Wireless World

Editorial Conventions

We recently received a letter from an American reader asking why some authors writing in the journal use pseudonyms. In general the answer is that those who use pen names are not permitted by their employers to write under their own name as the company does not wish its name to be associated with the particular subject, or maybe project, under discussion. On the other hand there are those who could not write as freely under their own name as under a pseudonym; outstanding instances of this are our late columnist "Free Grid", and currently "Vector". In our reply to our American enquirer we also pointed out some other journalistic conventions such as that unsigned articles are staff-produced and in general express the journal's opinion. Reports on exhibitions, conferences, etc., are typical staff contributions and as such are accepted as the "party line". If, as sometimes happens, a member of the staff, or an appointed correspondent, attends a demonstration, conference, lecture or what-have-you, and in his report adopts a particularly critical or laudatory line, which it would not be politic for the journal to adopt, the writer's initials are then appended. We recall the occasion some years ago when a member of the staff attended the demonstration in London of several different colour television systems preparatory to a decision being taken on a system for this country. Our reporter came down heavily in favour of PAL, which was in opposition to the opinion held in most circles. His initials were added to the report as this was, in other words, the considered opinion of the contributor and, as with published "letters to the editor", was not necessarily endorsed by the editor.

In this issue we depart from a convention which has been established for many years in the journal. Although the fact may be unknown to readers, it has been our unswerving practice to ascribe to a constructional project the title "*Wireless World* this or that" only when the design, development and prototype construction has been undertaken in our own laboratory-workshop or in the home of a member of the staff of *W.W.* under our supervision. The project which is introduced in this issue under the title *Wireless World* Desk Calculator was not conceived or developed by us. It is, however, true that when the idea of such a kit construction was put to us by Advance Electronics we saw it as the fulfilment of an idea which we had been investigating for very many months. Here was a proven design which, by collaboration between the designer's company and ourselves, would make available to readers a calculator fairly simple to construct (it takes about five hours) and yet professional in appearance. In this issue is the first of the two articles which discusses the design philosophy of electronic desk calculators, and next month full constructional details and operational procedures will be given together with particulars for obtaining the complete kit of parts. The articles are the joint work of Roger Alexander, the designer, and Brian Crank, who was deputy editor of *Wireless World* until early July, when he left to go into public relations.

The Wireless World Desk Calculator

1. Design philosophy

by Roger Alexander and Brian Crank†*

The *Wireless World* Desk Calculator is the result of close co-operation between *Wireless World* and Advance Electronics Ltd who designed the machine. The article describing this machine is divided into two parts. This month the evolution of low-cost calculators is discussed and then the circuitry and operating principles of the calculator are described. In part two of the article, next month, full constructional details will be given. These instructions will apply to a kit of parts which will be available for £39.25 plus 75p postage and packing charge. The kit is complete in every detail and includes the printed circuit board and a moulded plastics case. Details of how the kit may be ordered will be given next month.

For a long time engineers have realized that if a small automatic calculator could be produced at a fairly low price the market would be enormous. The first automatic calculators were mechanical and had no chance of achieving this ideal, for they were large and cumbersome and needed regular maintenance. Even more modern mechanical calculators could not make much impact on the potential market as they were large, noisy and expensive.

In the early 1960s, with the advances being made in computer technology, electronics engineers realized that the ideal calculator was much more likely to be realized using electronic techniques. Early design work was aimed at producing an electronic analogy of the mechanical machine of the time. The reason was that the people responsible for planning argued that users would be suspicious of anything that was radically different from the norm.

The machines were constructed from the discrete components available at the time and, in general, were large, difficult to use and expensive. In fact they did not offer many advantages over the mechanical machines they were intended to replace.

When t.t.l. integrated circuits became available at low cost these were employed in calculators to great advantage; a turning point had been reached. The performance of the t.t.l. machines was better and the price came down. The main trouble was a fairly large number of t.t.l. integrated circuits was required and the amount of labour involved in constructing the machines was quite high. The result was that the price could not be reduced to the level needed to reach the mass market.

The semiconductor industry then began producing medium scale integrated circuits and it was not long before i.c. designers turned their efforts to calculators. The result was a range of calculators based on only four integrated circuits. These calculators appeared in 1970/71; they were small, compact and practical but they were too expensive for the domestic market. A stage had been reached where just about all the electronic design problems had been solved and the electronic circuitry was potentially capable of a higher performance than was really required. Who could tell the difference between a calculation time of 100 and 200ms?

Electronic calculator design was also restricted by the need for a cheap method of displaying the numerical result of a calculation. Most of the calculators used gas discharge display tubes which required complex drive circuitry, inconvenient high-voltage supplies and, of course, they did not "interface" easily with integrated circuits.

The introduction of the gallium arsenide seven-segment digital display was a major step in the right direction as far as the calculator manufacturers were concerned. The displays were small and only required low voltage, low power drive. However, they were not without their difficulties because it proved too expensive to produce the large sized displays sometimes required by office users, but, nevertheless, they were admirably suitable for the small calculators we are discussing.

Semiconductor manufacturers had by this time started producing large scale integrated circuits based on m.o.s. technology and in 1971 Texas Instruments produced a calculator on a single chip: one integrated circuit which contained all the arithmetic, control and storage functions necessary to

*Advance Electronics Ltd.

†Formerly deputy editor, *Wireless World*, now with T. J. Burton Associates.

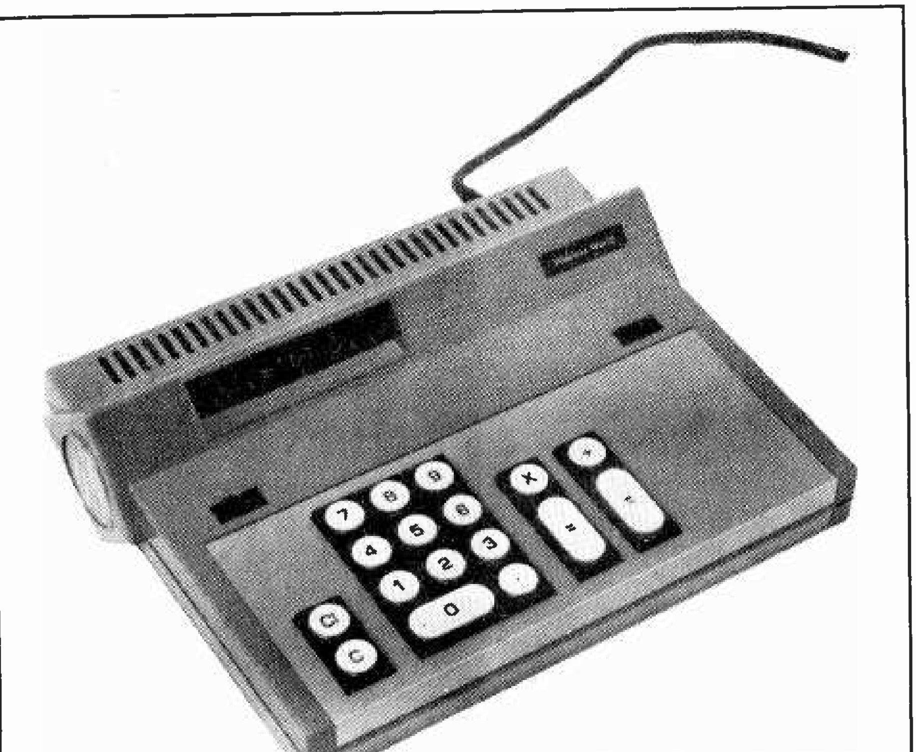


Fig. 1. The completed machine.

build a calculator. This represents the current state of the art, but is still not the ideal. For instance, the calculator chip will not drive the display directly, so suitable interface circuitry must be included and two power supply voltages are required.

What will happen next is a matter for conjecture but if the predictions made last month in the editorial (Electronic Calculators, p. 357) materialize the future will see a calculator based on a liquid crystal display driven directly from a c.m.o.s. integrated circuit. The whole thing would be mounted on a single substrate and would be powered by a single battery.

Whichever way calculator evolution proceeds, it is fair to assume that within a few years the price will have fallen to a level within reach of practically everybody. Calculators will probably be available with sufficient versatility to replace the engineer's slide rule at the price, say, of a small transistor radio receiver.

Once calculators are available in very large numbers, at low cost, it would be fair to ask what effect they might have on society. Will it be on a par with the computer revolution? Will it be necessary to teach arithmetic any more? The prospect certainly provides food for thought.

Wireless World desk calculator description

As mentioned earlier the *Wireless World* calculator (Fig. 1) is based on the calculator integrated circuit manufactured by Texas Instruments and employs nine gallium arsenide seven-segment displays. In addition to being able to add, subtract, multiply and divide the calculator has several other features. It will perform chain operations, which means that a number of operations can be carried out consecutively without the need to reset the machine; for instance $Y(X^2+Z)$. The machine also allows any number of calculations to be made with a constant factor which is required to be entered only once. If one needed to convert 12 and 27.5 inches to millimetres they have to be multiplied by 25.4 (the constant). The calculator is switched to the 'constant' mode of operation and 25.4 is entered on the keyboard, followed by the instruction to multiply (the calculator now stores $\times 25.4$). The first figure 12 is entered and the 'equals' key is pressed; the result of 12×25.4 is displayed. Next 27.5 = is keyed and the calculator will display the result of 27.5×25.4. There was no need to enter the 25.4 again for the second calculation or 'clear' the machine.

