map. Determine percentage coverage and also information allowing additional plans for any unserved areas.

6 Agree target dates of survey, field trials, IBA and Home Office applications, site commissioning tests and trials. Discuss and agree engineering and contractor

costs for installation and planned maintenance schemes.

These six points are examined in more detail in the following, un-numbered, paragraphs but in the same order. A suitable receiving site was established 380 metres from the transmitting site, with line-of-sight reception from the Mendip transmitter providing aerial terminal signal levels of approximately 65dBμV per channel.

Receiving site

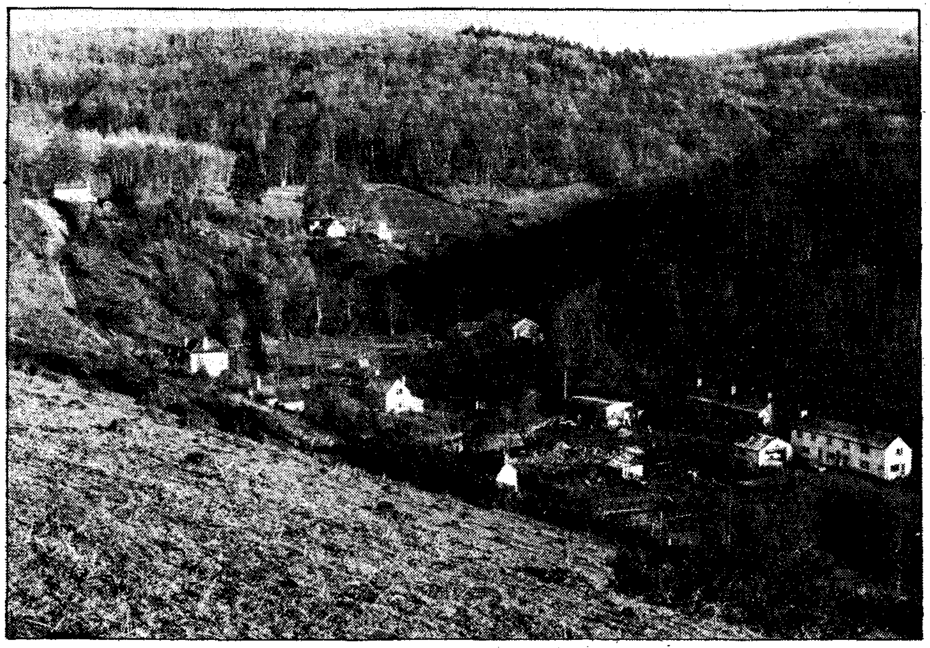
National grid reference	SO534108
site height	125 metres above ordnance datum
Programme source	Mendip (11,000)
Channels	54, 58, 61, 64
Polarization	horizontal

A Wolsey HG36 aerial was mounted approximately 4m above ground level on a concrete lamp post sunk 2m into the ground and concreted in place. An aerial input filter suppressed Wenvoe signals and a Countryman system of amplifiers routed the signals at appropriate levels to the transmitting site.

The transmitting site chosen produced a view of both areas involved and in conjunction with a reception area survey diagram outlining required horizontal and vertical transmitting aerial polar responses, a decision was made to use two aerials (Colour King) fed from separate amplifiers. These were built into two weatherproof equipment housings in the Wolsey laboratory and tested with a spectrum analyser to check spurious output signals. On site the pre-aligned and tested transmitting cabinets were bolted back-to-back around a further concrete lamp-post fitted in a similar manner to the receiving aerial mast.

Calculation of minimum transmitter output voltages required was based on a minimum reception area aerial terminal voltage (high gain QR18 aerial type) specified by Wolsey Engineering. This level combined with a low-noise masthead amplifier is calculated to present a minimum signal-to-noise ratio of 42dB to the receiver input with an average coaxial down lead loss of 4dB. Addition of line-of-sight propagation loss to the minimum received aerial terminal voltage, plus the gain of transmitting and receiving aerial, allows simple calculation of transmitter output voltages into 75 ohms. From this type of calculation the Redbrook equipment was planned with the signal levels as shown in the diagram.

With the type of community relationships involved at Redbrook no problem existed with receiver/transmitter site locations and access permission. Powering was achieved by feeding 55V ac from a farmhouse situated 120 metres from the transmitting site along an overhead pair of 5A wires. Originally this supply voltage was planned to be fed via coaxial cable feeding the farmhouse with signal but at the field trial stage a good direct signal was measured at this point. Two transformer units stepped the voltages down from 240 to 55V and back up again to feed the equipment housing transmitter amplifiers. The Countryman aerial and system repeater amplifiers were fed with 24V dc supplied from the common power unit dc power rail via the receiving-to-transmitting site coaxial cabling on a line powering basis.

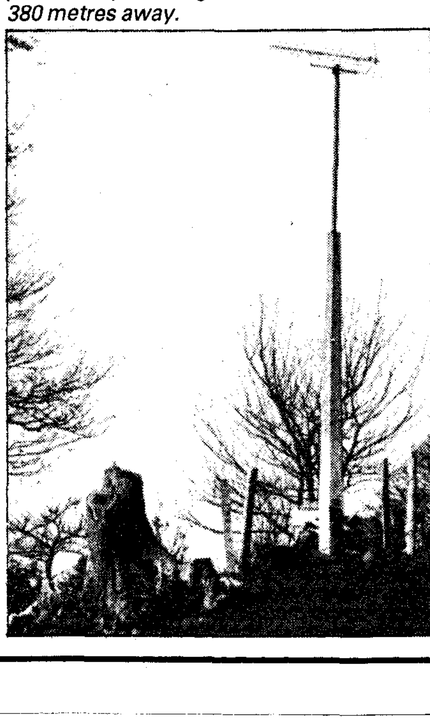


Both block and channelized converters were too expensive for this small area of the village; channelized amplifiers together with cross polarization and aerial directivity provide a cheaper alternative.

Transmitting site

National grid reference	SO535106
Site height	100 metres above ordnance datum
Transmitting aerial height	12ft above ground level
Transmitting power	area 1: 53mW into 75Ω per channel area 2: 13mW into 75Ω
Aerials	two directional, oriented on approximately 50° and 170° east of true north.
Channels	as received
Polarization	vertical

Receiving site with aerial mounted on concrete lamp-post 125 metres above datum to receive signals from Mendip, filters interference from Wenvoe and passes amplified signals to the transmitter 380 metres away.



For the field trials a test transmitting aerial system simulated final arrangements, fed from amplifiers operated at the planned output levels. A petrol electric generating set powered the equipment, which was in use for a 12 hour period. Communication by radio telephone is essential for these schemes. Received signal level readings were taken using a QR18 aerial mounted on a 5 metre mast, with a signal level meter. Readings taken at middle and extremity points of both areas showed that target minimum receiver signal levels were achieved in about 95% of the tests. One area in the lea of the hill supporting the television transmitting mast was almost totally screened and it was agreed that this area, representing about 5% of the total, would be cabled from a good signal area about 100 metres distant. This coverage was considered to be excellent, especially when compared with published minimum percentage coverage of villages.

Discussion on the possibility of a cabled system continued up to June 1980 when the Home Office changed its policy regarding active deflector systems. At a village meeting on 16th August, the committee were authorized to proceed with the active deflector systems. Wolsey were now pressured to work as quickly as possible toward the field trials stage but there was uncertainty regarding firm commitment of capital expenditure due to Home Office advice that expense should not be incurred before "approval in principle". An operating licence will not be issued until the BBC/IBA engineers confirm that the system does not cause interference to other viewers or services. Following official application, the Home Office initiates clearance procedures to check that other users of radio frequencies in the UK have no objection to the use of the site and frequencies chosen, and similarly internationally, with the additional registration of the station by the International Telecommunication Union.

But despite all the problems, which may appear insurmountable to a layman, the system was successfully finally commissioned and tested by Wolsey engineers and the contractors, TeleRadio of Bristol, on November 22 1980. Yet to date, despite praise of the system by all concerned, Redbrook has still not received an operating licence – a matter of some concern to those involved.

Summing up

Cable systems have many advantages over active deflector systems including capability for multiple channels, fm radio, teletext, future channel allocations, as well as conservation of air space, control of system radiation and immunity problems, outlet user control, precictable system planning, cost and signal quality, and predictable timing of installation and licencing. But there are many situations where active deflector systems are the obvious choice; I strongly recommend that full consideration be given to all the factors involved, particularly the commercial aspects which can be the most misleading.

The system cables were of distribution – quality taped and braided types with an integral aluminium barrier for the underground cable runs. This alminium barrier is a precaution against the ingress of moisture but its full effectiveness is debatable if the outer cable covering is broken. The new bonded-shield cables, now manufactured by Raydex cables and marketed by Wolsey Electronics, provide an additional protection which is well worth considering for either scheme.

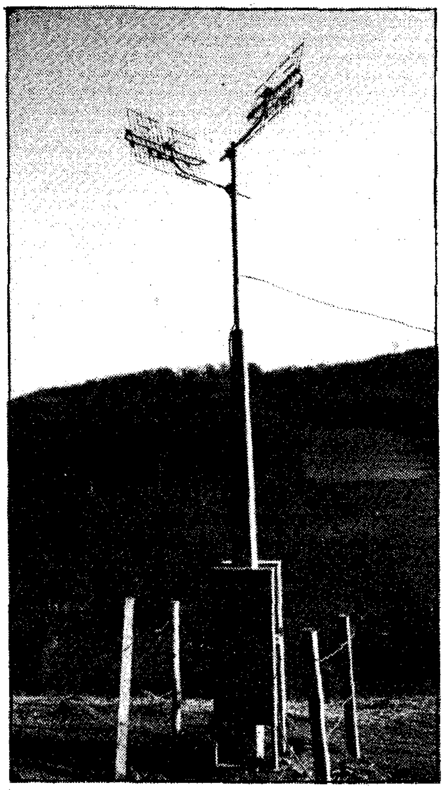
During the field trials a receiving – transmitting site separation of about 10 metres existed with the receiving aerial directed toward the back of the transmitting aerial. Under these conditions, with the high output transmitting amplifier connected, loop feedback presented a problem. This was resolved with the receiving aerial in its final position some 380 metres distant. Directivity and cross polarization protection ratios will not provide much assistance if the aerials are closely sited or co-sited employing the same channels. Transmitting amplifier to aerial feedback can be an additional problem where transmitting mast heights are low leading to possible close proximity of operation. Under these conditions good screening and matching of components is essential, especially during commissioning when the equipment and component access covers may be removed.

Only the non-technical jobs should be attempted by the community members who are normally unskilled, even these jobs should be carefully discussed and supervised. Despite detailed instructions for the cable laying task at Redbrook some of the underground cables had to be replaced after open circuits occurred due to cable stretching by a tractor-type machine!

Even with first-class community relationships social problems can occur which can be almost impossible to overcome. Where interference to existing viewers, can be a major obstacle irrespective of the quality of reception, full publicity and consideration must be afforded to the active deflector system. Take note that outside of the contributing members, viewers cannot be prevented from receiving from such a private system – especially if their original reception has been marred by it. This is, of course, a community problem and system installers would be well advised to leave it to them.

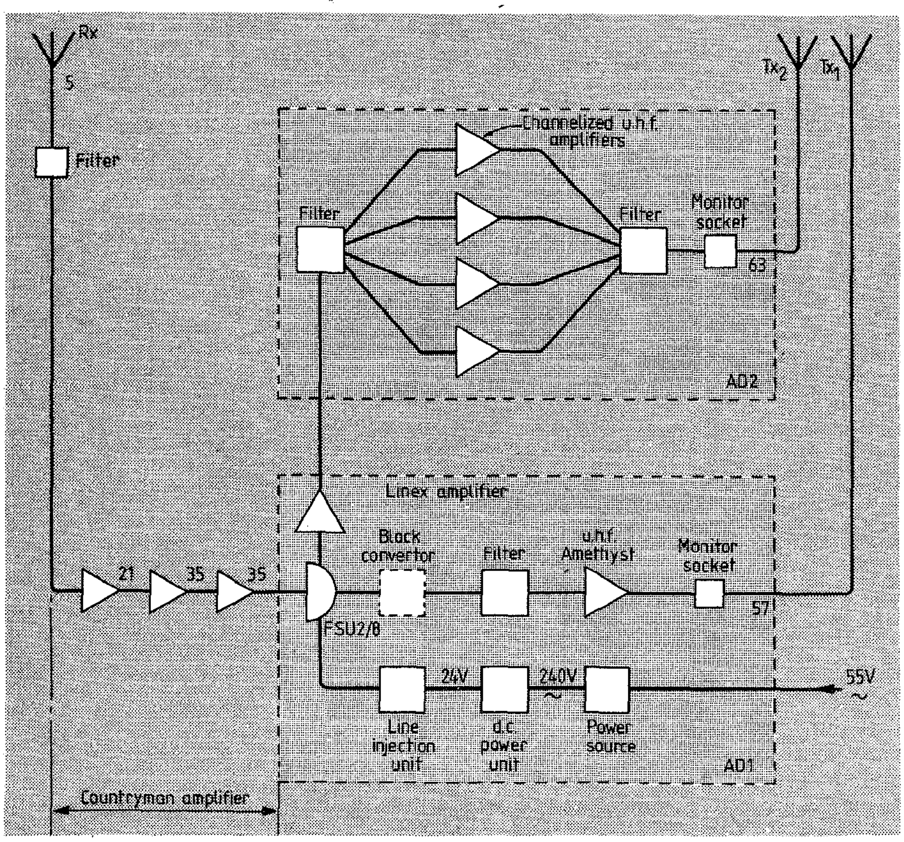
With the complications encountered in this type of system, approved installation contractors are demanded as also is strict compliance to recommended equipment if engineering liability is involved. The system depends on low-level received signals which dictates the use of low-noise masthead amplifiers and high gain aerials with specified gains, directivity and cross polarization specifications.

We did not expect to profit from this pilot scheme but we were obliged to charge direct engineering time and system engineering time involved from many points of view. Maintenance contracts and insurance are considered essential to protect both community and supplier/installer and to operate the system on a correct business basis. With phase three of the national plan projected to 1986, it is anticipated that many schemes of all sizes will be entered into in future years. Whether cable or active deflector systems are chosen, they should be planned on a sound technical/commercial basis to guarantee satisfactory results and service.



Back-to-back equipment housings around the transmitter mast contain blocks AD1 and AD2 in diagram below, one for each area.

Power is supplied to the transmitter by overhead 55V a.c. wires from nearby farmhouse and by coaxial cable to the receiver.



Interfacing microprocessors

Hardware connections for common microcomputers

Microelectronics Educational Development Centre, Paisley College of Technology

by **J. D. Ferguson,** B.Sc., M.Sc., M.Inst.P., **J. Stewart and P. Williams,** B.Sc., Ph.D., M.Inst.P.

The general purpose interface board has been designed for direct connection to either of the Acorn systems or, via a suitable connector, to any microcomputer based on the 6502. This final part of the series gives connections details for the popular 6502 systems and outlines the modifications necessary for use with other microprocessor families such as the Z80.

A rack-mounted Eurocard system directly accepts the interface board and provides an excellent system for expansion in the laboratory or for industrial applications, and the availability of a bus-compatible 6809 c.p.u. board emphasises this. In general the interface can be directly adapted to most 6502 systems, and the similarity of the 6809 bus signals makes conversion for this processor straightforward. For three of the most popular microcomputers, Apple, Pet and Aim 65, only a cable and socket adaptor is required to change the order of the pins from one bus to another. Using the interface board with an Acorn Atom is even easier because the Atom can have an edge connector mounted directly on the printed circuit board and the pins can be adjusted for connection to an external unit. The alternative connections are shown in Fig. 1.

6502 systems are now common because this microprocessor is the basis of many microcomputers for the educational and personal computing market but the 6809 offers an internal hardware multiplier, 16-bit operations and an extended operating instruction set. Programming models of the two microprocessors are compared in Fig. 2 and the following points are of interest. The four pointer-registers in the 6809 (index (x and y) and stack (u and s)) are all 16-bit as is the program counter, and these are matched by internal 16-bit arithmetic functions making this device a convenient introduction to true 16-bit systems. The direct page addressing mode allows any page to be used in the same way as the zero-page of the 6502, i.e. for faster access (what might be called a floating zero-page). Other addressing modes such

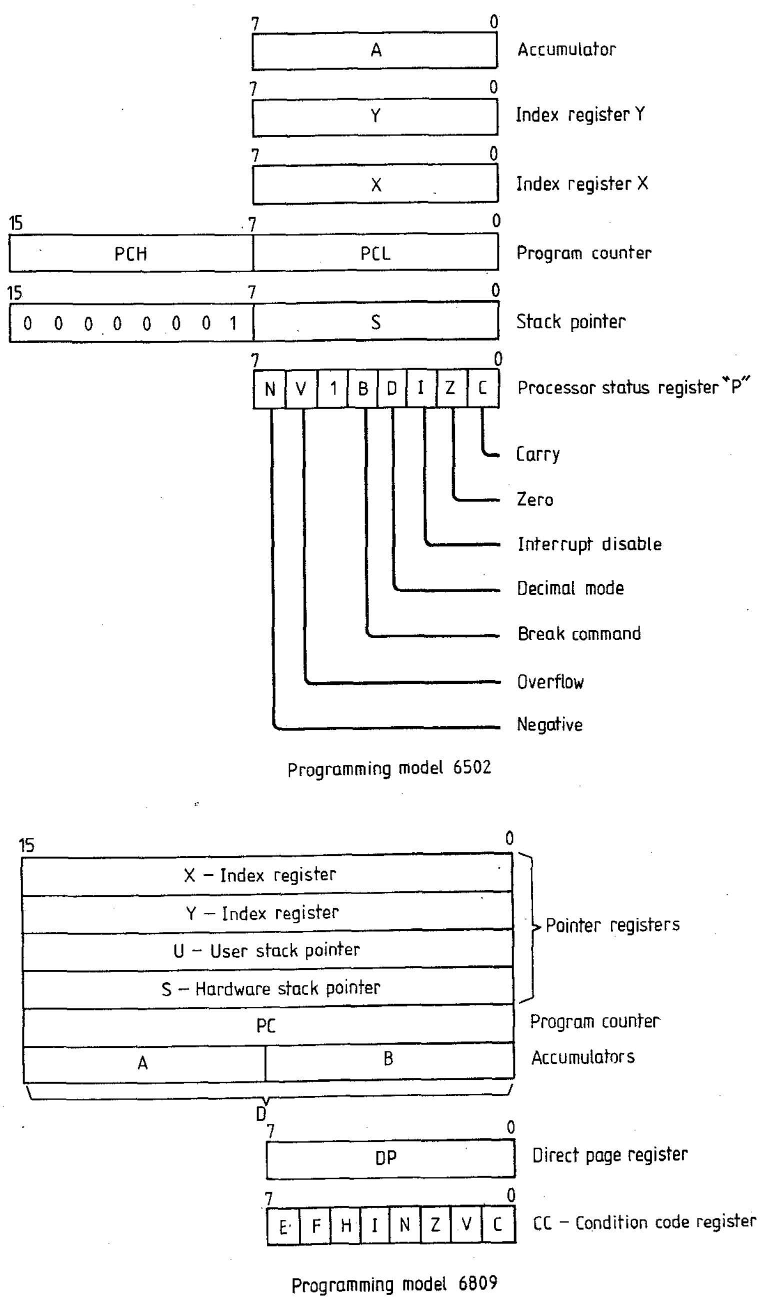


Fig. 2. *Programming models of the 6502 and 6809.*

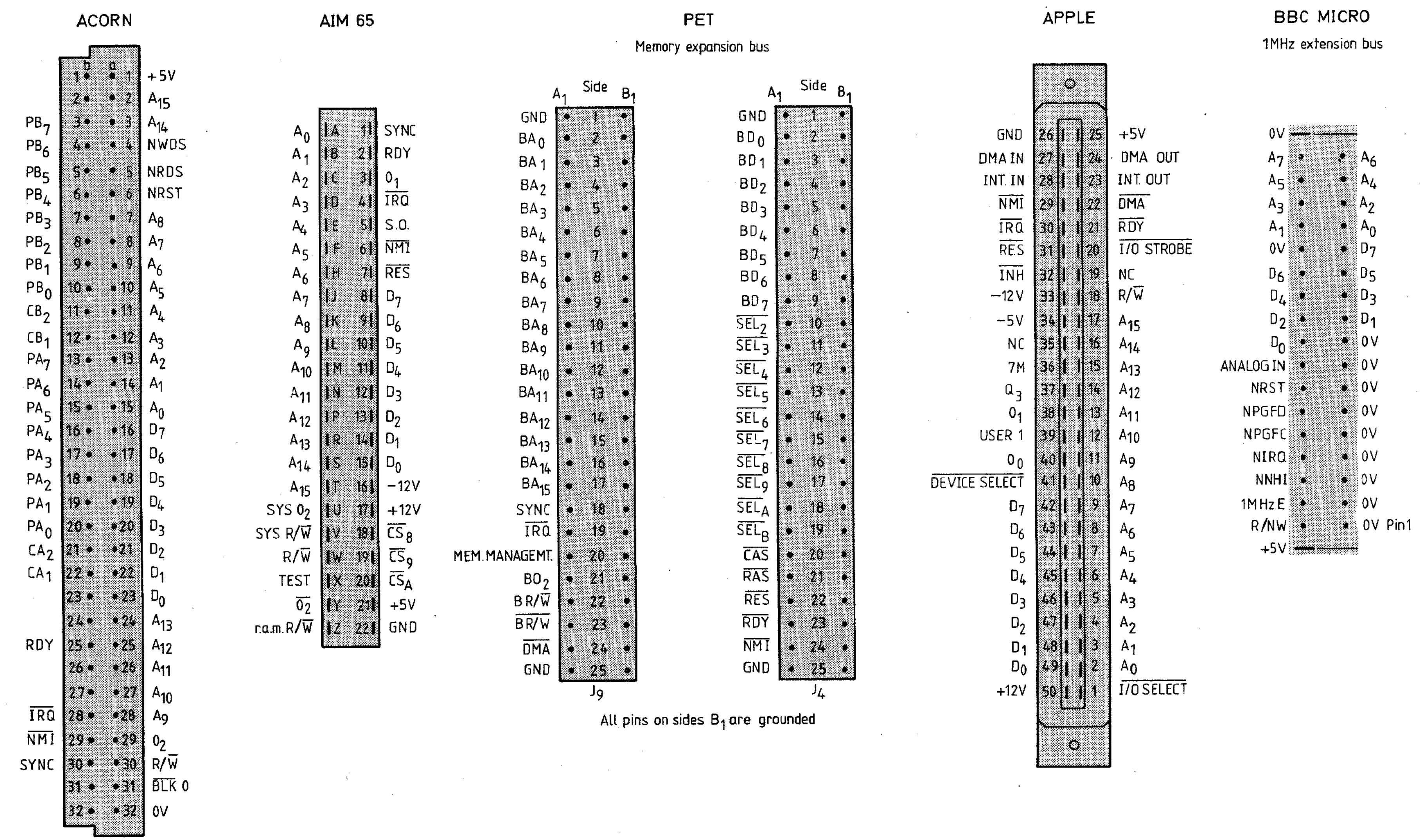


Fig. 1. *Connection details for common 6502-based microcomputers.*

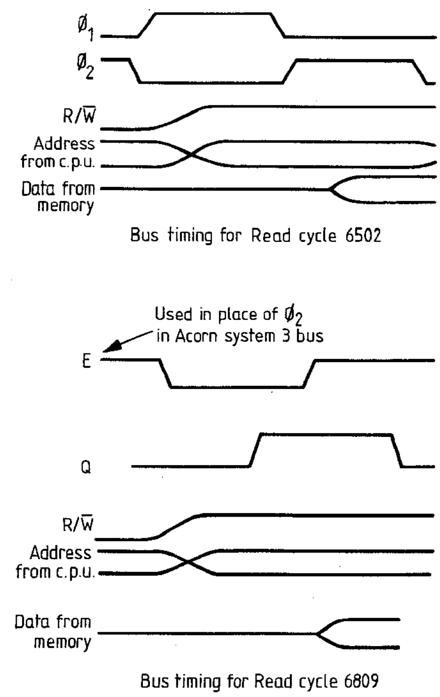


Fig. 3. *Bus timing for a Read cycle of the 6502 and 6809.*

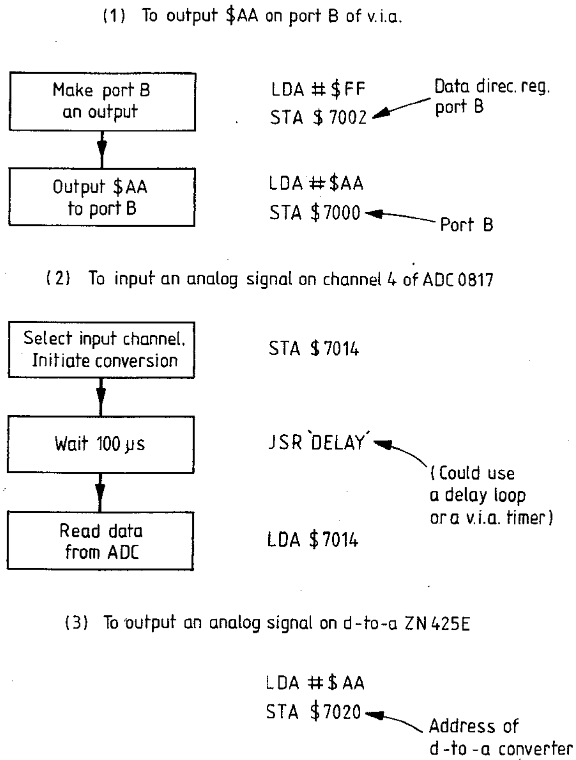


Fig. 4. *Simple software examples for driving the interface board with a 6809.*

as auto increment/decrement (including single or double steps) make it easy to step through tables of 8-bit or 16-bit data. Similarly, block-moves of data become easier as do software stacks. Even more important is the 6809's support of position independent code. Programs can be loaded anywhere in memory and run without re-assembly to a fresh origin. In addition, relative branching, both long and short, gives position independent transfer of control, and workspace on the stack gives position-independent temporary storage as an alternative to fixed r.a.m. locations.