In a similar way division by a constant number can be performed and the machine can be made to store $\div X$, where X is a number within the range of the machine. It must be noted that the "constant" and "chain" modes of operation cannot be carried out at the same time.

The calculator will square a number automatically, will indicate an error or an overflow and it suppresses insignificant leading zeros. The brightness of the display is variable in steps and the result of a calculation can be as large as 16 digits to the left of the decimal point. With numbers as large as this the eight most significant are displayed.

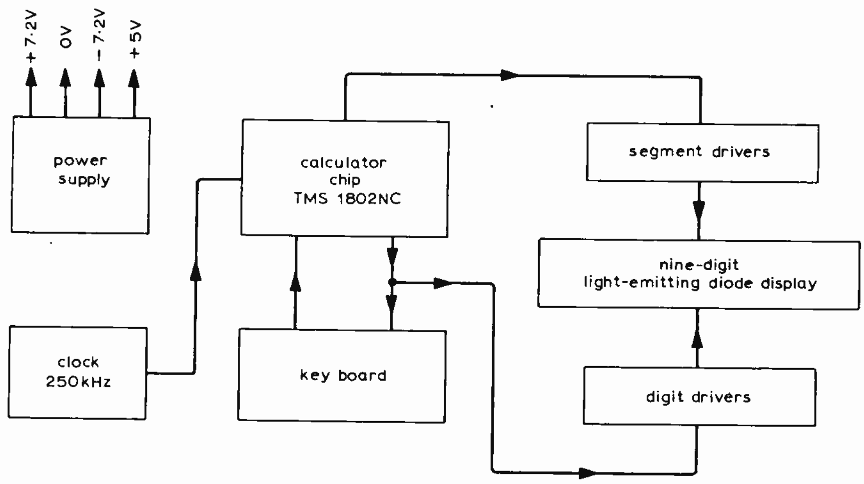


Fig. 2. Block diagram of the calculator showing the main sections.

The calculator operates in the floating point mode, that is, the decimal point can appear anywhere in the display, the least significant digit appearing at the extreme right of the display. The rest of the specification is detailed in Table 1 and a block diagram appears in Fig. 2, which should be referred to during the following description.

Table 1

Desk area	200 × 240mm
Power supply	110 or 240V a.c. at 9VA
Weight	1.1kg
Temperature range	0 to 50°C
Calculation time	<200ms

Calculator integrated circuit

The Texas Instrument's TMS1802NC integrated circuit is the heart of the calculator. It is a single silicon chip, only 6.56mm square, containing more than 6,000 transistors, and it is one of the most complex m.o.s. large-scale integrated circuits ever produced using the silicon nitride process.

A block diagram of the chip is shown in Fig. 3. The timing block is formed from a programmable logic array (p.l.a.). This means that an array of transistors is laid down selectively on the chip. The function this array performs is decided by the pattern of transistors laid down at a stage during the production. In other words the function that the array performs is defined by the thin oxide mask used by the second layer, in this sense, the array is programmable. The timing block contains such things as the master counter and circuitry for producing timing pulses for the keyboard and for time multiplexing the display (more about that later).

The programme read-only memory (r.o.m.) stores details of all the operation algorithms and issues them to the control p.l.a. where they are decoded and used to control the timing decoders and the arithmetic logic unit. Feedback from the control p.l.a. to the programme r.o.m. allows the programme to loop and branch within the algorithms. The data random access memory (r.a.m.) and arithmetic logic unit stores the data (numbers entered by the keyboard and results of calculations) and has all the necessary logic to perform arithmetic functions as defined by the programme r.o.m. It is interesting to note that the arithmetic unit consists of a set of logic gates and not a sequential adder, b.c.d. shift registers, etc., made up from flip flops. The output p.l.a. suppresses the leading zeros in the display and changes data which is to be displayed into a form suitable for driving the seven-segment indicators.

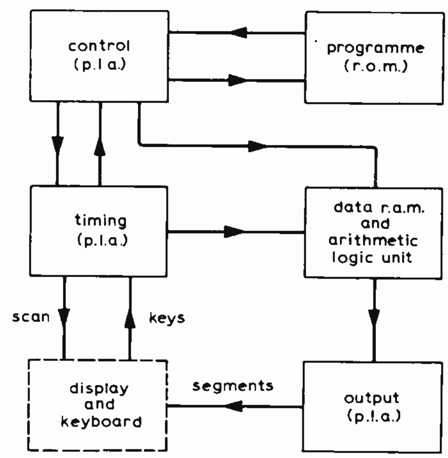


Fig. 3. Block diagram of the calculator integrated circuit.

Keyboard

The keyboard consists of 17 keyswitches and two slide switches. The key switches contain sealed reed relays and were designed and built by Alma Components Ltd to a specification produced by Advance Electronics. At the time the machine was being designed all reed key switches were very long because the reeds were mounted vertically and were actuated by circular magnets contained in the button. A "slim line" machine was needed, so radical changes had to be made to the existing key switch designs. The problem was resolved by mounting the reed diagonally, in a horizontal plane, in a square box which forms the base of the key switch. The final size of the base was fixed at $19mm^2$ and the height of the switch at 15.9mm, enabling the machine to be reduced to about 30mm in depth.

The switch plunger houses a small magnet which actuates the reed switch, and herein

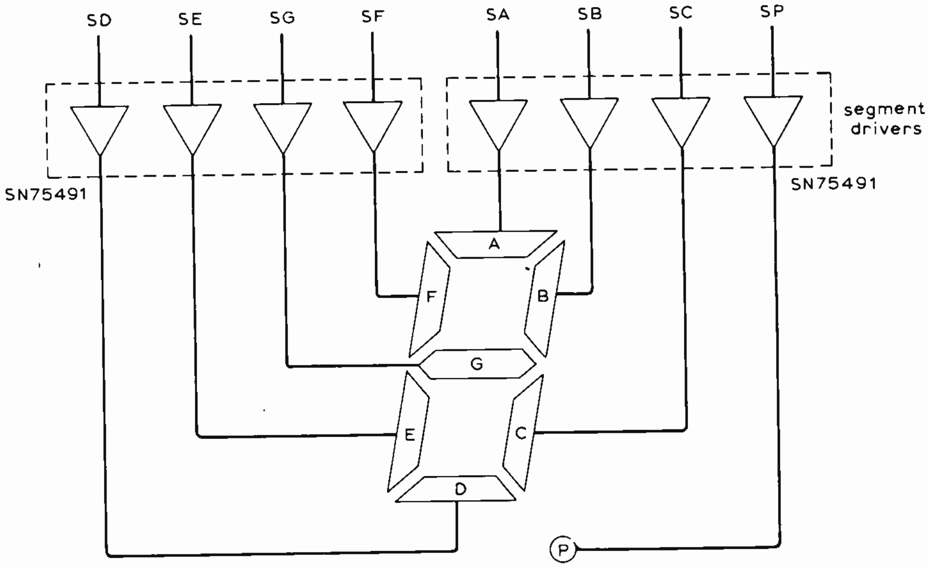


Fig. 5. Layout of the light-emitting diodes in one indicator. The segment lettering A, B, etc., corresponds to the calculator i.c. outputs SA, SB, etc.

lay a problem. The low overall height limited the plunger traverse to about 3.2mm and there was a danger that the reed would not open when the plunger was at the top of its travel because of the close proximity of the magnet. This problem was solved by incorporating a metal keep at the top of the plunger travel which effectively removed the magnetic field from the reed. It also gives a bonus which Advance Electronics call negative feel. A small force is required to start the plunger moving because the magnet is in contact with its keep. This resistance disappears as soon as the plunger moves, giving a very pleasant feel to the keyboard.

The plunger has a tapered top, onto which are pushed double-shot moulded key tops in black and white. The base of the key switch contains the two contacts for printed circuit board mounting and two orientation pips. This allows the switch to be mounted in one plane only so that no interaction occurs between adjacent switches.

Two slide switches are mounted directly on the printed circuit board; they are the constant-mode switch and the brightness level adjustment. The constant-mode switch is single-pole, two-position and the brightness switch single-pole, three-position. These switches are operated through the case top via moulded switch tops, engraved with the operating function.

The key switches are connected in a matrix as shown in Fig. 4; one has to imagine a contact operated by each button which connects a horizontal wire to a vertical wire. Fig. 4 shows all the keys which can be connected to the calculator chip; not all these keys are used in the *Wireless World* calculator.

The calculator i.c. produces 11 timing signals which appear at the 11 pins one after the other. These pins on the i.c. are connected to points D1 to D11 on the keyboard diagram (Fig. 4). The calculator i.c. is also connected to the points KN, KO, KP and KQ. Let us assume that the ÷ button is pressed. During timing signals on D1 and D2 there will be no output from the various K outputs to the i.c. but during the time a signal is present on D3 there will be an output on KO. When there is an output at KO during this time the i.c. interprets it as an instruction to divide. In a like manner, when any button is pressed the i.c. can determine which one it is.

The calculator i.c. has an input sensing programme which provides protection against transient noise, double entry (two buttons pressed at the same time) and switch bounce.