For many applications including the important area of digital signal processing, an internal hardware multiplier is a great advantage. Two 8-bit numbers in the A and B registers are multiplied with the result appearing as a 16-bit number in A and B which is then used as a 16-bit register called D. The usual add, compare, subtract, etc. functions are all available in 8- and 16-bit form.

The 6502 retains certain advantages in applications such as industrial control where the necessary data manipulation is simpler and 8-bit arithmetic is sufficient, which enables programs to be shorter and faster. However, these advantages are lost if the control function is to be accompanied by extensive processing of the results. Tests were carried out on a 6809 c.p.u. card installed in an Acorn system 3 rack which uses the bus structures and timing relationships of the bus signals shown in Fig. 3. In the absence of an assembler, the routines were hand-coded and the basic board functions were exercised as shown in Fig. 4. The operating system with which the 6809 commonly runs is Flex, and the editor-assembler should permit efficient data collection and processing.

With unrelated families of microprocessors, connection to the interface board is more difficult. For the Z80, one problem is the higher clock frequency commonly used. At 4MHz neither the 6522 i/o device nor the a-to-d converter can cope. Other problems include the multiplexed bus of 8085 systems. Reducing the clock frequency may be feasible with new designs such as single-board controllers, particularly if the overall speed is uncritical. The existing clock frequency could be used with a divider to drive the board, but the timing relationships would remain difficult. The insertion of wait states can be applied to existing systems but this causes other problems. Driving the interface board from i/o ports which are used to generate address and data-bus equivalent signals is another possibility. The provision of a software clock signal is also feasible, but to sustain it over the period necessary to take the multichannel a-to-d converter through all of its channels would be a burden on the microprocessor. If a counter device is available, this could be used to generate the necessary clock signals. It is important to note that these compromises are far from ideal, but they do offer a means of using a board designed for one family of microprocessors, with another family.

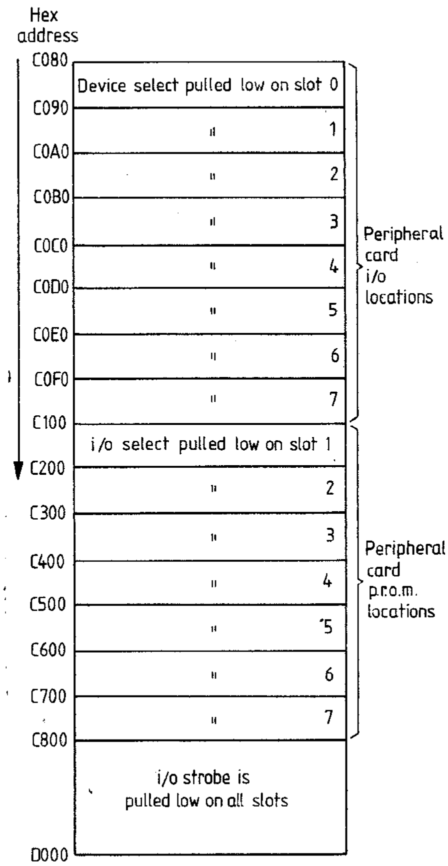


Fig. 5. *Apple slot memory assignment.*

BBC microcomputer

The BBC microcomputer designed by Acorn departs from Acorn's standard bus structure and provides two extension buses running at 1MHz and 2MHz. The 2MHz bus, called the tube, is planned for use with a second processor and expansion memory, while the slower bus is designed for specific peripherals such as an IEEE controller, daisy-wheel printer interface or Prestel modem. The 1MHz bus does not contain address lines A15 to A8, but page select signals FCxx and FDxx are provided instead (known as Jim and Fred on the prototype). Provision of these sig-

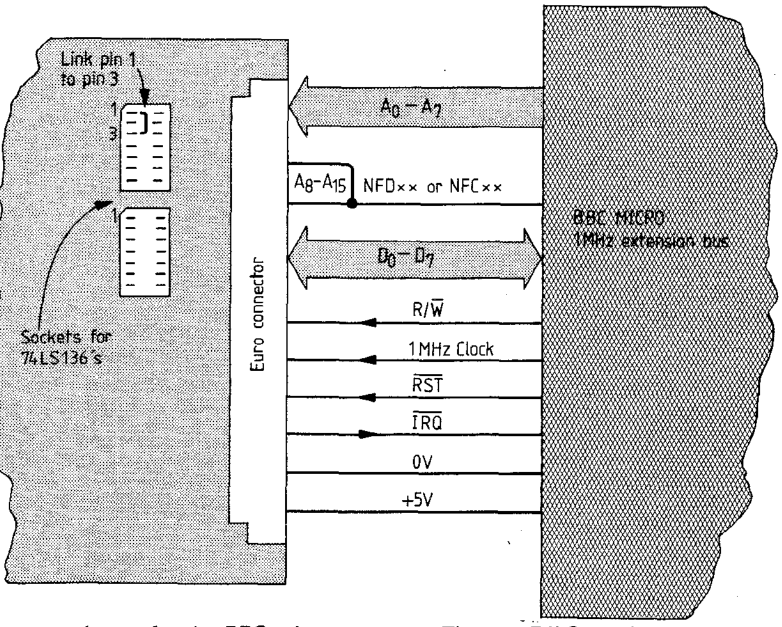


Fig. 6. *Bus adaptor for the BBC microcomputer. The two 74LS136 devices are removed and pins 1 and 3 are linked in one socket.*

nals allows the block and page select circuit on the interface board to be omitted. Fig. 4 shows one possible arrangement for an adaptor where the two 74LS136 i.cs and the block and page select switches have been removed. A link between pins 1 and 3 on one of the i.c. sockets carries the page select signal to the final stage of the address decode circuit on the board.

The board has so far been tested with the common 6502 microcomputers and has been successfully extended to the 6809 via the Acorn rack-mounted system. However, it is anticipated that it will be equally applicable to the BBC computer. Although there is a higher resolution a-to-d converter with this machine, it appears to be slower and restricted to four channels.

Apple 11

The Apple motherboard contains eight expansion slots (labelled 0 to 7) for a variety of peripheral interface boards. Each slot has defined memory locations associated with it which, if accessed, will generate device select and input/output select control signals for that slot, see Fig. 1. This arrangement eases the problem of address decoding when designing accessories specifically for the Apple. However, when using the general purpose interface board, these signals are not required because the board already has address decoding which selects its position in the memory map irrespective of which slot it is located in.

Fig. 5 is useful to ensure that the 64 bytes used by the board do not coincide with other cards installed in the Apple, e.g. if slot 4 is vacant the address-decode circuit could be set to block C, page 4. Building an adaptor for the Apple is simple because all the signals required by the board are present except for 0_2which can be replaced by 0_0.

Acorn system 3/Atom

There are no interfacing problems with either of these systems. The board can be mounted directly into the rack of the system 3, and can be driven by either the 6502 or 6809 c.p.u. card. The position of the board in the memory map will depend on the existing configuration of the system. A suitable position might be in block 0, page C, where the 6522 v.i.a. could be used as a parallel printer driver using software routines already present in the system 3 monitor.

As previously mentioned, mounting the board externally requires the modification of a Eurosocket wire-wrap connector.

Pet

Fig. 1 gives the pin arrangement of the memory expansion connector on later Pets. All the signals required by the board are provided except for +5V which can be found on the J3 connector, pin B2.

Cepstrum analysis

Theory, applications and calculation

by R. B. Randall B.Tech., B.A. and **J. Hee,** M.Sc. Brüel & Kjaer

The cepstrum is a spectrum of a logarithmic (amplitude) spectrum. It can be used for detection of any periodic structure in the spectrum, from harmonics, sidebands, or the effects of echoes. Effects convolved in the time signal (multiplied in the spectrum) become additive in the cepstrum, and subtraction there results in a deconvolution. After a discussion of the basic theory, this article describes applications of the cepstrum, including the study of signals containing echoes (seismology, aero engine noise, loudspeaker measurements), speech analysis (formant and pitch tracking, vocoding) and machine diagnostics (detection of harmonics and sidebands). Calculations need an FFT analyser and a desk-top calculator.

First defined as the power spectrum of the logarithmic power spectrum[1] in 1963, the cepstrum was proposed to determine the depth of the hypocentre of a seismic event from the echoes in seismic signals. The reason for defining it in this way is not clear; in the original paper it is compared with the autocorrelation function which can be obtained as the inverse Fourier transform of the power spectrum. Later, another definition was given as the inverse Fourier transform of the logarithmic power spectrum, thus making its connection with the autocorrelation clearer. At about the same time, another cepstrum-like function was defined as the inverse Fourier transform of the complex logarithm of the complex spectrum[2] and to distinguish it from the above cepstra it was called the complex cepstrum, while they were renamed power cepstra. Reference 3 contains a good discussion of the definitions and properties of the various forms of the cepstrum, and a guide to some of the applications. This paper summarizes that material, adds newer applications and indicates how the cepstrum calculations can be performed using a modern FFT analyser in conjunction with a desk-top calculator. This is a relatively low-cost system which has the power and speed of a minicomputer based system but which is more flexible in its uses and gives more direct contact with the signals being analysed.

Applications of the cepstrum can be divided into the purely diagnostic, such as the determination of an echo delay time, or sideband spacing from the position of a peak in the cepstrum, and those involving editing, where by removal of certain components in the cepstrum it is possible to remove information about their causes. This would include removal of the effects of echoes from a spectrum or time signal, or in speech analysis removal of voice effects leaving only the resonances of the vocal tract formants. For the diagnostic applications, either definition of the power cepstrum may be used, whereas for the applications involving editing it is essential to use the definitions based on the inverse Fourier transform. Where it is desired to return to the time function or include phase information in the frequency spectrum the complex cepstrum must be used.

The applications discussed include the processing of signals containing echoes (seismic and underwater signals, measurements on loudspeakers in a reverberant environment, aero engine noise including ground reflections, measurement of the properties of a reflecting surface) speech analysis (formant and voice pitch tracking, vocoding and speech synthesis) and machine diagnosis (determination and monitoring of families of harmonics and sidebands in gearbox and turbine vibration signals). A more mathematical application is in calculating the minimum phase spectrum corresponding to a given log amplitude spectrum (i.e. a Hilbert transform). This could have application to loudspeakers, where minimum phase characteristics are often desired, and comparison between actual and ideal phase characteristics could be made.

Terminology

The name cepstrum is derived by rearranging the word spectrum and was proposed in the original paper along with a number of similarly derived terms. The reason was presumably that the cepstrum is a spectrum of a spectrum, but of course the same applies to the autocorrelation function, and the really distinctive feature of the cepstrum is the logarithmic conversion of the original spectrum. Even so many of the original terms are still found in the cepstrum literature the most common of which are

cepstrum	from	spectrum
quefrency		frequency
rahmonics		harmonics
gamnitude		magnitude
saphe		phase
lifter		filter
short-pass lifter		low-pass filter
long-pass lifter		high-pass filter

Not all the above terms are necessary or useful; Quefrency, for example, is now well established as the x-axis of the cepstrum, though it is identical with time. In principle there is no difference between quefrency and the τ of the autocorrelation function. Even so, it is useful to speak of a high quefrency as representing rapid fluctuations in the spectrum i.e. small frequency spacings and low quefrency for more gentle variations. It can also be useful to distinguish between the rahmonics in the cepstrum and the effect in the cepstrum of a family of harmonics in the spectrum.