While the calculator is switched on but is not being used, the calculator i.c. goes into an "idle" routine. It sequentially scans the KO and KN lines, "looking" for a non-quiescent condition. When such a condition is detected the i.c. waits 6.8ms and goes into a routine called TPOS which again scans the input to determine if the signal was the result of a keystroke or was just transient noise. If the latter was the case the i.c. reverts to the idle condition.

If the TPOS routine proves a keyboard entry the i.c. responds accordingly and performs the task demanded of it. The i.c. then enters a programme called TNEG, which scans the keyboard looking for a non-quiescent condition; if this test is successful the i.c. returns to the idle condition.

If the TPOS routine discovers a KN input the programme jumps to the 'NBR' (number) routine and the data are entered into the random access memory. If a KO input had been detected the programme would have jumped to 'OPN' (operation) and the calculator would have executed the instruction. Keys must be pressed or broken for a minimum of 25ms.

Display decoding and driving

The result of calculations is presented on gallium arsenide seven-segment indicators of the form shown in Fig. 5. Each segment (numbered A to G) and the decimal point (P) is formed by a light-emitting diode. All eight diodes are mounted on a substrate and are connected as shown in Fig. 6. In the calculator the display packages measure about 6.9mm high by about 5.6mm wide with a digit height of 3mm. The displays are magnified to provide a final digit height of approximately 5mm.

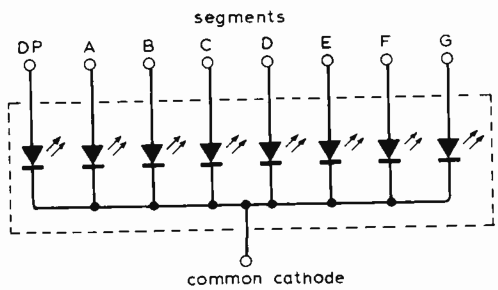


Fig. 6. The circuit of one of the gallium arsenide indicators showing how the eight light-emitting diodes are connected.

The calculator chip has eight outputs, each output intended to illuminate one of the light-emitting diodes so as to form a complete figure as shown in Fig. 7. The letters SA to SG and SP represent the outputs of the calculator i.c. and SA is the output which drives segment A (as defined in Fig. 5) in an indicator and SB drives segment B, and so on. Fig. 7 serves two functions. It shows which segments have to be illuminated to form the various numerals and symbols, and it also shows which outputs from the calculator i.c. are needed (indicated by the thin bars) to form those numerals and symbols.

There are 9 indicators (8 for numerals plus an indicator for symbols), each containing 8 light-emitting diodes, and yet there are only eight outputs from the calculator chip. One would have thought that, as each indicator requires 8 inputs, one would need 8 × 9 outputs from the calculator chip. However, if one thinks about it, 72 output

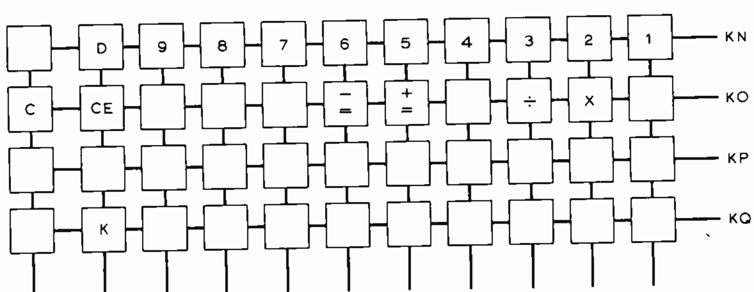


Fig. 4. The keyboard matrix. Timing signals from the calculator i.c. are fed to the matrix at D1 to D11. The calculator i.c. scans the K outputs and can determine which key has been pressed because each key will provide an output for a unique combination of timing signal and K output.

leads from one i.c. would not be practical.

What happens in the calculator is this. Each indicator is selected in turn and the calculator provides the correct combination of output signals necessary to drive that particular indicator. Therefore only one indicator is illuminated at a time. However, the circuitry switches from one indicator to the next so quickly that to the eye all the indicators appear to be illuminated continuously.

It was mentioned earlier that the calculator i.c. could not supply enough power to drive the indicators directly. Integrated driver amplifiers are therefore used to interface the calculator i.c. with the indicators. These same drivers also allow the indicators to be switched on and off sequentially, or to be time multiplexed as the technique is known. We still need a means of knowing when to switch a particular indicator on and still need some sort of control signal to do it. In the calculator the timing signals on D1 to D11, which were used to scan the keyboard, are employed to select the correct indicator as well, as shown in Fig. 8. Notice that D9 and D10 are not used.

Two kinds of indicator driver are used. One type, known as digit drivers, are used to select the individual indicators and are therefore driven by the timing signals on D1 to D11. The second type, segment drivers, are connected between the SA to SG and SP outputs of the calculator and the individual indicator segments.

All the drivers are housed in four dual-in-line bipolar integrated circuits. The segment drivers employ two type SN75491 i.cs, each i.c. containing four Darlington circuits cap-

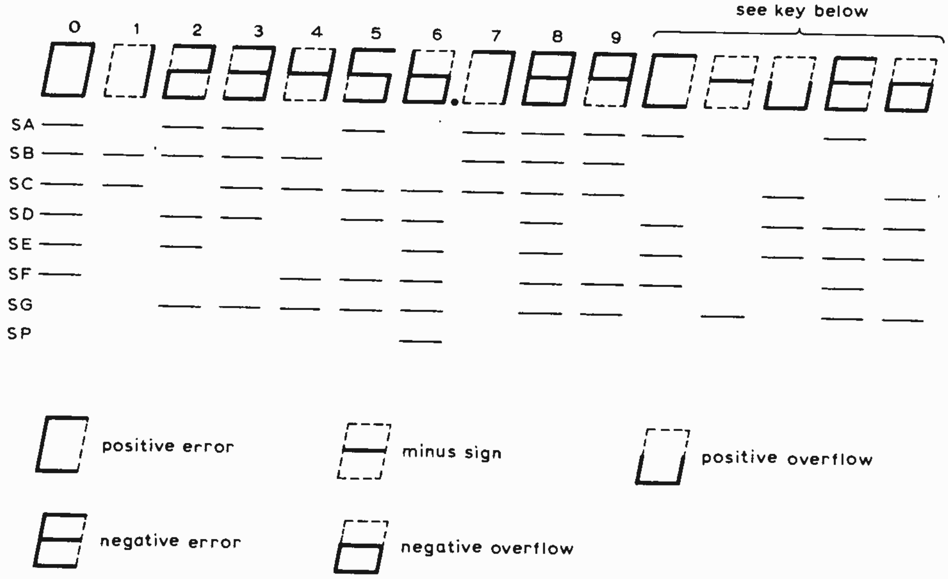


Fig. 7. The display font and how the various numerals and symbols are made up from the light-emitting segments. The bars opposite the letters SA, SB, etc., represent outputs from the calculator chip. The symbols appear only in the ninth, extreme left-hand, indicator of the calculator and inform the user of several fault conditions and also give the sign of the indicated result.

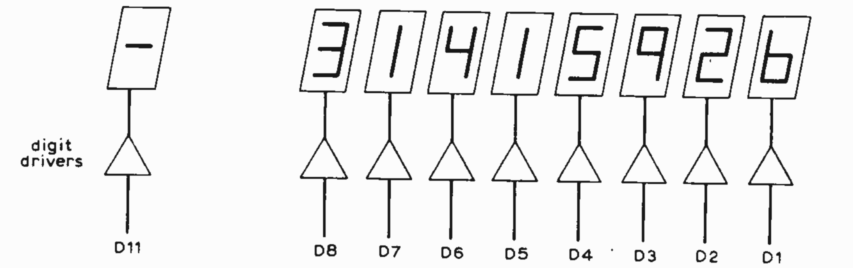


Fig. 8. How the timing signals that were used to scan the keyboard also feed the digit driver amplifiers so that only one digit at a time is illuminated. Timing signals D9 and D10 are not used for display purposes.

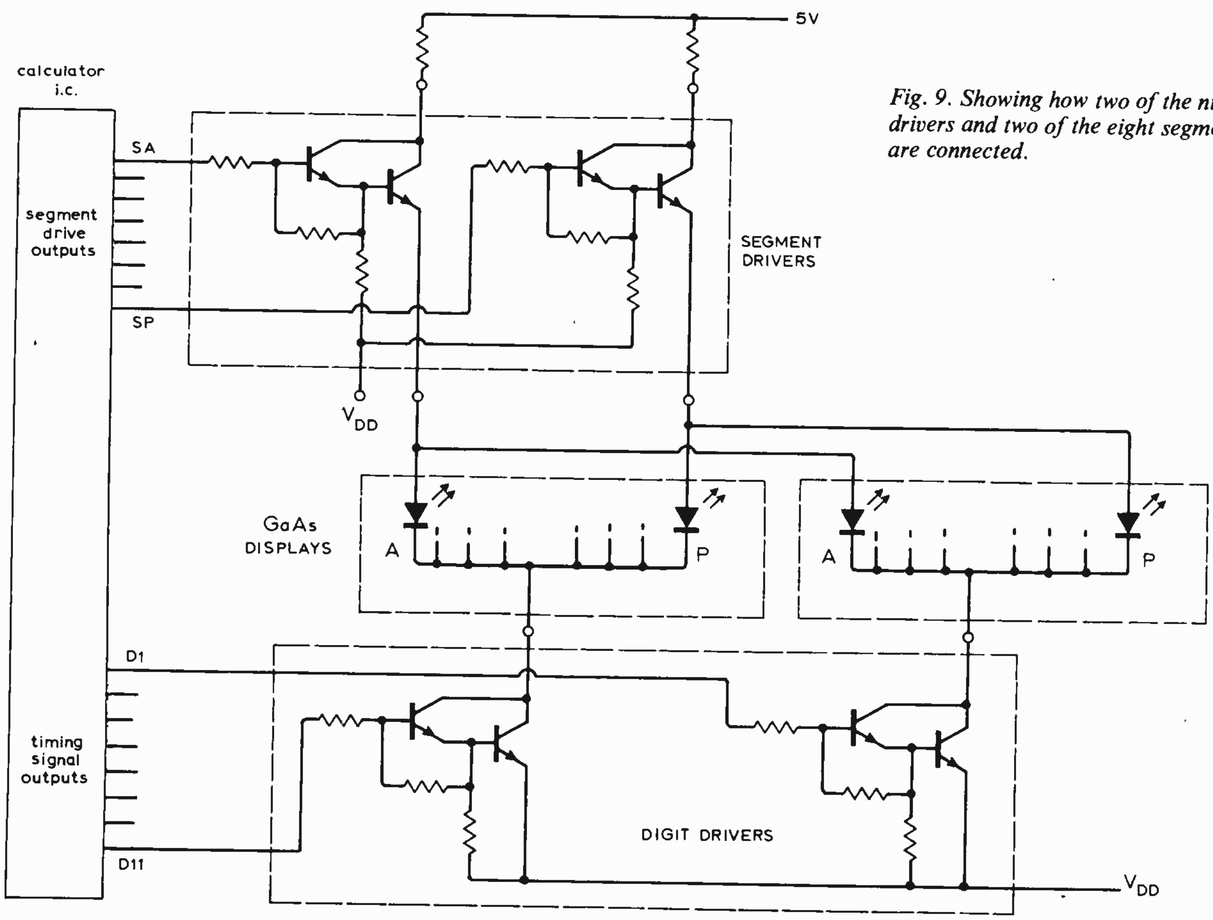


Fig. 9. Showing how two of the nine digit drivers and two of the eight segment drivers are connected.

able of switching 40mA with a maximum 800mV drop across them. These take care of the seven segments and the decimal point. The digit drivers employ two type SN75492 i.cs, each i.c. containing six Darlington circuits capable of switching 320mA with a maximum 900mV drop across them. There are 12 digit drivers in the two packages and, as we require only nine, three are left spare.

Fig. 9 shows the circuit for two of the eight light-emitting diodes in two of the nine indicators. The indicator on the left-hand side of the calculator display is the one which is used for symbols as it is driven by the D11 timing pulse. The right hand digit is the least significant and is driven by the timing pulse on D1. In Fig. 9 the segments are A; the decimal point shown being driven by SA and SP.

Clock generator

The clock generator needed to drive the calculator i.c. is a hybrid thick film integrated circuit, which makes for easy construction and reliability; it measures only 25 × 12 × 3.5mm. The internal circuit of the i.c. is given in Fig. 10; operating frequency is 250kHz.

Power supplies

The circuit of the power supply is given in Fig. 11. Two bridge rectifiers connected to two separate secondary windings on the mains transformer provide the inputs to two voltage stabilizer integrated circuits. The output of the 15V regulator has a zener diode and a 220Ω resistor in series across its output to provide outputs of plus and minus 7.5V relative to the zero line for the calculator i.c. The second stabilizer provides a +5V output. Both the stabilizers exhibit excellent transient response.

If one of the transformer secondaries is short circuited the fuse in the primary circuit will blow. This is to ensure that the calculator will meet the British Standard Specification for office equipment which requires that the transformer should not reach more than 100°C when the secondary is short-circuited. The two 1.8nF capacitors and the two coils L_1 in the transformer primary circuit protect the calculator from pulse interference in the mains. The coils consist of four turns each on the same toroid core.

The two transformer primaries are connected in series for 240V operation and in parallel for 110V operation. The fuse size is 50mA for 240V and 100mA for 110V supplies.

The power supply has a much higher capacity than needed for this particular calculator because it was designed as a standard sub-assembly for a range of machines.

Because of the extensive use of integrated circuits construction of the calculator is a simple matter.

Full constructional details will be given in the concluding part of this article next month which will also describe operating procedures with worked examples.

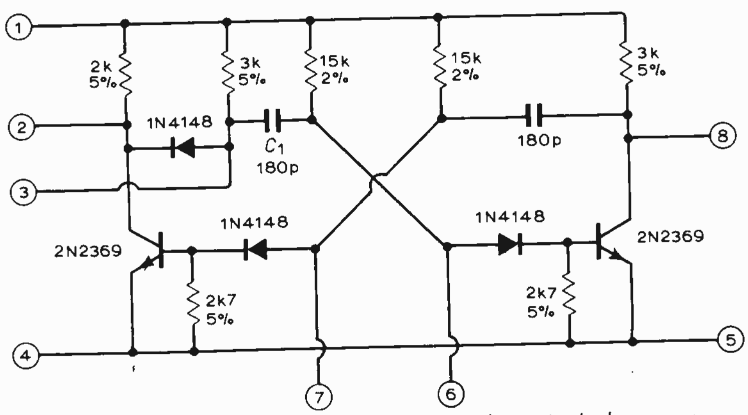


Fig. 10. Circuit diagram of the hybrid thick film integrated circuit clock generator.

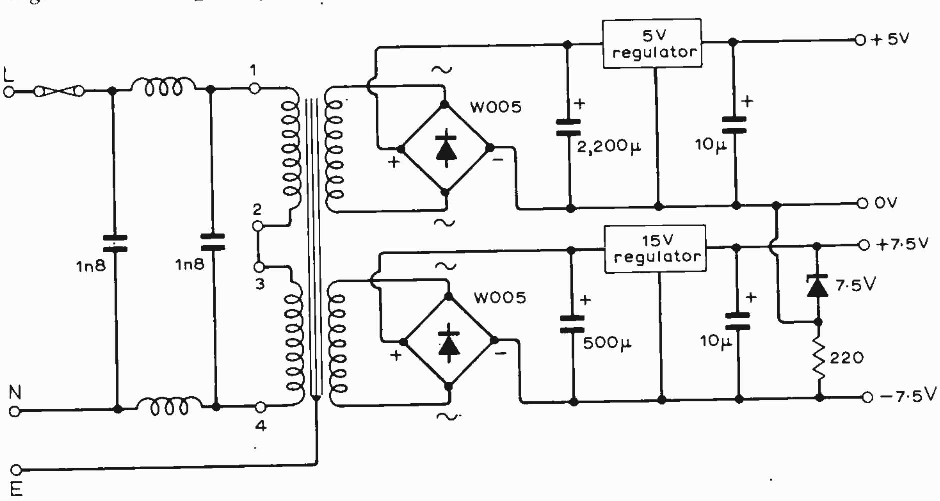
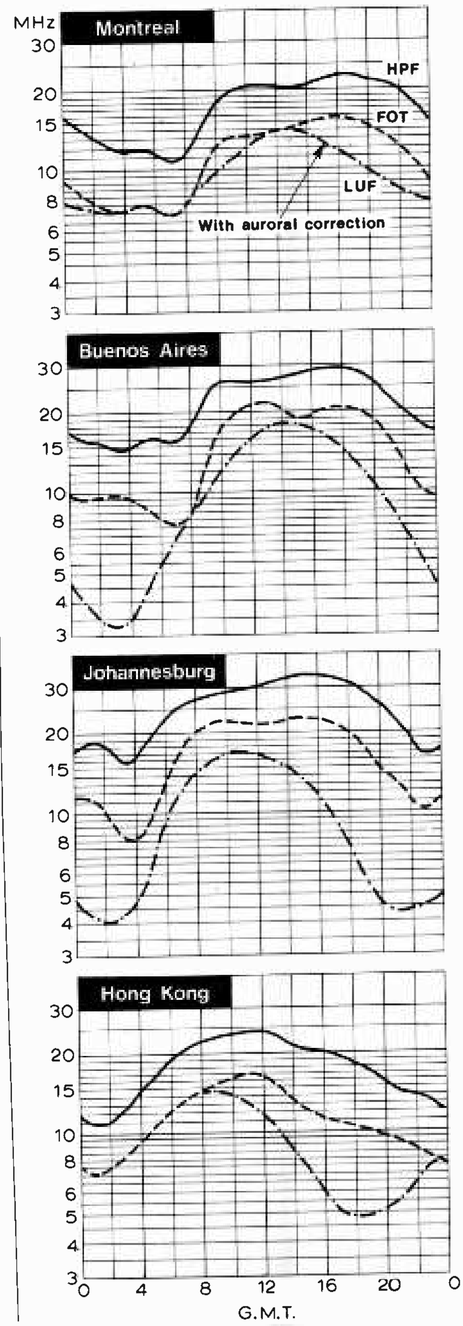


Fig. 11. Power supply circuit diagram.

H.F. Predictions — September

There is little change in the charts from those for August. Optimum daylight frequencies are slightly higher, optimum night-time frequencies a little lower and lowest usable frequencies (LUFs) remain almost the same. There should be an improvement in working across the North Atlantic as ionospheric conditions are near identical at both ends of the route throughout the 24 hours. The LUF shown for this path allows for periods of high attenuation which can persist for several days, normally the LUF is around 3MHz lower at mid-day. HPF is the highest probable frequency and FOT the optimum traffic frequency.