Basic theory

Using the terminology $\mathcal{F}\{\}$ to indicate the forward Fourier transform of the bracketed quantity, the original definition of the cepstrum is

$$c(\tau)=|\mathcal{F}\{\log F_{xx}(f)\}|^2 \qquad (1)$$

where the power spectrum of the time signal $f_x(t)$ is

$$F_{xx}(f)=|\mathcal{F}\{f_x(t)\}|^2 \qquad (2)$$

The new definition of the power cepstrum is

$$c_p(\tau)=\mathcal{F}^{-1}\{\log F_{xx}(f)\} \qquad (3)$$

while the autocorrelation function is

$$R_{xx}(\tau)=\mathcal{F}^{-1}\{F_{xx}(f)\} \qquad (3a)$$

A further useful definition of the cepstrum is the amplitude spectrum of the logarithmic spectrum, or

$$c_a(\tau)=|\mathcal{F}\{\log F_{xx}(f)\}| \qquad (3b)$$

This can be interpreted as the square root of equation (1) or as the modulus of (3) as for a real even function such as a log power spectrum, the forward and inverse transforms give the same result.

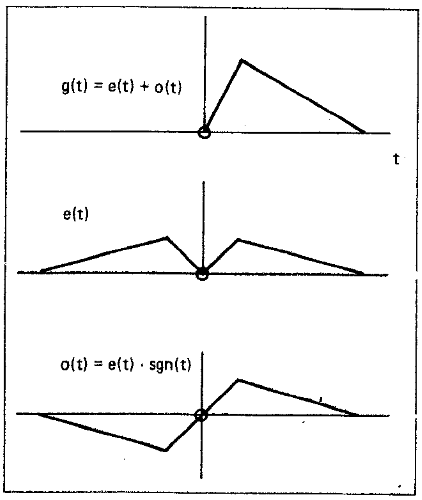


Fig. 1. Division of a causal function into even and odd components.

The complex cepstrum may be defined as

$$c_c(\tau)=\mathcal{F}^{-1}\{\log F_x(f)\} \qquad (4)$$

where $F_x(f)$ is the complex spectrum of $f_x(t)$ i.e.

$$F_x(f)=\mathcal{F}\{f_x(t)\}=a_x(f)+jb_x(f)$$
$$=A_x(f)e^{j\phi_x(f)}$$

in terms of real and imaginary components and amplitude and phase. From this the complex logarithm of $F_x(f)$ is

$$\log F_x(f)=\log A_x(f)+j\phi_x(f) \qquad (5)$$

Where $f_x(t)$ is real, as is normally the case, then $F_x(f)$ is "conjugate even", i.e.

$$F_x(-f)=F_x(f)$$

from which follow

$$\left.\begin{array}{ll} a_x(f) & \text{is even} \\ b_x(f) & \text{odd} \\ A_x(f) & \text{even} \\ \log A_x(f) & \text{even} \\ \phi_x(f) & \text{odd} \end{array}\right\} \qquad (6)$$

From equations (5) & (6) it follows that log $Fx(f)$ is also conjugate even and thus its inverse transform, the complex cepstrum, is a real-valued function despite its name.

For calculation of the complex cepstrum the phase function $\phi_x(f)$ must be continuous rather than the principal values modulo 2π, and this "unwrapping" of the phase spectrum can present problems in many practical situations[3]. From equation (2), $F_{xx}(f) = A^2_x(f)$ and thus the power cepstrum of (3) is virtually the same as the complex cepstrum of (6) for functions whose phase $\phi_x(f)$ is identically zero.

Another practical problem is whether the power spectrum should be one-sided or two-sided in frequency. In cases involving editing and transformation in both directions the two-sided spectra should presumably be used, but for some diagnostic applications it is advantageous to use one-sided spectra (negative frequency components set to zero). The theoretical background for this is tied up with the theory of Hilbert transforms and so a brief introduction follows.

From the general theory of Fourier transforms[4] the spectrum of a real even function is real and even, and of a real odd function is imaginary and odd. As any real function can be divided into even and odd components it follows that the real part of the Fourier transform comes from the even part of the time signal and the imaginary part from the odd part of the time signal. In the particular case of a causal time signal, i.e. one equal to zero for negative time, a special situation arises. As illustrated in Fig. 1 the even and odd components must be identical for positive time that they will cancel and give zero for negative time. Thus the even and odd compo-

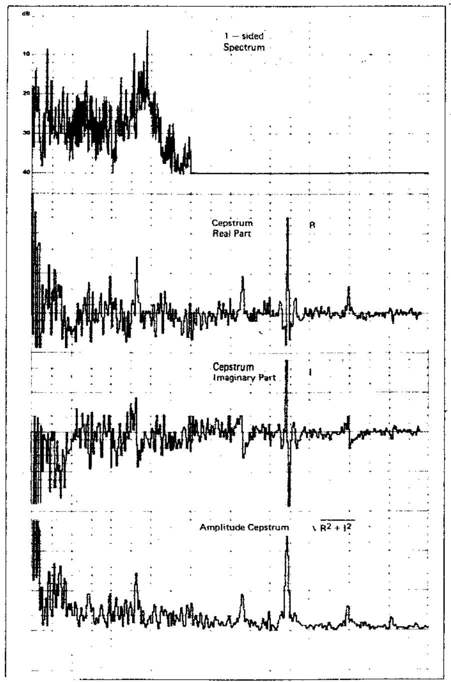


Fig. 2. *Cepstrum calculation procedure for a one-sided spectrum.*

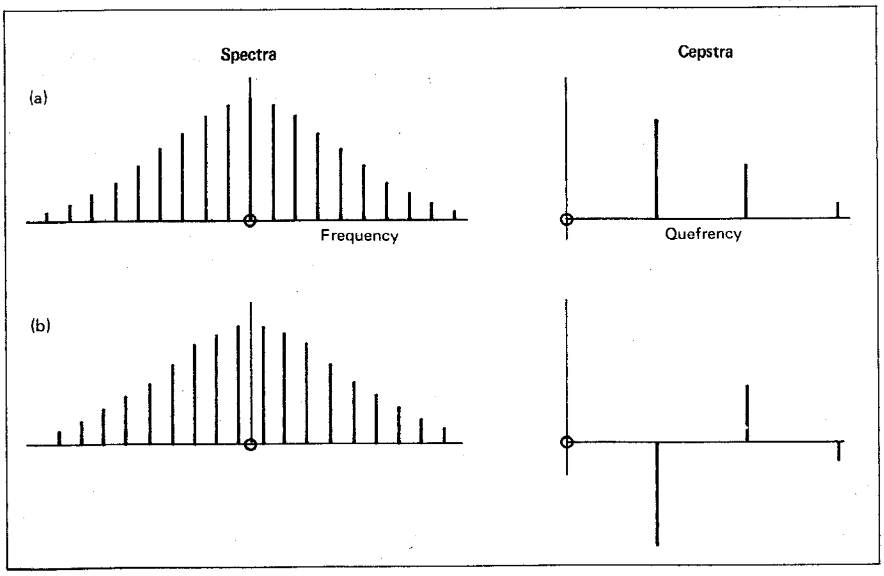


Fig. 3. *Cepstra of harmonic series (a) and odd harmonic series (b).*

nents are related by the sign function and the real and imaginary parts of the Fourier transform are no longer independent. The imaginary part can be obtained from the real part by convolution with the Fourier transform of the sign function, viz. a hyperbolic function in the imaginary plane. This convolution constitutes the (inverse) Hilbert transform, which is thus the relationship between the real and imaginary parts of the spectrum of a causal function or more generally of any one-sided function.

For minimum phase functions the log amplitude and phase spectra are also related by the Hilbert transform[4]. It follows directly that the time signal obtained by inverse transforming the complex spectrum having log amplitude as real part and phase as imaginary part must be causal, i.e. the complex cepstrum of minimum phase functions is right-sided only and is identically zero for negative time (quefrency).

Thus one way of deriving the minimum phase spectrum corresponding to a given log amplitude spectrum is to first calculate the power cepstrum according to equation 3. This is a real even function but can be considered as the even part of the (one-sided) complex cepstrum of the equivalent minimum phase function, which can thus easily be derived by doubling the positive quefrency values and setting the negative quefrency values to zero. A forward transform of this cepstrum will thus have the original log amplitude spectrum as real part, and the desired phase spectrum as imaginary part. An example of this for a loudspeaker is given later.

Returning to the question of whether the power cepstrum should be obtained from a one-sided or two-sided power spectrum, Fig. 2 shows the result of forward transforming a one-sided spectrum. The real part of this transform comes from the even part of the original function by analogy with Fig. 1 and thus the true cepstrum of the two-sided spectrum can be simply obtained by doubling the real part and discarding the imaginary part, see Appendix B1. On the other hand, the imaginary part will be the Hilbert transform of the real part and thus has interesting properties. There are zero crossings in the imagi-

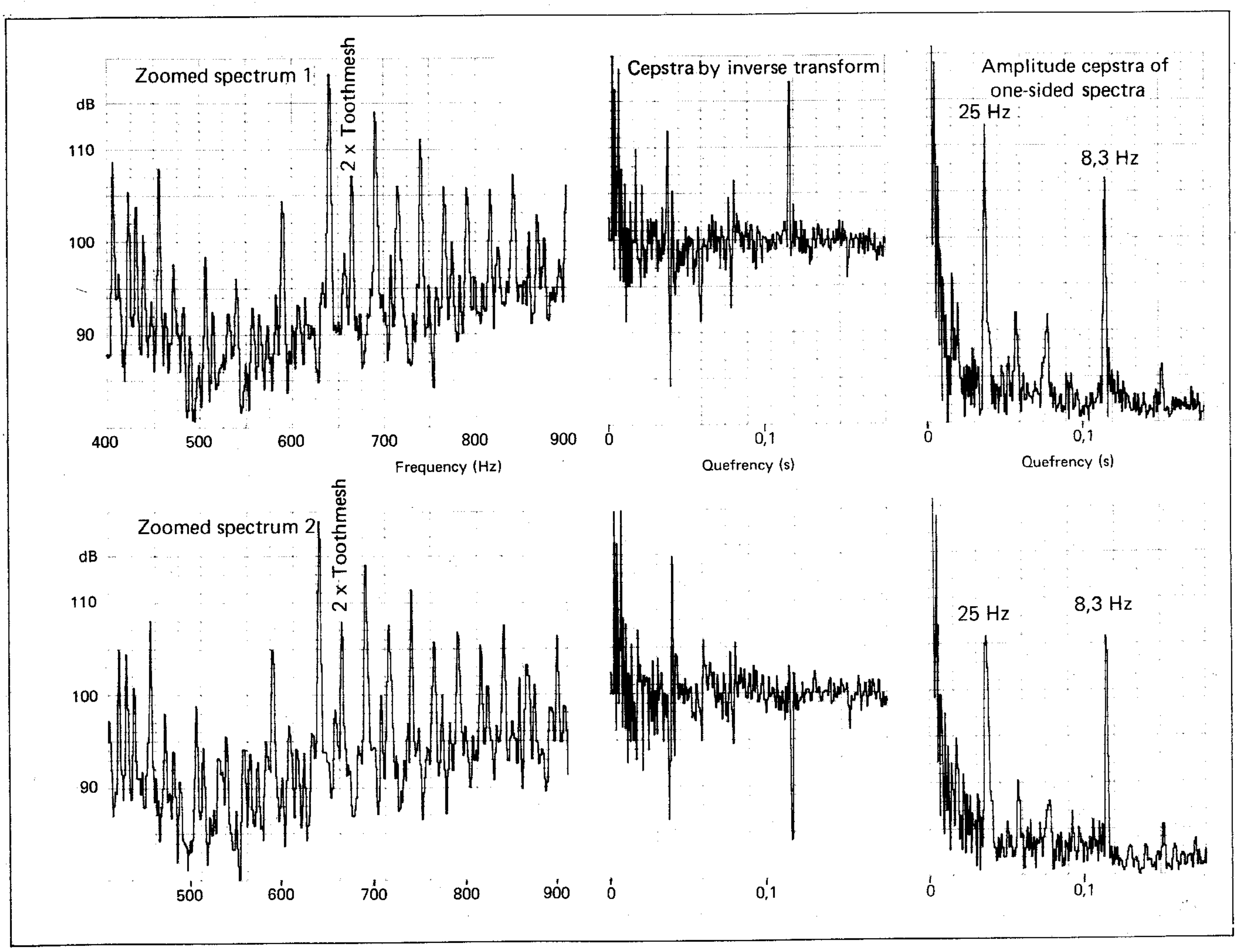


Fig. 4. *Cepstra of slightly displaced zoom spectra.*

nary part corresponding to the main peaks in the real part, these peaks corresponding to the spacing of the sidebands in the original spectrum. The peaks are positive and in the real plane because the original sidebands fall at exact harmonic frequencies. Referring to Fig. 3(a) this shows that the cepstrum of an harmonic series is a series of positive rahmonics. On the other hand if the periodic components in the spectrum are displaced a half spacing, e.g. a series of odd harmonics Fig. 3(b) the cepstrum consists of an alternating series of rahmonics with the first one negative. Halfway between these two situations the true cepstrum peak would be at right angles to the real plane and would thus correspond to a zero-crossing in the real part of the cepstrum (but a peak in the imaginary part which is its Hilbert transform).

To summarize, whenever the periodic structure of the spectrum does not correspond to a true harmonic series (in principle including zero frequency) the best version of the cepstrum to use is the amplitude cepstrum of the one-sided spectrum. The peak in this function will always indicate the true spacing of components in the spectrum, independent of their displacement along the frequency axis. A situation where this is relevant is in the calculation of the cepstra of "zoom" spectra where the lower limiting frequency of the zoom range is interpreted as being zero frequency. Fig. 4 shows an example of cepstra calculated from slightly displaced zoom spectra from the same signal. The cepstra calculated according to eqn 3 vary considerably as a result of this slight displacement, whereas the amplitude cepstra of the one-sided spectra are much more similar and easy to interpret. On the other hand it would not be possible to edit in these amplitude cepstra and return to the power spectra.

Deconvolution

Several applications of the cepstrum involve deconvolution. Fig. 5 shows schematically how the output signal $f_y(t)$ from a physical system can be considered as the convolution of the input signal $f_x(t)$ and the impulse response $h(t)$ of the system

$$f_y(t) = f_x(t) * h(t).$$

By the convolution theorem this transforms to a multiplication in the frequency domain

$$F_y(f)=F_x(f).H(f) \qquad (7)$$

and by taking logarithms this multiplication transforms to a sum:

$$\log F_y(f)=\log F_x(f)+\log H(f).$$

Because of the linearity of the Fourier transform, this additive relationship is

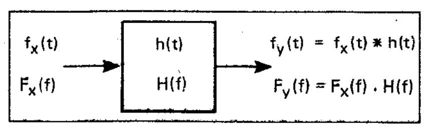


Fig. 5. *Input/output relations for a linear system.*

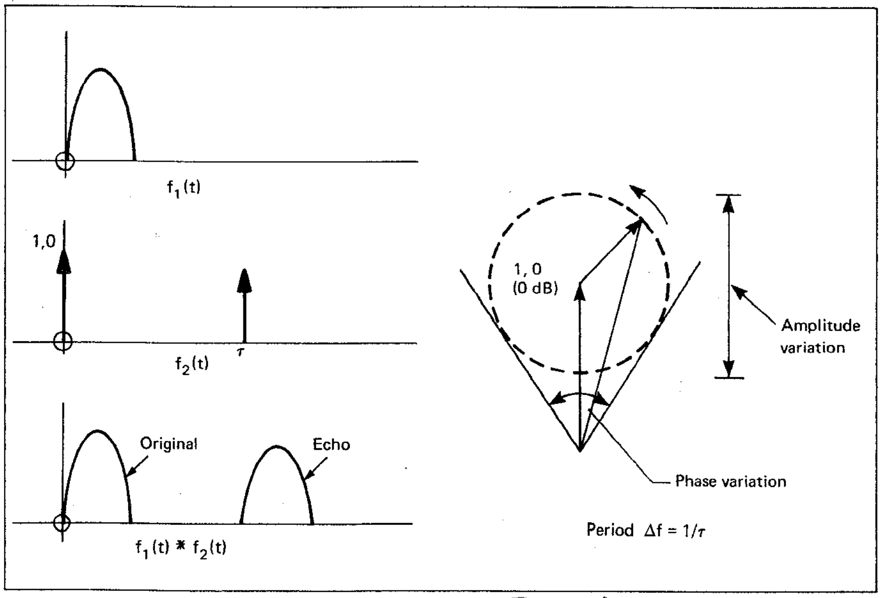


Fig. 6. *Modelling a signal with echo as a convolution.*

maintained in the complex cepstrum, i.e.

$$\mathcal{F}^{-1}\{\log F_y\}$$
$$=\mathcal{F}^{-1}\{\log F_x\}+\mathcal{F}^{-1}\{\log H\}.$$

By squaring the amplitudes in equation (7),

$$F_{yy}(f)=F_{xx}(f).|H(f)|^2 \qquad (8)$$

and so the same additive relationship also applies to the power cepstrum:

$$\mathcal{F}^{-1}\{\log F_{yy}\}$$
$$=\mathcal{F}^{-1}\{\log F_{xx}\}+\mathcal{F}^{-1}\{2\log|H|\}.$$

This means that if the effect of one of the factors (source or transmission path) is known in the cepstrum, then subtracting it there will result in a deconvolution in the time domain. As an example, echoes give a series of delta functions at known locations in the cepstrum. By subtracting these from the cepstrum all information about the echoes is removed and by transformation back to the spectrum, and even to the time signal if the complex cepstrum has been used, the echoes will also be removed there.

Note that the autocorrelation function would be obtained by inverse transformation of eqn 8 and that the multiplication there would transform to a convolution of the source and transmission path effects, as opposed to the additive relationship of the cepstrum. This represents one of the advantages of the cepstrum over the autocorrelation function. Another is that echoes are more readily detected in the cepstrum, in particular when the power spectrum is not flat[1].

References

1. Bogert, B. P., Healy, M. J. R. & Tukey, J. W., Quefrency analysis of time series for echoes: cepstrum, pseudo-autoco-variance, cross-cepstrum and saphe cracking, in Proceedings of Symposium on Time Series Analysis by Rosenblatt, M., (Ed.), Wiley, 1963, pp. 209-43.
2. Oppenheim, A. V., Schafer, R. W. & Stockham, T. G. Non-linear filtering of multiplied and convolved signals, *IEEE Trans. Audio & Electroacoustics,* vol. AU-16, Sept. 1968.
3. Childers, D. G., Skinner, D. P. & Kemerait, R. C., The Cepstrum: a guide to processing. *IEEE Proc.,* vol. 65, October 1977, pp. 1428-43.
4. Papoulis, A., The Fourier Integral and its Applications, McGraw-Hill, 1962.

Books

Operational Amplifiers

by I. E. Shepherd
318pp., hardback
Longman, £25.00

As the author points out, analogue circuits still have their uses and need a certain skill in circuit design to extract their full potential: they are not quite as simple as they appear. In this text, the author has adopted the practical approach, which is not to say that mathematics are avoided when they are helpful, but maybe of the more lengthy an laborious proofs are confined to appendices.

All the more important aspects of circuit design using op-amps are well covered, together with a section on applications and a complete chapter on active filters, after a preliminary discussion of the characteristics of op-amps and their use in general terms. Recent developments in technology, such as current-mirror inputs, micropower amplifiers and BIFET and BIMOS types are briefly described. A long list of references and a bibiliography are included.

PET Interfacing

by J. M. Downey and S. M. Rogers.
262 pp., paperback. Prentice-Hall International, £11.85.

Many users of the PET microcomputer have found the need to use peripheral devices on input and output for which no interface exists. This book explains the use of three ports on the PET – the user, memory expansion and IEEE 488 ports – to connect the peripherals by means of simple interface circuits. The use of the three ports is thoroughly described and several examples are given for each. All PETs with the 25 by 40 character display will accept the interfacing information given in this book. An IEEE 488 printer interface is presented and there is some information on graphic plotting using d-to-a converters.

Clandestine radio – the early years

The underground networks proliferate

by **Pat Hawker,** G3VA

Pat Hawker continues the history of wartime, covert radio operations. This second and final part takes the story towards the end of World War II.

A major problem for the clandestine transmitters was power to run them. Valve equipment, well before the semiconductor era, was difficult to run for long from dry batteries. Heaters and filaments of the 1930s consumed considerable power, and replacement batteries for either h.t. or l.t. were heavy and in short supply. Mains supplies were unreliable and not available for Maquis/partisan type operations; for urban working, selective power cuts in different districts, apartment blocks etc could be (and was) used by the ORPO to trace locations. The ORPO could also d/f on radiation from receiver oscillators.

A variety of methods were used, mostly with vibrator-type power units running from 6V vehicle batteries. To keep these charged, the static pedal-cycle generator was often used, since bicycles were available, and up to 100 watts can be generated by an energetic "cyclist" for short periods – and rather less over quite long periods. In Western Europe, the Mark VII had two separate power units, one for mains, the other for 6V batteries; the B2 and B2 Minor had combined power units. SOE also used (though only on a limited scale in Western Europe) such techniques as hand generators, wind generators, petrol-driven generators and even steam-driven generators. The steam generator used a boiler suspended in a brazier, coupled to a twin-cylinder steam engine and could charge a 6V battery at 4A. For use in the Far East, SOE developed a folding "beach chair" generator that could be folded into a back pack. When required the user sat in the deck-chair and pedalled.

The "control" stations in the UK used high-performance receivers with the operators having access to transmitters ranging from about 100-watts to 1kW, usually at a moderate distance from the receiving station. By modern standards, or even by the standard achieved by 1944-45, the early control stations were primitive, particularly in depending upon very simple aerials: it was not until later that rhombics and vee directional receiving aerials were put up. In this matter, as in so many others, Special Communications were forced to give priority to interception of enemy traffic rather than to the clandestine links – a logical requirement in view of the value of the Sigint product, but tough on the agents trying to get their weak, low-power signals through to "London" with the skilled ORPO d/f teams closing in.

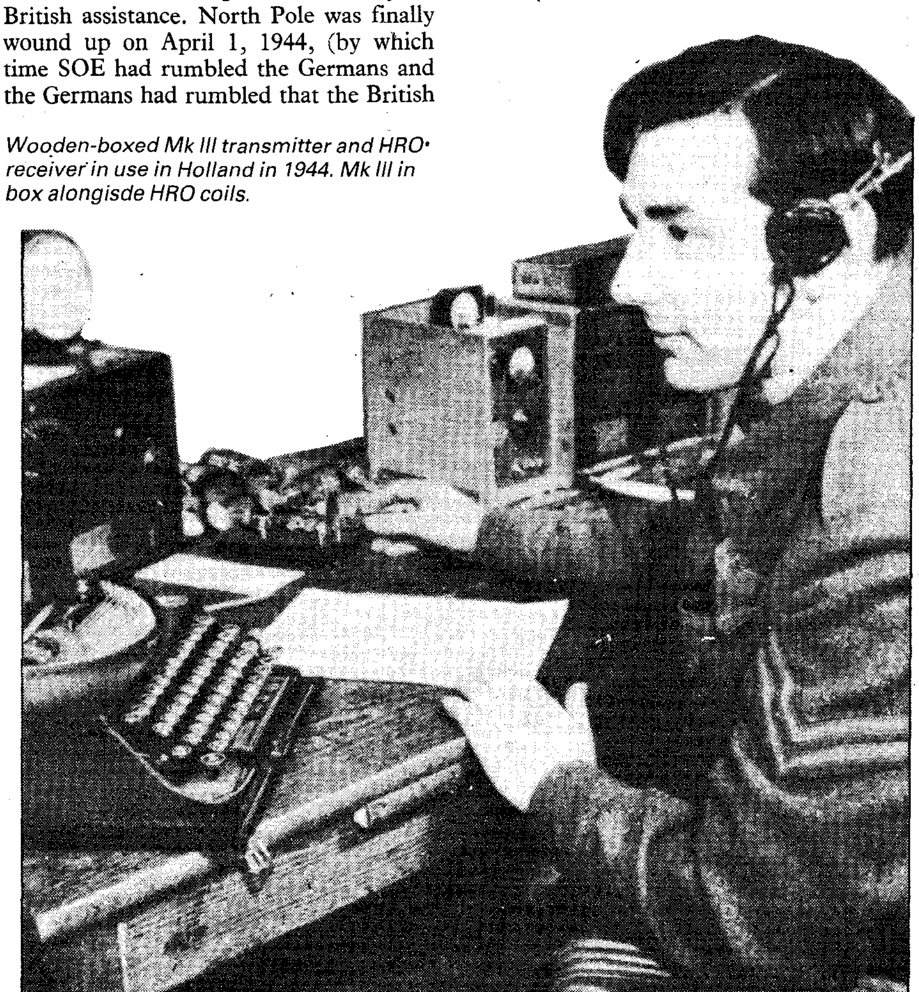
By 1944, things had improved. SOE, for example, had a 40-position station at Poundon equipped with HRO, AR88 and Marconi CR100 receivers and with broadband, high-power linear amplifiers that made possible virtually instantaneous frequency changing and simultaneous operation on several circuits – a considerable technical achievement. For SOE, but not for Special Communications, many of the base operators were girls, often enrolled in the First Aid Nursing Yeomanry (FANYs) whose role in so many covert organizations had little to do with nursing!