The charts are usable for large areas around London and the destinations as marked. Certainly to plus and minus ten degrees with useful indications to plus and minus twenty degrees of latitude and longitude.



News of the Month

B.B.C. sound programmes go on p.c.m.

Consistent quality of transmitted sound at all places in the U.K., regardless of the distance of transmitters from studios, will be provided by a pulse-code modulation distribution system which the B.B.C. is now installing for its sound radio programmes. An experimental service, carrying audio signals for Radios 1, 2, 3, 4 and Radio London, will be operating between Broadcasting House and the Wrotham, Kent, v.h.f. transmitters by mid-October. It will, of course, include two channels for Radio 3's stereo broadcasts, and, in addition, four "contribution circuits" by which programme material originated in London can be sent, via Wrotham, to regional studio centres. This first stage of the p.c.m. distribution scheme will be followed by an extension to the Midlands by the end of the year and one to the North by early 1973.

The advantage of using p.c.m. for this purpose is, of course, that the original waveform is preserved, regardless of sending distance, provided the streams of digits representing amplitude samples of the waveform are received free of errors. The equipment therefore includes a means for protection against errors. A characteristic of p.c.m. is that it requires a large bandwidth for a given audio signal bandwidth, though this is compensated by the ruggedness, stability and resistance to circuit noise of the system. The London-Wrotham scheme, with a capacity of 13 audio channels, uses the equivalent of one television channel — approximately 5.5MHz. However, the B.B.C. already has available a network of wide-band cable and microwave links, so that this bandwidth requirement is no great problem.

In the p.c.m. equipment, developed by the B.B.C's Designs Department, the audio signal is sampled at a frequency of 32kHz and each amplitude sample is encoded into a 14-bit pulse train, 13 bits for information and 1 parity bit for error protection. The encoded signals from 13 sources are then combined in time-division multiplex to give a single "bit stream" with a bit rate of 6.336M bits/second. Each audio channel has its own coder and decoder, so the 13 available channels can be allocated to stereo pairs and/or mono channels as required. Each has a bandwidth of 40Hz to 15kHz, a peak signal/peak weighted noise ratio of 69dB and provides at the receiving end a standard 1mW signal into 600Ω.

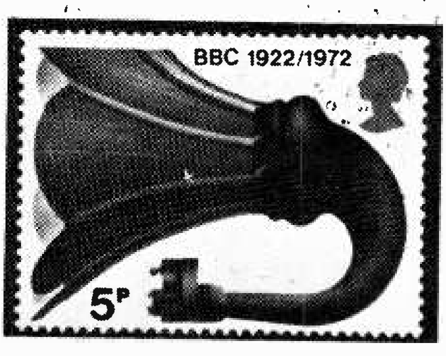
This distribution system can be seen as one further stage in the progressive "digitization" of all sound and vision signal distribution for broadcasting — a process which started with the "sound-in-syncs" system for television developed by the B.B.C. about four years ago (see *W.W.* January 1969, p.38).

Changes to B.B.C. medium-wave services

To prepare for the introduction of commercial local radio, to enable the B.B.C. local stations to broadcast on medium waves and to release an additional wavelength for the B.B.C's external services, it is necessary to make certain changes in the wavelengths used by some of the B.B.C. transmitters on medium waves. Most of the changes will be made from the start of broadcasting on 2nd September. The changes involve the medium-wave services of Radio 3 and Radio 4 only and listeners who use v.h.f. will not be affected.

Instead of the present seven wavelengths used for Radio 4 in England, only four wavelengths will be used and all medium-wave Radio 4 transmitters in England will broadcast the same programme. The wavelengths to be used are: 434 metres (692kHz), 330 metres (908kHz), 285 metres (1052kHz) and 261 metres (1151kHz). The Radio 4 services in Scotland, Wales, and Northern Ireland, will not be affected by these changes.

There will be only one medium wavelength for Radio 3, namely 464 metres (647kHz). Of the existing transmitters at present radiating Radio 3 on 194 metres (1546kHz), some will close down and others will change to 464 metres. The local radio stations will be broadcasting on medium waves and this means that many listeners who do not have v.h.f. receivers will be able to hear their local stations for the first time.



One of three special stamps, in values of 3p, 5p, and $7\frac{1}{2}$p, which go on sale at Post Offices throughout Britain on 13th September to celebrate the jubilee of the beginning of daily broadcasting in the U.K. by the British Broadcasting Company. A fourth stamp (9p) marks the 75th anniversary of the experiments by Guglielmo Marconi and George Kemp that culminated in 1897 in the first successful radio tests across water — nine miles across the Bristol Channel.

The 1972 Audio Fair

Wireless World is once again actively involved in this year's Audio Fair which will be held at its usual venue, Olympia between Monday October 23rd and Saturday October 28th. As in previous years, the festival organizers are providing a lecture theatre where, during the afternoon and evenings, a variety of demonstrations, lectures, etc will take place. The *Wireless World* programme includes lectures by Dr. Ray Dolby on noise reduction systems, Professor P. B. Fellgett, of the University of Reading, on multi-channel sound reproducing systems, Dr. A. Bailey, of Bradford University, projecting the future of audio, and J. R. Stuart, known to readers for his articles on a tape recorder design and noise reduction circuits, will speak on the trends in audio systems. Full details of the Audio Fair programme of lectures and a summary of exhibitors will appear in a future issue of the journal.

The B.B.C. will be presenting demonstrations of stereo reception for which the p.c.m. distribution system already mentioned will be used.

The Future of the electrical and electronics industry

A report issued by an American study group within the I.E.E.E. reveals some interesting details about the economic state of the electrical, electronic and allied industries in the United States. Whereas the growth rate of the American Gross National Product displayed strong upward tendencies between 1950 and 1970 – despite a nasty null in the period 1955-60 – the growth rate for employment of engineers declined sharply from 8.1% to 2.4% in the same overall period. Using the accumulated data from their own researches and available Government sources, the study group were then able to make a prognosis for the next 10 years.

significant feature of the analysis required for such an exercise was an examination of the U.S. budget expenditure which showed a switch from a 50% national defence allocation in 1960 and only 27% on human resources to 1970 figures of 34% for National defence and 42% on human resources. Despite this massive change of emphasis the study group felt there was room for cautious optimism and that most industries in the area studied, would show a growth rate of up to 8% per annum. The strongest areas for advance are telephone and telegraphs, computers, radio and television.

On the international front it was accepted that Japan and Germany have shown the greatest technological advance in the past decade and that the gap between them and the U.S. is rapidly narrowing. One of the reasons for this is the reduction in government spending on R. & D. and the committee members strongly recommended a change in this policy.

It is widely held that the trends displayed by American industry are mirrored at a later date by similar changes in the economic state in the U.K. For this reason the report may make a valuable contribution towards predicting our future as well as pointing out the marketing attitudes likely to be adopted by our American friends. The report is titled "Economic Conditions in the U.S. Electrical, Electronics and related Industries — An Assessment" and is available from I.E.E.E. Inc., New York, N.Y.10017, U.S.A.

Advanced design cassette machines

The Japanese company, Nakamichi Research Inc., have recently announced in Japan and America, the development of two new cassette recorders claimed to be superior to their previous models. Several unique features appear to be incorporated in the designs and surprisingly one of the two models is specifically designed for the professional market. Model ZETA as the professional version is named, is a dual capstan deck using a magnetic, phonic wheel, servo system, with the phonic wheel forming part of the motor shaft. By using large flywheels and this form of servo control, together with a second motor for tape spooling, wow and flutter figures of less than 0.1% are claimed.

Another unusual feature is the use of a three-head line-up, enabling optimal head gaps to be selected for record (5 microns) and playback (0.5 microns). A B. & K. frequency response plot supplied by the manufacturers shows a remarkable 35Hz to 20kHz, 3dB response using CrO_2 tape and a 3dB roll off at 18Hz using standard tapes. The expected list price in the U.S.A. is in the region of $800 for model ZETA and between $300 to $400 for the domestic version, to be called SLX. Since Nakamichi only supply the basic deck and electronics, these products are likely to be seen eventually in a number of manufacturers' equipment. Both Rank Wharfedale and Bell and Howell are marketing machines made by Nakamichi. First samples will be delivered to America in October and production on a limited scale will progress from that date.

An interesting footnote is that both products will be available with either Dolby "B" processors or the Philips DNL incorporated. This is the first offer of DNL in high-quality machines that has been seen.

1972 broadcasting convention

The programme for this year's International Broadcasting Convention, which is to be held from 4th to 8th September, includes five main sessions under the following titles: management and engineering training; origination and recording; distribution and satellites; sound broadcasting and transmitters; educational broadcasting, propagation and receivers.

The Convention, to be held at Grosvenor House, Park Lane, London W.1, will comprise the technical sessions and an exhibition of television and sound broadcasting equipment.