Initially most of the West European links were organized, controlled and equipped from the UK, but as the war progressed, several Resistance groups, including the Dutch and the Danes, increasingly developed their own ideas, as indeed did the Poles from an early stage, adopting such techniques as high-speed "squirt" automatic keying which reduced the time on air but required higher-power transmitters.

In Holland, the debacle of Operation Northpole (Englandspiel), a well-organized German funkspiel or radio game that ran from 1942-44 and resulted in the capture and subsequent execution of almost 50 young Dutchmen sent in by SOE, left Dutch resistance organizations wary of British assistance. North Pole was finally wound up on April 1, 1944, (by which time SOE had rumbled the Germans and the Germans had rumbled that the British had rumbled them) by a series of plain-language messages sent to London over six German-controlled radio links. The two main Dutch Resistance groups – OD, which drew its main support from former Dutch army groups and the right-wing; and RVV, representing left-wing political groups – began themselves to prepare for the internal radio links they foresaw would be needed during the eventual liberation of their country – though they did not foresee that Holland was to remain cut in two, following the tragic failure of the airborne 'Market Garden' operation at Arnhem, throughout the bitter winter of 1944-45. OD and RVV were loosely linked together through the Netherlands Intelligence Department (BI) but their activities remained separate.

Between September 5, 1944 to the end of the European war in May 1945, the two clandestine radio networks (Binnenlandse Radiodienst) in the north (occupied) zone, transmitted or received over 120,000 cipher groups, working to two separate control stations in Eindhoven (RVV had a back-up station in Nijmegen) on frequencies between 2700-3200kHz. Unfortunately the operation had been planned in the belief that the stations would only be on

Wooden-boxed Mk III transmitter and HRO receiver in use in Holland in 1944. Mk III in box alongisde HRO coils.



the air for a few days or at most weeks. The locally-built transmitters, drawing on the Philips Eindhoven factories, used relatively high power, typically 70-100 watts from push-pull, self-excited power oscillators, crudely disguised as medical diathermy equipment. They were usually installed in remote farm-houses, using rotary converters and heavy vehicle batteries. The low frequencies demanded longish, outdoor aerials and the stations were bulky and difficult to transport.

The closing months of 1944, supported by a private telephone link to the occupied zone over the power supply network, and also by several links direct with the British Special Communications stations in the UK, brought a stream of information to Dutch Intelligence. The Dutch clandestine operators included several ex-marine and ex-Service operators and also a number of former radio amateurs, but only a few had any previous experience of the dangers of covert operation. In Eindhoven, British assistance was accepted from January 1, 1945 but rather reluctantly (the British operator for the RVV base station was asked not to come in on the day that Prince Bernhard visited the station!) and BI would not heed advice that the stations were staying on the air too long.

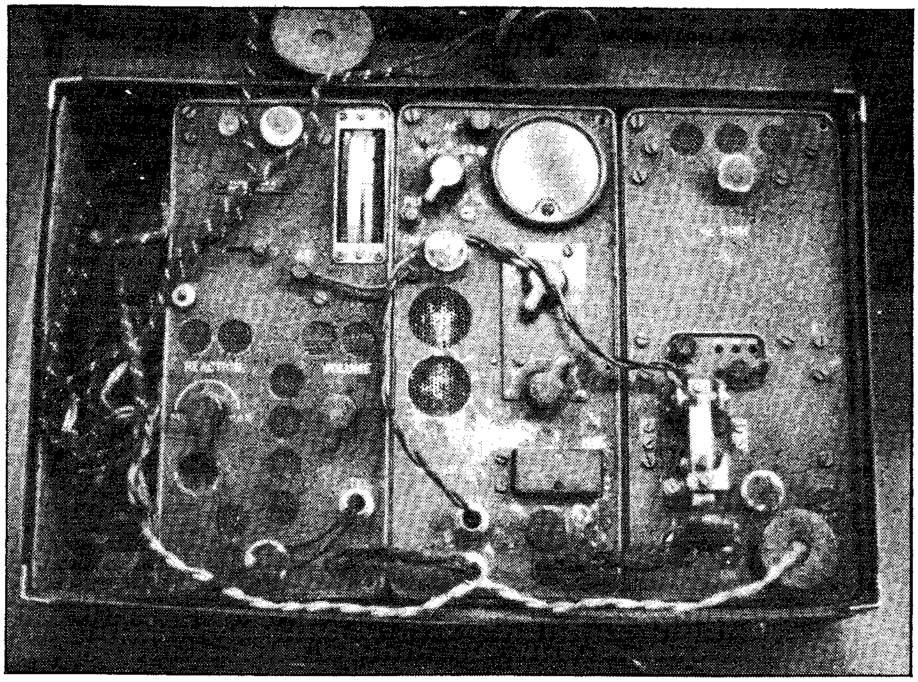
In January/February 1945 disaster struck. In three weeks, eight transmitters, mostly with several operators apiece, were lost in a series of German raids. Some of the operators were executed on the spot, some in front of their families, some together with their families; some were executed after imprisonment; some perished in the final holocaust of the concentration camps. For a period, virtually all traffic with Eindhoven was cut. But one who survived was the main RVV operator in Amsterdam, who had already had considerable experience working across the North Sea to Buckinghamshire. A professional operator, his traffic to Eindhoven on occasions *averaged* over 27 cipher groups per minute over several messages – a rate of transmission that could seldom, if ever, have been exceeded on other manually-operated World War II clandestine links.

In March/April more volunteers came forward and several of the links were re-established. It was over the Amsterdam link that news first came through that the Germans were willing to negotiate a surrender in Holland, and this was carried out on the link now finally operating with German connivance.

How successful?

There can be little doubt that the German Abwehr/RSHA Intelligence services would have been vastly more effective if they had never heard of h.f. radio! At every turn they were frustrated by the ability of the Allies to read over their shoulder, and what Bletchley Park did not discover of their activities form their own messages they were able to ferret out through the XX Committee form the several dozen double-agents using w/t to communicate with their supposed German masters.

But what of our own activities? The Germans also successfully ran



Early French Resistance suitcase set in three units, now in Toulon museum. Possibly Mk IV.

"controlled" agents not only for the Englandspiel that ended in the tragic deaths of almost 50 young Dutchmen, but also, less spectacularly, by penetrating and "playing back" some French Resistance and SOE Section F links. The Germans similarly derived much information from their cryptanalysts (including the illuminating reports of the American diplomats in Cairo) though there is little evidence that they ever broke into MI-6's hand ciphers, including the less than ideal agent-transposition ciphers based on memorized "keys" in the form of poems, etc. Not that MI-6, in its cryptography, was always beyond reproach – important operational plans were occasionally sent in insecure ciphers by officers unwilling to wait until the messages could be enciphered on a one-time pad.

But MI-6, unlike SOE or German Intelligence, did make a serious effort to keep codes and signals operators apart – a practice that may have at times made for bored radio operators but did make it less easy to set up and run a successful funkspiel.

It could be argued that few organizations really understood how to use clandestine radio and radiomen well: does not Smiley (in John Le Carré's "Tinker, Tailor, Soldier, Spy") say: "We all have our prejudices and radiomen are mine. They're a thoroughly tiresome lot in my experience, bad fieldmen and overstrung, and disgracefully unreliable when it comes down to doing a job".

Harry Rée, one of F Section's best "organizers", has put it rather differently: "I hadn't a radio operator with me, and to be quite honest I didn't want one. It would have been an added responsibility, and added risk. They had a terrible time . . . . The sets had to be sent from England, they were terribly heavy in those pre-transistor days. They were built into big harmless-looking suitcases. The poor devils of operators couldn't stay for long in any one place for fear of being detected by radio vans, so they had to hump those great things around. It was a very dangerous, but also a terribly dull job."

While one may question the extent to which anything that is terribly dangerous (and calls for skill) can at the same time be terribly dull, there is logic in the view that many SOE "circuits" would have gained rather than lost effectiveness if they had had no radio.

Sir Colin McV Gubbins, who controlled the SOE radio activities, claimed that without h.f. Morse links, SOE would have been "groping in the dark".

Perhaps so, but who can contemplate unmoved, the thought of F Section sending the brave, but dreamy and inadequately trained, Indian princess Noor Inayat Khan ("Madeleine") into the dangerous Paris area where she kept, in a school exercise book, both in plain language and in cipher, every message that she sent or received – a prize indeed for the Germans.

When any agent, well or indifferently trained, well or indifferently equipped with radio, is infiltrated into enemy-occupied territory, any of a number of things may happen. He or she may succeed in the mission and, at least for a time, remain at liberty and on the air; the agent may believe he has succeeded, yet in fact be working under secret surveillance or transmitting messages stemming from the enemy who may have already secretly penetrated the network; the agent on arrival may be quickly captured, possibly being met by an enemy-organized reception party – he or she may then volunteer or be persuaded to act as a double-agent; the agent may in fact already be a double-agent using this means of returning to his masters or penetrating a Resistance network; he may reach is destination but then be unable to make radio contact with the control station, possibly due to faulty equipment, inexperience, loss of nerve or crystals (or he may have been given the wrong crystals for his signals plan!). Both sides endeavoured to guard against "controlled" radio games by providing security checks or even double security checks, by "fingerprinting" the Morse, etc. but no form of security check can, for example, be proof against the agent who determines to co-operate fully with the enemy. The radio game can, however, be made more difficult if the operator is deliberately distanced from both intelligence gathering and the cryptography.

It is difficult to assess with any degree of certainty how MI-6's radio compared with that of SOE, though there can be little doubt that some – though not all – of the Special Communication circuits did achieve remarkable (one is tempted to say curiously remarkable) successes, using (on the whole) less advanced technology. Over several years, MI-6 communications with the key Jade group in the Paris area handled may hundreds of messages; the Belgian "weather service" group were never shut down despite daily skeds; Norwegian communications were on the whole effectively maintained; as were the links with the Vichy police, the Polish-French Interalliée network (before this was heavily penetrated), and the large network run by "Colonel Remy" (Renault-Roulier). In some cases these networks were able to call on indigenous, experienced radio-operators; in others they avoided frequent carrying of conspicuous suitcases by having a number of different transmitters in prepared locations working the same signals plan. Even so one wonders if there were other factors involved – some of which may never be known. There has for instance been some speculation that the Jade group may have provided a link with the Schwartz Kapelle ("the black orchestra") a German anti-Nazi organization that included a number of Abwehr officers – was this the link that has been alleged to have existed between "C" (Sir Stewart Menzies), head of SIS, and Admiral Canaris, had of the Abwehr, one of the many who were executed following the July attempt on Hitler's life? What is more certain is that London penetrated and was kept informed of some of the activities of the German radio security teams and was able sometimes to warn their agents of pending disaster if they continued working to their signals plan.

Not all communications were by Morse. MI-6 made use of early American f.m. equipment on about 30MHz to speak directly from high-flying aircraft to Resistance groups (the system might have been more successful had not German military vehicles been using the same channels). SOE developed the 450 MHz "S-phone" super-regenerative transceiver, which was also used in conjunction with Rebecca and Eureka navigational aids. S-phones were used across the Dutch rivers and, for example, in August 1944 one hundred aircraft were flight-controlled using S-phones during a massive drop of supplies to Marshal Tito's partisans in Northern Yugoslavia. But Morse was the dominant mode including its use by the Poles and Danes for high-speed "squirt" transmissions – with the Poles running up to 1kW with pairs of 813 valves.

"Belgian met. service for the RAF" – group of Belgians who provided daily weather reports for several years. Station never closed down: not all group survived.

What about the Russians? The Sorge ring in Japan, the large "Red Orchestra" network in Western Europe and the Lucy ring in Switzerland were all eventually broken up largely as a direct result of intercepting and tracing the clandestine radios. For several years the Lucy ring provided Moscow with extremely accurate information. It has been suspected (but denied) that some of this information stemmed from Bletchley Park, with London using this means of sharing some ULTRA secrets with its Russian ally without revealing its source. The Englishman Allan Foote, one of the trusted clandestine radio operators for the Lucy group may have been an SIS "Z" penetration agent – and indeed may not have been the only Lucy operator working for more than one master. In the Russian networks the radio operator was traditionally nearer the centre of things than their British counterparts: they also learnt quickly the vital importance of having a sufficiently complex signal plan, although at first were often allowed or even encouraged to stay far too long on the air at each session.

Then again too many of the messages, transmitted at great risk, were either too verbose or unessential. At least one Section F operator refused to handle any more messages for a network that insisted that everything should be sent in full. There is still a view that equates communications efficiency with the data rate – regardless of the contents. The true art of clandestine radio is to make every word count: if there is nothing vital to communicate stay off the air; use a single-group "crack" signal rather than a 50-group message!

The technology

If radio was a double-edged weapon for Intelligence, what did its close involvement in covert activities do for the technology? In effect it led to professonal interest in what had previously been a largely amateur field – that of transportable, low-power communications equipment with a range of several hundred miles. By 1945, a whole new class of "military" equipment for long-range patrols and infiltration had begun to appear: very different from the tactical or strategic long-range military communications that had dominated British signals planning in the 1930s. Spies, "private armies" and underground resistance movements created a new form of radio. Sadly one has to admit that Resistance of the 1940s also fathered the techniques of the urban terrorists of the 1970s and 1980s: the plastic explosives, the assassination squads, the art of silent killing, the suicide pills, the hostage-taking. Only clandestine radio has been downgraded by the availability of the international telephone.

Equipments such as the crystal-controlled Mark III and B2 proved more dependable in use than the orthodox military No 19 sets; similarly the Americans found that by using equipment designed for amateur radio including the Hallicrafters HT4 transmitter (BC610) they were able to put together the outstandingly successful SCR299 series of signals vehicles (and again in the Vietnam war hurriedly adapted the Collins KWM-2 amateur transceiver for their Special Forces).

It also showed that the main problem was (and still is) the provision of electric power. Semiconductors have reduced power consumption but there is still a requirement on h.f., it can be argued, for a minimum of several watts of r.f. power if only to combat the ever-varying propagation conditions that often make it easier for signals to be heard a thousand miles away than it is to pass traffic, at fixed times of the day, in different seasons of the year, to a base station just 100 to 500 or so miles away, often within the "skip zone".

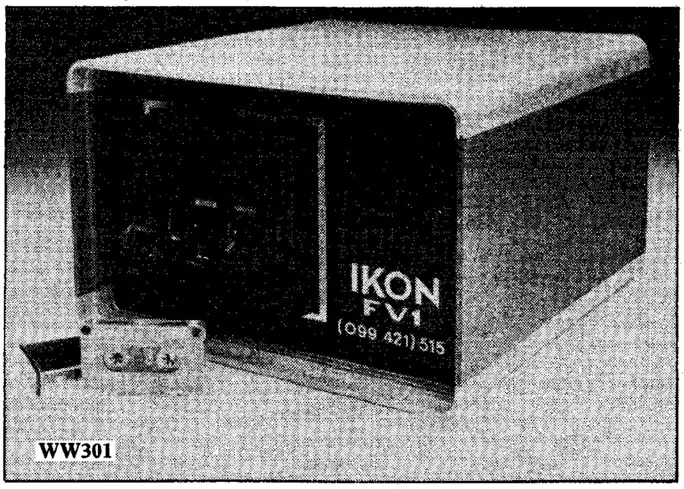
But above all, it showed that skill and experience as a radio operator was an essential key to success, provided that it was supported by the necessary deviousness and instinct for conducting covert operations on the part of the organization concerned. Some men and women, with little or no previous experience of radio communications, *did* acquire the necessary experience quickly, but successful h.f. radio operating requires more than an ability to send and copy Morse at some given rate of transmission. Added to this, they needed also the ability to live clandestinely in enemy-occupied territory without cracking under the stress, though liable to be "blown" at any time by those they were forced to trust, pawns in an infinitely complex game of chess. Far too many of those who volunteered to provide the radio links with England in 1941-45 lacked the training or experience they needed to survive. We remain in their debt.

New Products

Alternative storage medium

Baud rates from 110 to 4800 are selectable on the mini-cassette based FV1 storage medium from Ikon Computer Products. The unit has internal buffering and a microprocessor-based operating system to organize data, provide error checking and keep software control requirements from the host computer to a minimum. Up to 100K-bytes of data can be stored on one mini cassette – in a maximum of 104 files – and two files may be accessed simultaneously. RS232 communications with the computer are through a 7-pin DIN plug which also doubles as a baud rate selector. Data can be transmitted in blocks of between 1 and 99 bytes. The FV1 responds to 11 commands and can return one of 18 error codes. Apart from non-volatile, cheap data storage, the unit can also be used to transfer data from one computer to another.

Ikon Computer Products, Kiln Lake, Laugharne, Carmarthen, Dyfed, Wales.
WW301



WW301

Automatic multimeter

Among a number of new products recently introduced by Philips is the PM2521 automatic multimeter. This 4½ digit microprocessor controlled instrument is fully auto-ranging and gives readings of frequency, time, temperature, dB, diode forward voltage, resistance, alternating/direct voltage, a.c. and d.c. on a 4½ digit display. Voltage, current and resistance resolutions are 10μV, 10μA and 10mΩ respectively and the meter's basic accuracy is around 0.03%. When current measurements up to 20mA are made, an active circuit keeps the voltage over the input terminals to less 25mV by feeding a current in the opposite direction. The bandwidth for r.m.s. measurements is 100kHz. On direct-voltage, resistance, diode-check and dB ranges, the zero reading can be offset to a reference value. The 2521's price is £295 plus v.a.t. (in the UK). Among other instruments recently introduced by the same company are a 10Hz to 100kHz RC oscillator with 0.02% distortion at 1kHz and floating outputs, and a range of line conditioners. Extensions to their range of miniature thermocouple assemblies have also been made.

Pye Unicam Ltd, York St, Cambridge CB1 2PX.
WW302

Varactor amplifier

Low input-bias current and high input impedance, at ±10fA and $3\times10^{11}\Omega$, 30pF respectively, are the main features of Intech's inverting-only varactor amplifier. The AMP310J has a minimum slew rate of 0.4V/ms and its input noise-voltage figures are 10μV p-p for 0.01 to 1Hz and 10μV r.m.s. for 1 to 100Hz. Metal packaging is used and the device is pin-compatible with the A/D310J. In quantities of over 100, the price is £19.50 each.

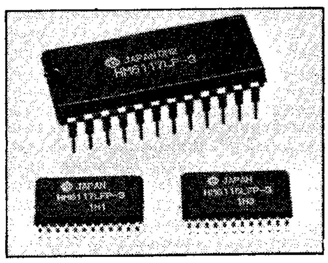
Teknis Ltd, Teknis House, Meadrow, Godalming, Surrey GU7 3HQ.
WW303

Small outline r.a.ms

The package used for these 16K static r.a.ms is 60% smaller in area and 50% thinner than conventional 24-pin d.i.l. devices. Because of the reduced size, Hitachi, the manufacturers, expect that there will be a strong demand for the i.cs in applications requiring dense component layouts such as pocket computers, point-of-sale terminals and other portable electronic equipment.

The HM6116FP/LFP and HM6117FP/LFP series memories will be available from Hitachi UK in the near future.

Hitachi Electronic Components (UK) Ltd, P.I.E. Building, 2 Rubastic Road, Southall, Middx UB2 5LL.
WW304



High-voltage electrolytics

Both radial and axial-lead LS series electrolytic capacitors from Matsushita are available through Compstock. These components can be obtained in values from 0.47μF to 10 000μF with working voltages ranging from 6.3 to 500V d.c. Other operational characteristics are a worst-case temperature range of −25° to +85°C, leakage current of $0.02CVs^{-1}+3\mu A$ and ripple current ratings of 10mA to 1100mA r.m.s. according to the capacitor's working voltage and value. Small size is a feature of the LS series – the 10 000μF, 6.3V type measures 18mm diameter by 40mm long.

Compstock Electronics Ltd, Compstock House, London Road, Stanford-le-Hope, Essex SS17 0JU.
WW305

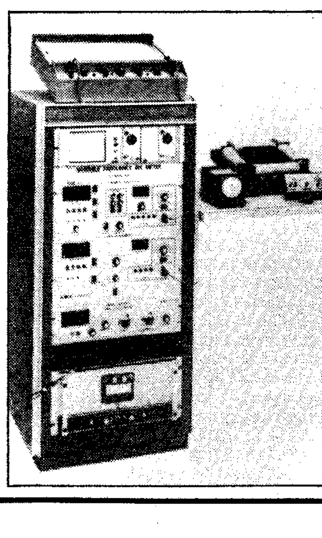
Heat-conducting epoxy

Electrically-insulating, thermally-conductive epoxy developed for bonding substrates to their packages is available from Epoxy Technology Inc. This black resin, described quaintly as 'somewhat flexible', is claimed to be suitable for bonding materials with dissimilar coefficients of expansion such as aluminium and alumina, and gold-plated Kovar and alumina. Epo-Tek H65 also adheres to other surfaces including those made from ferrous and non-ferrous metals, glass, ceramic and semiconductor materials. Curing takes about 30 minutes at 150°C and the resin can be subjected to 300°C intermittently without degradation, the manufacturers claim. A 3oz trial sample costs $20.

Epoxy Technology Inc., 15 Fortune Drive, PO Box 567, Billericay, Massachusetts 01821, USA.
WW306

BH meter

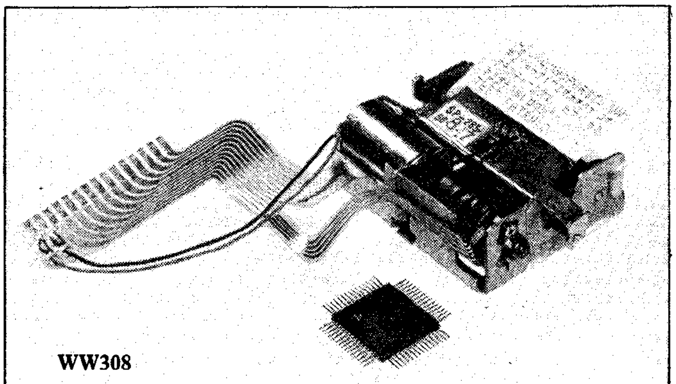
Induction and coercive force measurements can be made on soft magnetic materials such as ferrites, tape-wound cores, transformer laminations and special steels using the model 7000T variable frequency BH meter from LDJ Electronics. Frequencies from 10Hz to 10kHz (up to 20kHz is optional) or d.c. can be used for measurements. Fixed-frequency values, induction (B), coercive force (H) and permeability readings are shown on a 3½ digit panel



meter while an oscilloscope displays the BH curve and variable frequencies from 10Hz to 10kHz. The system was primarily designed for use in the laboratory but it can also be used in quality control applications.
LDJ Electronics Inc., UK Division, 4 Somerset Way, Semington, Nr Trowbridge, Wilts BA14 6LD.
WW307

Small thermal printer

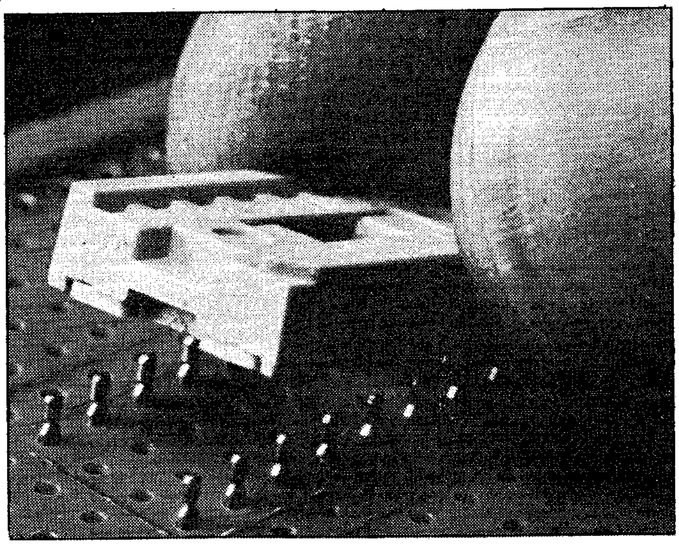
Two items make up the SP-285 kit from Roxburgh, a 16-column thermal printer for 38mm wide paper, and an 8-bit microprocessor controller i.c. The former prints at rates of up to two lines per second using a 5 by 7 matrix and the latter has a 64 ASCII character set and interfaces the printer to an 8-bit bus. For quantities of 99 upwards, the kit costs less than £17 per unit.
Roxburgh Printers Ltd, 22 Winchelsea Road, Rye, E. Sussex TN31 7BR.
WW308



WW308

Breadboarding kit

At the heart of 3M's breadboarding kit is a 24-contact plug strip which can be snapped off to the desired length, thus reducing the number of different components in the kit and simplifying ordering. These strips are plugged into the circuit board in parallel pairs to mate with a range of d.i.l. sockets which accept either the legs of an i.c. or a d.i.l. plug carrying a discrete component. On the underside of the board, the plugs have insulation displacement connectors to accept one or two wires. The basic kit includes a single-height Eurocard board, a selection of dual sockets, plug strips, solder strips, tools and 25ft of 30a.w.g. wire.
Electronic products group, 3M UK PLC, 3M House, PO Box 1, Bracknell, Berks RG12 1JU.
WW309