New Goonhilly aerial

Goonhilly 3, Britain's new £2¼M satellite communication aerial, has been handed over to the Post Office by its maker Marconi Communication Systems Ltd. Engineers at the station on Goonhilly Downs, Cornwall, are familiarizing themselves with the aerial and its control requirements, to perfect their handling of the giant 97ft dish, on its 62ft concrete tower, and to master the complex telecommunication equipment before bringing the system into service. Faults are being simulated to test the monitoring circuits and automatic facilities for bringing reserve equipment into use. A line-up of the system to work through the Intelsat IV F2 satellite, in geo-stationary orbit over the Atlantic Ocean, will eventually make available 1800 extra telephone circuits (400 at first) for calls to America, Africa and the Middle East.

Chewing it over

A radio transmitter powered by a tiny mercury battery, small enough to fit in the space of a molar, has been developed in America to study the chewing habits of patients. The development work was undertaken by Dr. Irving Glickman, a dental specialist and Tufts University professor of periodontology (the study of diseases of supporting structures around the teeth: tissues, gums and bones).

The transmitter consists of a Hartley oscillator circuit, a Mallory mercury battery and a multilayered switch with sections less than 3/100th of an inch apart. Embedded in a dental bridge, replacing a single tooth, the elements of the device all fit into the artificial molar. The transmitter is activated when a pinpoint gold inlay on the high point of a tooth in the opposing jaw hits one of the switch's individual layers, each of which emits at a different frequency. The signals are then picked up by an aerial and fed to a six-channel oscillograph. The oscillograph recordings tell the dental investigators where the opposing teeth actually make contact, while the radio-equipped subject is chewing foods of different consistencies. This method can locate up to five spots that touch in chewing on that molar, and also distinguishes between single contacts and gliding or rubbing movements. The subjects are also fitted with what looks like a dog collar that has built-in sensors for strain measurements. These are connected through a transducer amplifier to another channel of that same oscillograph. The dentists are also studying how teeth come together every time one swallows.

Nuclear-powered pacemakers

Two patients at a hospital in Buffalo, New York, have received radioisotopic-powered pacemakers which have been jointly developed by Medtronic, Inc., of Minneapolis, and Alcatel, a French firm. The device has an expected lifetime of ten years and has been used in a number of patients in western Euope. Radiation is insignificant and comparable to luminous dial watches.

The first nuclear-powered pacemaker was implanted in a male patient on 2nd March 1971. Designed for long-term bipolar stimulation with mycardial leads, the nuclear-powered pacemaker incorporates the same basic circuitry that is used in an earlier unipolar pulse generator. An electronic convertor has been added to increase the source voltage. The inhibited demand pulse generator has a preset rate of 72 pulses per minute and a fixed amplitude of five volts. The capsule containing the power source is made of tantalum with an inner core of platinum for gamma radiation shielding, sufficient space being provided between the two layers for helium buildup as the plutonium disintegrates. The power source and electronics are encased and hermetically sealed in titanium, shielding it from most sources of magnetic, electromagnetic, and radio-frequency interference. The plutonium 238 power source has a half life of 87 years, theoretically providing sufficient power for at least ten years and it will now be possible to offer patients a device free from the need of frequent operations as required by conventional pacemakers with chemical batteries.

Directional Information in Reproduced Sound

by P. B. Fellgett*

Directional information over a 360° sound-stage can be coded onto two channels, but encounters a difficulty involving the phase shift round the azimuth circle. Methods of coding this information onto ordinary two-channel media are necessarily of this kind, and are not four-channel as is often implied. Three channels enable the phase difficulty to be avoided. Existing techniques of f.m. stereo broadcasting use a modulation bandwidth of three times audio bandwidth, and could carry three-channel 'pantophonic' signals by exploiting both phase and amplitude modulation of the stereo sub-carrier. Tape cassettes can provide four channels, but the usually-assumed 'discrete' pair-wise blended coding is not ideal, and the phase difficulty recurs here, as it does for any even number of channels. The best possibilities for future development seem to lie in three-channel coding, with compromise towards two channels when required by the limitations of the medium. For serious musical applications, the coding must be consistent with a practicable microphone technique for picking up true reverberation and ambience. This condition ensures that 'pop' effects and existing four-channel pair-wise blended coding can be catered for within their respective implicit limitations; the converse is not necessarily true.

The earliest high-quality systems for the reproduction of sound aimed at giving the listener information about the waveform that would have been heard at the original performance, but no information about direction of arrival. Although such one-channel or monophonic systems could feed more than one loudspeaker at the listening position, the signals from all loudspeakers are, of course, equivalent.

The first development of sound reproduction to give directional information employed two channels of transmission, and the two stereophonic channels are implemented on a vinyl disc as two orthogonal directions of stylus motion. These may be regarded either as two orthogonal directions each at 45° to the surface of the disc, interpreted respectively as left and right stereo channels, or equivalently as lateral modulation representing the monophonic signal, and hill-and-dale modulation representing the stereo difference. Similarly in f.m. broadcasting, the mono signal is transmitted as base-band modulation, and the stereo-difference signal is modulated onto a subcarrier.

Stereophonic systems essentially provide directional information over a sound-stage subtending an angle of between about 30° and 90° centered on the direction in front of the listener. It seems certain that the directional information in reproduced sound of high quality, originally provided by the extension from mono to stereo, will be extended to full omni-directional information, or at least an approximation to this that is aesthetically satisfying. Systems which give this information over the full 360° of azimuth may be called 'pantophonic'. (Gerzon has used the term 'periphonic' for systems which give height information as well, but these are not considered here.) A system that is symmetrical throughout the 360°, both at source and in reproduction, may be called strictly pantophonic. This implies that a source at any azimuth will be dealt with equivalently, and similarly that in reproduction there are no preferred azimuths for the loudspeakers as seen by the listener. It cannot be assumed that strict pantophonic systems are necessarily best, but the onus to show an advantage lies with the proposer of any system not having this property.

Current thinking about pantophonic reproduction has followed a rather curious historical path resulting, unfortunately, in the perpetuation of a number of misconceptions. This historical approach derives from the practice of the recording industry of using four-channel master tapes. In so far as such tapes represent simply discrete sources there is of course no difficulty and any competent pantophonic system can accept these inputs as readily as any others, and can associate a chosen azimuth with each of the four channels.

A difficulty however arises from the convention, which appears to have grown without systematic thought being given to it, of associating the four channels with cardinal azimuths (usually 45°, 135°, 225° and 315°) and introducing cross-talk between channels associated with adjacent cardinal azimuths so as to associate particular sources with azimuths intermediate between these cardinal directions.

This pair-wise blending, in imitation of stereo blending between two channels, unfortunately has a number of disadvantages. To anticipate some of the argument which will follow, this method is non-optimal in the sense that it does not exploit the full capabilities of four channels in giving directional information. It has indeed been shown that its capability varies between that of three and of four channels according to the azimuth of the source. Secondly, there are discontinuities in the azimuth-weighting function resulting from the fact that signals are blended only into an adjacent pair of channels. To feed such a system correctly, it would be necessary to use microphones having a sinusoidal lobe of response over 180° of azimuth, and a zero response over the other 180°; such microphones do not exist. Evidently, no subsequent part of the reproduction chain can remove the degradation introduced by pair-wise stereo blending.

Systems of reproduction using four channels of information having audio bandwidth between the source and the listener could correctly be called 'quadrisonic'. To link with the accepted terms 'monophonic' and 'stereophonic', however, the term 'tetraphonic' should be used. The term 'quadraphonic', which has unfortunately gained some currency, is linguistically indefensible.

Linguistics apart, it is bad physics to call any system using the conventional vinyl disc, for example, four-channel. In its present state of development the vinyl disc provides, as we have seen, two channels of information of audio bandwidth. It is not converted into a system having any other number of channels simply by having four, eight, or a thousand inputs, nor by connecting four or any other number of loudspeakers to its output.

The true position is that the vinyl disc in conventional form provides a two-channel system, as does conventional stereo broadcasting. As the stereo sub-carrier has double sidebands, sufficient bandwidth is available to provide a three-channel system. This possibility could be realised compatibly by incorporating the third channel as phase-modulation of the stereo sub-carrier. The limited extension of the bandwidth of the vinyl disc in the TMX system of Cooper and Shiga[1] provides a third channel of restricted

*Professor of Cybernetics and Instrument Physics, University of Reading.

bandwidth. (This could be called 'sesquisonic'.) Tape cassettes are of course in principle capable of providing four or more channels; the maximum number is limited by signal/noise ratio considerations, and pair-wise stereo blending is at a marked disadvantage in this respect in comparison with optimal coding.

Perhaps the most serious misconception is that it is possible to encode four (or three, or more) channels onto the two channels of a vinyl disc, and subsequently recover the information. This however could be done only by magic—it is clearly physically and informationally impossible. An associated fallacy is to suppose that the objective in pantophonic reproduction is to imitate a blended four-channel master tape. If indeed such a tape is the primary source, then an ideal system will evidently sound very similar to this master; some of the defects of the sub-optimal coding may be mitigated but they cannot be overcome. Nevertheless this is not a true statement of the problem. Monophonic reproduction has been referred to as "listening through a hole in the wall". Analogously, stereo provides two holes, and a four-channel master four holes. What is wanted is something different, namely the best directional information that is possible with the available capacity.