WW309

Radio-code clock

Model RCC8000 is a microcomputer controlled instrument which receives, decodes and analyses all time-coded standard frequency transmissions to provide an accurate, secure and automatic time/calendar or synchronization system. The standard unit offers a range of outputs including RS232, 16-bit parallel, f.s.k. for magnetic tape and pulsed or serial data for slave displays. A keypad programmer is also available which independently programs seven parallel output lines to switch at precise times/dates and for exact durations. Other facilities, such as GMT/BST or GMT operation and variable hours offset, can be selected with internal switches. Accuracy of the clock is within 5ms of atomic time and the receiving range is claimed to be greater than 1500km. Applications include energy management, computer real-time clocks, master/slave-clock systems and synchronization of separate equipment or events.
Circuit Services, 6 Elmbridge Drive, Ruislip, Middx.
WW310

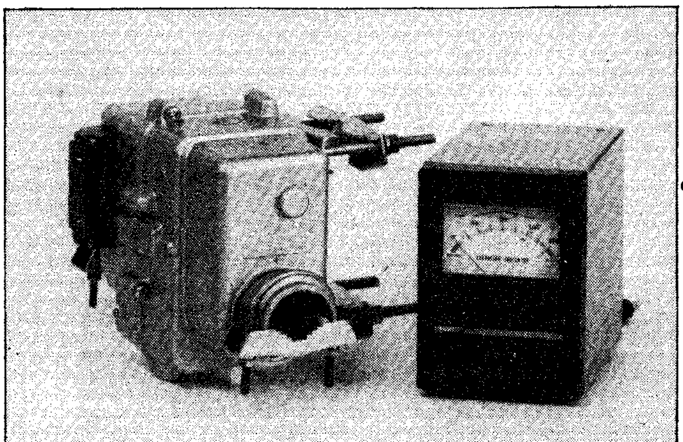
WW310

Microcomputer keyboard

Apart from one or two teething problems, the ZX81 is quite a feat of electronic design when you consider its price. But as the most tedious task in computing is considered to be typing in a program, many owners may find the 81's keyboard a bit of a bind. So Computer Keyboards are supplying what they call a 'professional' keyboard which plugs directly into the ZX81 for around £28.95 including v.a.t. and postage. Each keyboard has a transparent cover under which the key legend is placed. As from 1 January, 1982, a case for the keyboard will be available from the same company.
Computer Keyboards, Glendale Park, Fernbank Road, Ascot, Berks.
WW311

Antenna elevation rotator

According to South Midlands Communications, the KR500 is, 'the only purpose-designed elevation rotator in production'. Of course, until fairly recently, transmitters were firmly fixed to the ground and there wasn't much demand for such a rotator – but times are changing. This antenna rotator is part of a range of similar but vertical axis drives recently introduced by SMC and manufactured by Kenpro. Also available are rotators adapted to work in conjunction with wind vanes, and with built-in hysteresis to control the direction of scientific instruments.
South Midlands Communications Ltd, SM House, Osborne Road, Totton, Hants SO4 4DN.
WW312



WW312

Tv deflection i.c.

A single integrated circuit containing sync. circuit, oscillator, ramp generator, flyback generator, protection circuits and power output stage is manufactured by SGS. The TDA1670 provides current outputs of up to 3A peak-to-peak with supplies of up to 36V and a flyback voltage of 60V. Thermal overload cut-out protects the i.c., and the blanking generator can be used to cut off the beam current to avoid tube damage if vertical deflection collapses. A 15-lead version SGS Multiwatt package houses the i.c.
SGS-ATES (UK) Ltd, Walton Street, Aylesbury, Bucks.
WW313

Waves

By Ariel

Far from free speech

The annual problem of what to buy the loved one for her next birthday has been partially eased by the news of an exciting development from the firm of Nippon Electric. It's a machine, selling at a paltry $15,000, which is claimed to be capable of 'hearing' words and then printing them out via a word-processor.

But don't think your monetary outlay is going to stop there. The fact that this contrivance recognises no other language – not even English, by gad sir – means that you'll have to treat the loved one to a crash course in Japanese as well. I'm assuming, by the way, that you have a word-processor kicking about the house somehere. So you'll be saved that expense.

Certain other small items may also crop up. For instance, when using the machine the LO will need to speak at slightly less than her normal conversational speed and to pause after the delivery of each syllable. Now, I can't speak for your LO, but the one I've got at home is internationally notorious for the rate at which she punches out the words and phrases. In fact, in the interests of science, we once took her to one of those firing ranges in the back of beyond and measured her performance against that of an old Gatling machine loaded with a belt of 100 rounds. We wrote the LO a speech of precisely 100 words, blew a whistle, and they were off, the pair of them.

You'll never believe this, but the LO was knocking back a restorative gin before the Gatling had stopped chattering. So if she-to-whom-you-would-give-your-all comes into this supercharged speech bracket you may have to foot the bill for expensive depressant drugs, to be taken prior to a session on Nippon's wonder box. If that doesn't work she could try chewing on a hard apple or a chunk of stickjaw while mouthing her piece. If it does nothing else it should make her Japanese sound interesting.

Ghastly to have met you, Mum

In the September issue of *Wireless World* I tactfully suggested that British Telecom (which Parliament has now made respectable) would be performing nothing less than a public service if they applied the excellent (in this case) principles of euthanasia to that illiterate disgrace to the ornothological world, Buzby. He has, I submitted, polluted our screens for too long.

I never seriously hoped that my subtle hint would be acted upon, but I wasn't prepared for BT's reply: the sudden appearance one night of Buzby's mother. The fact that he had one was the biggest surprise. I always imagined him to be a clinical laboratory lash-up that went wrong.

Now, having seen these two monsters together, I can only say they thoroughly deserve each other. But we still do not.

Influence of angels

Have you noticed the growing element of commercial sponsorship that's creeping into our TV fare nowadays?

It shows up mostly in the sports sector: show jumping, athletics, table tennis, badminton . . . Even cricket, which was the last bastion of all that's clean-limbed and county, has fallen under its spell. And as for motor racing, it's a miracle to me that the cars ever make it round the circuit under the weight of all those posters and emblems.

Not only sport is affected, however. Those dreadful contests are too. You know the sort of disaster I'm talking about: "The Miss Beautiful Bicuspids of Great Britain" competition, presented in association with the Dire Dental Floss Corporation. "The Year's Hairiest Chest", a tasteless exercise aided and abetted by the maker of some tonsorial stimulant or other.

Those responsible must be living in a world of their own if they really believe these trivial affairs, designed mainly to promote private commercial interests via the back door, add up to entertainment for the majority of viewers.

Another thing that puzzles me is that very often these angels of the box seem to have nothing in common with the programme they're backing. What, for example, is a cigarette manufacturer doing putting up a trophy for cricket? Surely this is one area where scar-free lungs and plenty of puff are basic essentials. And it is appropriate that a well-known firm of tv setmakers – whose product is designed to be used in a sitting position – should award its patronage to an athletics meeting?

In the show jumping arena you'll come across even more marked incongruities. There's the Titmarsh Teamaker Trophy, the Cornish Cultural Circle Cup, the Redditch Roadmending Company's Rosebowl and the Sussex Sheet Metal Workers' Shield. Just what, tell me, have cups of tea in bed, mind-improving activities, mending holes in the road and metal work got to do with sweating nags and jump-orfs?

For those who have eyes to see, the writing is on the wall. Before we know where we are we'll have jewellery stores offering cut-price wedding rings to bridegrooms willing to fly advertising pennants from their toppers as they step up the aisle. We may yet see manufacturers of fire extinguishers promoting spectacular blazes. Who knows?

Window-knocking

If you want to relieve the tedium of waiting for your loved one (here she is again) while she pops into the local department store for a couple of hours to try on some dresses, try strolling down the High Street and treating yourself to a session of window-knocking. This is akin to window-shopping, but a good deal more purposeful. It consists of thoroughly inspecting window displays and then deciding which one constitutes the biggest shambles.

You have to make allowances when arriving at your judgment, of course. Obviously, a milliner's window containing nothing but one hat on a pedestal and a card bearing the words "Hier Man Spricht Deutsch" has an inborn advantage over the one displaying assorted ironmongery.

Over the years (because my LO likes trying on dresses in department stores) I've had the opportunity of regularly making comparative studies of window-dressing techniques and have become a knocker whose views are not lightly dismissed. And while there are notable exceptions, I hasted to add, I'd say that the area of retailing most likely to benefit from a fundamental window rethink is the radio, tv and domestic appliance trade.

I'm prepared to admit the dealer's problems are numerous. The goods he sells come in infinite variety and in lots of different shapes – some of them awkward. Front-loading washing machines and pop-up toasters do not, I will also concede, lend themselves to imaginative presentation. Any more than you can avoid the stark truth that a TV set is merely a box with knobs on.

Allowing for all these handicaps, there is still bags of room for improvement and I'm sure the trade would do well to look around and take a leaf or two out of the books of other retailers. For instance, the angularity of video cassette recorders could be softened by small tastefully-deployed pot plants. There is no valid reason why the larger portable radio sets should be stood in rigid line like a unit of the Guards. Let them be arranged casually, possibly on a bed of polystyrene chips. Curtaining is a much-neglected material in this area of display. Heavy, richly-coloured drapes will not only provide a dramatic and compelling effect, but will also serve to mask scratches on cabinets, inclicted by Saturday-only assistants.

These are only a handful of suggestions. The enterprising dealer is bound to come up with lots more if he sets his mind to it. But don't overdo it, lads. Otherwise I shall have nothing to knock while I'm waiting for the LO to reappear with a full shopping bag and an empty purse.

Some of the TELOMAN PRODUCTS RANGE

NEW PRODUCTS

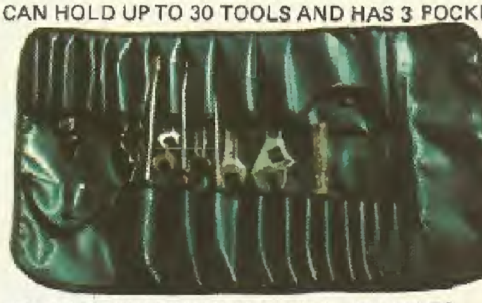
£43.70 inc VAT

SIZE TL100 19"x 14"x 6"

TL99 17"x 12" x 6" £39.90

TOOL PALLET FOR SPANNERS (PRICES ON APPLICATION)

TOOLROLL £8.95 inc. VAT P&P £1.00
MEASURES 23"x13" WHEN OPEN. MADE FROM PVC. IT CAN HOLD UP TO 30 TOOLS AND HAS 3 POCKETS.



SALES PRESENTERS £7.48 inc. VAT P&P £1.50
IT CONTAINS 3 DOCUMENT POCKETS 4 RING BINDER BOARD CLIP WITH QUICK RELEASE. **SIZE A4**

THE **TL100** HAS BEEN DESIGNED FOR THE PROFESSIONAL ELECTRONICS, TV OR INSTRUMENT TECHNICIAN WHO NEEDS TO CARRY A LARGE NUMBER OF SPECIALIST TOOLS. CONSTRUCTED FROM HARD WEARING ABS WITH STRONG ALUMINIUM FRAMES, TWIN HANDLES AND TOGGLE LOCKS. A MOULDED TRAY IN THE BASE, A COMPREHENSIVE 2 SIDED TOOL PALLET THAT IS REVERSIBLE WITH SPACE FOR OVER 40 TOOLS. THERE IS SPACE FOR DOCUMENTS AND A HEAT SINK FOR A HOT SOLDERING IRON TO PREVENT ANY DAMAGE BEING CAUSED.
ALSO AVAILABLE IS THE **TL99** WHICH IS A SMALLER VERSION OF THE **TL100**.

TLW4 TOOLWALLET MEASURES 11"x14"x2½" WHEN CLOSED. MADE FROM REINFORCED PVC WITH A HEAVY DUTY INDUSTRIAL ZIP.

£13.75 inc VAT P&P £1.50.

- ***DISCOUNT STRUCTURE FOR MULTIPLE USERS ONLY**
- ***CUSTOM MADE TOOL PALLETS (ONLY FOR LONG RUNS) TOOLS WILL BE REQUIRED FOR MEASURING BUT WILL BE RETURNED.**

Mail Order

Please send ____________________

Enclosed my cheque £ ________

TL 100/TL99 P&P £2.60 extra)

Name ____________________

Company ____________________

Address ____________________

____

Tools NOT included. British made.
Money back guarantee. Allow 7-21 days for delivery.

TELOMAN GUARANTEE 2 yr

TLW4

Teloman Products Ltd
'Wychwood' 2 Abbots Ripton Rd, Sapley, Cambs. PE17 2LA
Tel: (0480) 65534