For serious aesthetic use, the problem is accordingly to reproduce in the neighbourhood of the listener's head the sound-field he would have experienced at the actual performance. If this is possible, then it is equally possible to associate the signal from each separate microphone in a multi-microphone recording with a prescribed azimuth by an extension of the conventional method of pan-potting so as to imitate sound coming from this direction. Unless a microphone technique is available which enables the true ambience and direction of arrival to be recorded the system is not a candidate for recording serious music, for example church music in which a full 360° sound-stage is traditionally used. Failure to recognize this distinction, or to implement it, in some of the current commercial proposals may account for a large part of the doubt that is still felt by some critical listeners as to the value of extending stereo to pantophonic reproduction.

Consider a microphone at point O in a sound field (Fig. 1). If the microphone is responsive only to pressure, it can record the intensity of the sound at O but nothing about its direction. Sound is however propagated as a longitudinal wave, so that a source at A gives rise to particle-motion along the line AO. The sum of the particle motions due to any number of sources can be resolved along convenient orthogonal directions, say Ox and Oy, and these two signals then carry complete information about the directional properties of the sound-field at O. Unfortunately a source at A′, 180° from A, produces particle motion which also lies in the direction OA, and consequently a 180° ambiguity exists in the absence of a reference of phase. This reference, as may be verified formally from the wave equation, is properly provided by measurement of the pressure at O due to the sound-field.

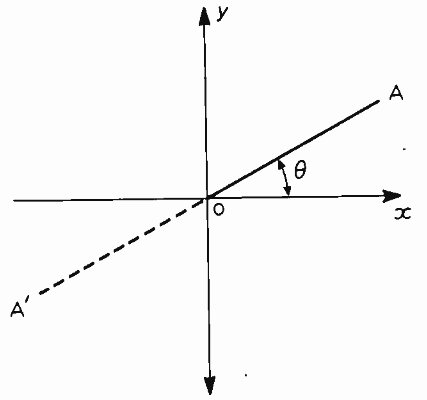


Fig. 1. Particle motions in a sound-field.

It follows that complete and unambiguous information about the sound field at a point can be provided using two velocity-sensitive microphones oriented at right angles, in combination with a pressure microphone, and feeding only three independent channels of audio bandwidth. Any theoretical analysis however represents an abstraction from the real complexities of nature, and the principal approximation in the present case lies in the fact that the human head is not a point. The argument is probably valid for frequencies below about 1kHz but may need modification at high frequencies. Both theoretically and empirically an optimal three-channel system seems to be inferior to four-channel pair-wise stereo blended arrangements by at most an insignificant amount. The argument for three-channel pantophonic f.m. broadcasting appears very strong, especially as full mono and stereo compatibility presents few problems.

In relation to the vinyl disc, the question remains how the three channels can be compressed into the two that are available on the disc. Around 1970 it was noticed by several people independently that the velocity components along Ox and Oy of Fig. 1 could be transmitted as the in-phase and quadrature components of an ambience channel, with the reference of phase being provided by a monophonic channel deriving its signal from the pressure microphone; or equivalently in the simulation by pan-potting. It has been confirmed that this phasor arrangement is capable of giving directional information acceptable in many circumstances. The introduction of phase shifts carries, however, a disadvantage that may be best seen by developing the argument from a different point of view.

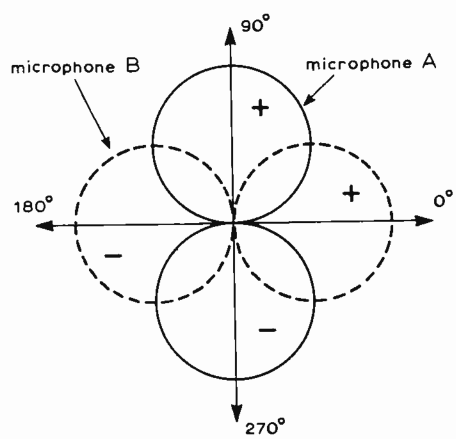


Fig. 2. Responses of microphones in a Blumlein configuration.

Consider two Blumlein microphones for stereo recording having figure-of-eight responses oriented for convenience as shown in Fig. 2. The respective signals from the two microphones from a source at azimuth θ are given by

$$A = a \sin \theta$$

$$B = a \cos \theta$$

The monophonic intensity information is evidently provided by the sum of squares of these two signals, while azimuth information is provided by their ratio in the form

$$\tan \theta = A/B$$

This azimuth information is complete apart from the 180° ambiguity corresponding to this last function being two-valued in the interval 0 to 360°. Provided therefore the sound-stage is confined to the front 180° semi-circle (strictly impossible for reverberation, of course) the original Blumlein microphone arrangement already provides the basis of a fully pantophonic two-channel system, and in this limited sense no further invention was necessary. It is this property that was exploited in the original Hafler surround-sound arrangement.

The 180° ambiguity may be removed, in principle, by mapping the azimuth range 0 to 180° (Fig. 2) onto a 360° circle. This transformation would result in the Blumlein microphone responses of Fig. 3. The amplitudes present no problem, but one microphone is required to change polarity suddenly at azimuth 270° onto which both the old 0° and 180° azimuth have been mapped. More sophisticated arguments show that this unwanted phase reversal is a general property of two-channel systems which cannot be avoided. It appears in various guises in the theory of such systems, and in particular requires a 180° phase shift to be distributed around the circle of loudspeakers used to reproduce the pantophonic sound. The only available choice is how this 180° of relative shift is distributed; the UMX[1] (both BMX and TMX) system distributes it uniformly, Sansui QS with rather less regularity, and CBS SQ with still greater irregularity.

The coding of directional information onto n channels has been analysed by two principal methods; the circular harmonic method of Cooper and Shiga[1], and the representative sphere or energy sphere of Scheiber[2] and Gerzon[3]. Both methods are applicable to any number of channels, but the energy sphere is particularly useful for two-channel systems because the sphere is then an ordinary three-dimensional one which can be more readily visualized than the hyperspheres needed in general.

According to this method, the two channels A and B are taken as normalized so that a source of unit strength gives rise to the same energy $A^2+B^2 = 1$ irrespective of azimuth. As only relative phase between the two channels is of interest, absolute phase

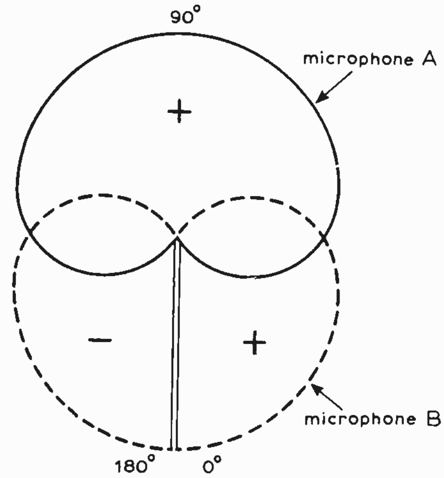


Fig. 3. Responses which result when the front 180° region (top part in the diagram) of Fig. 2 is mapped into 360°.

being irrelevant, the amplitudes in the two channels may be written without loss of generality as

$$A = \sin(\alpha/2)\exp(i\beta/2)$$
$$B = \cos(\alpha/2)\exp(-i\beta/2)$$

in which the angles α, β are defined as spherical co-ordinates on the representative sphere. Each azimuth θ corresponds to a point on the sphere according to the relations $\alpha = \alpha(\theta)$, $\beta = \beta(\theta)$ and the problem of encoding is to choose the function α and β which define the locus described on the sphere as θ explores the 360° of azimuth.

Similarly in decoding a representative point on the energy sphere is defined in spherical co-ordinates by

$$\text{output} = A\cos(\alpha'/2)\exp(-i\beta'/2) + B\sin(\alpha'/2)\exp(i\beta'/2)$$

in which α' and β' represent the amplitudes and relative phases of the respective contributions from the A and B channels to a given output of the decoder.

The energy sphere provides a geometrical picture from which many relations can be clarified. In the first place, fallacious 'four-channel' thinking has concentrated the attention of some designers only onto four points corresponding to cardinal azimuths. In fact of course the representative point of the encoder will explore the full 360° of azimuth, whether this was intended consciously or not, either in a system using a microphone technique to record the true ambience and direction of arrival of the sound, or in one in which separate microphones are pan-potted. If four (or indeed any other number) of points in a pair-wise blended system are fixed, the locus between these points is dependent on the phase-shifts between adjacent channels. Phase shifts of 90° between four cardinal channels have been used, for example, in the Sansui system to control the intermediate locus. (Reference to the block diagram shows that these 90° phase shifts may properly be regarded as outside the encoding matrix itself.)

A systematic approach is evidently desirable, and two forms of locus deserve initial consideration, the great circle and a locus passing through tetrahedral points. The 'modified Blumlein' amplitude coding described in connection with Fig. 3, and phasor systems such as BMX[1], give great circle loci. Indeed these two methods of encoding are informationally equivalent in the sense that one can be converted into the other, or *vice versa* at the decoder by means of the linear operations of 90° phase shift combined with addition and subtraction. For example the amplitude-coded signal can be converted into a phasor-coded one according to the following relationships

$$A = \sin(\alpha/2) \qquad B = \cos(\alpha/2)$$
$$B + iA = \cos(\alpha/2) + i\sin(\alpha/2) = \exp(i\alpha/2)$$
$$B - iA = \cos(\alpha/2) - i\sin(\alpha/2) = \exp(-i\alpha/2)$$

Although tetrahedral arrangements of cardinal azimuths have been considered, there seems to have been little or no discussion of the locus between these points. A desirable locus would be roughly in the shape of the seam on a tennis ball, crossing a great circle at four points half-way between the cardinal azimuths.

It can be shown that the cross-talk between decoder outputs is given by

$$-20\log_{10}\{\cos(\Omega/2)\}$$

in which Ω is the angle subtended at the centre of the sphere by the points representing the decoding of the two channels. In terms of energy, this formula represents a cardioid of revolution—rather like the top bun of a cottage loaf—with maximum at the representative point and zero 180° away from it on the sphere. It is important to note that separation is purely a property of the decoder, irrespective of the encoding which feeds it.

Separation has been an obsession in the discussion of pantophonic proposals, probably by false analogy with stereo in which cross-talk between channels represents a degradation of the intended system. In pantophonic reproduction however the objective is not to provide high separation between an irrelevant and in general mythical four-channel source, but to reproduce at the listener an approximation to the directional properties of the sound-field that would have been heard at source. Imitation of stereo operation is not the ideal way of achieving this, as stereo blending is itself non-ideal.

Consider a listener at the centre in Fig. 4 listening to a front-centre signal presented, as some authors have supposed to be desirable, as equal signals from loudspeakers *a* and *b*, while *c* and *d* remain silent. It is easy to see that in fact the particle velocity is along the desired centre line, but the pressure due to the sound-field is 3dB too high. This phenomenon may be the explanation of the slightly oppressive 'overhead' quality experienced by some listeners to normal stereo. Even with loudspeakers oriented so that one is in the front-centre direction as in Fig. 5, feeding a signal to this speaker alone is not the only way of establishing particle velocity along the centre line, accompanied by a correctly related pressure field. Fig. 5(a) illustrates another out of the infinity of ways in which this

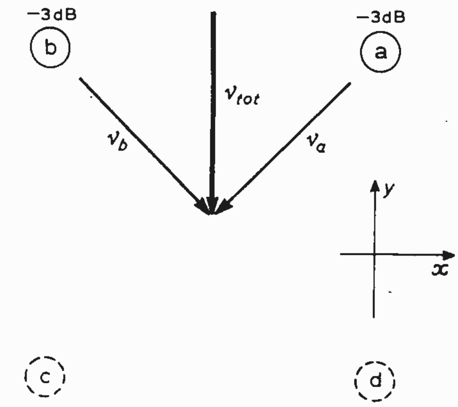


Fig. 4. Stereo speakers do not give exactly correct relationship between air pressure and velocity for a front centre source. For $a = b = 0.707$, $v_x = 0$, $v_y = 1.00$ and $p = 1.414$.

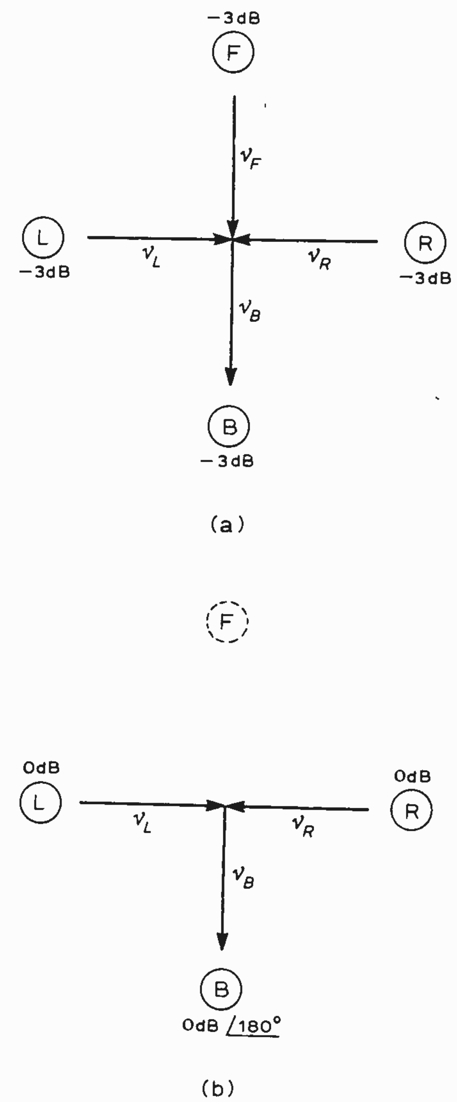


Fig. 5. Examples, one pathological, of configurations which give correct relationship between pressure and velocity for front centre source. $v_x = 0$, $v_y = 1$ and $p = 1$ not only when $F = 1$, $L = R = B = 0$, but in general when $F - B = 1$, $F + L = 1$ and $L = R$. Cases illustrated are for $F = L = R = \frac{1}{2}$, $B = -\frac{1}{2}$, and $F = 0$, $L = R = 1$, $B = -1$.

could be done, and Fig. 5(b) illustrates a particularly pathological example. It may rightly be objected that an extended listening area must be considered, and that the sound-field in the neighbourhood of the listener's head cannot be taken as equivalent to the behaviour at a point, particularly for frequencies above 1kHz. For these reasons it is certainly not recommended that the arrangement shown in Fig. 5(b) should be deliberately used.

The arrangement shown in Fig. 6, in which the speakers are once again placed in the 45° cardinal position, may however be preferable to 'infinite separation' between front and rear speakers, particularly below 1kHz.

The last equation shows that the false goal of infinite separation can be attained between any two decoder outputs by simply placing the representative point of each at the zero response of the other. There is however then a limit to the separation that can be achieved between both of these channels and any third channel. Table 1 shows that if

Table 1

Separation X to Y and X to Z (dB)	Maximum separation Y to Z (dB)
3	∞
4.5	10.7
10	6
∞	1.95

the separation between output X and both of outputs Y and Z is to be at least as great as that shown in the first column, then the separation between Y and Z is limited to the value shown in the second column. These relationships are further illustrated in Fig. 7, which shows as a function of azimuth the separation between adjacent pairs and across diagonals of the 'high-separation' tetraphonic coding proposed by Scheiber when the full pantophonic locus is considered. It may be doubted whether the extra complication and asymmetry of tetrahedral encoding, which in any case increases adjacent-speaker separation at the expense of the diagonals, is worthwhile over the whole locus. (Claims for separations exceeding those implied by the crosstalk equation can be true only in the sense of being achieved by non-linear or gain-riding techniques, the limitations of which appear as unacceptable for serious music as were those of processed stereo.)

The implication of this last equation for great-circle encoding is that each output of the decoder, which may be connected to its own loudspeaker, behaves as if it were effectively connected to a source microphone having a cardioid directional characteristic. It would be at most pointless, and probably undesirable, if it were possible to increase the directivity of this polar so that it exceeded the directional discrimination of practicable studio microphones. A three-channel pantophonic system can give 6dB between decoder output associated with azimuths 90° apart and complete separation across diagonals, or alternatively this polar can be slightly sharpened at the expense of a small secondary lobe in the rear direction.

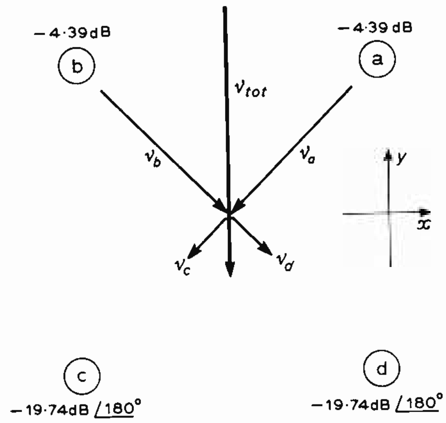


Fig. 6. Theoretically desirable configuration giving correct relationship between pressure and velocity. $v_x = 0.707\ (v_a - v_b - v_c + v_d)$, $v_y = 0.707\ (v_a + v_b - v_c - v_d)$. *For* $a = b = 0.603$, $c = d = -0.103$, $v_x = 0$, $v_y = 1.00$.

Table 2 shows the number of zeros on the effective polar of a decoder output that can be achieved with different numbers of channels. The phase shift problem, discussed in connection with Fig. 3, applies not only to two-channel systems but to all systems having an even number of channels, as stated in the third column of Table 2. This table suggests that three-channel pantophonic systems have a special claim to consideration, and underlines again the desirability of developing f.m. broadcasting in this mode.

In the original development of stereo reproduction, it was taken for granted that no adequate test of the new system could be made without adapting the studio technique to exploit the directional information that was then being provided for the first time. Pantophonic sound has unfortunately often been denied the courtesy of this logical necessity, and attempts have been made to express compatibility with existing systems in terms of the use of a sound-stage restricted by the limitations of conventional stereo. At the present stage of development, the source material is largely under the control of the proponents of any given

Table 2

No. of channels	No. of zeros in the output polar	Phase shift present
1	0	—
2	1	yes
3	2	no
4	3	yes

system, and may be chosen to illustrate the good points of the latter while concealing its defects. A compatible pantophonic system should fulfil the following requirements with good approximation:

- It should be capable of exploiting a full 360° sound-stage. In a demonstration, the material should be such as to exploit this capability artistically, for example church organ music with detached choir and echo organs.
- It must be capable of reproducing correctly the direction of arrival of sound at the original performance, including reverberant sound. The ability to pan-pot separate microphones may be taken for granted if this condition is fulfilled.
- A good monophonic signal should be available without the use of a special decoder. This signal should give equal weight to all azimuths, or nearly so.
- Played in stereo mode without decoder,

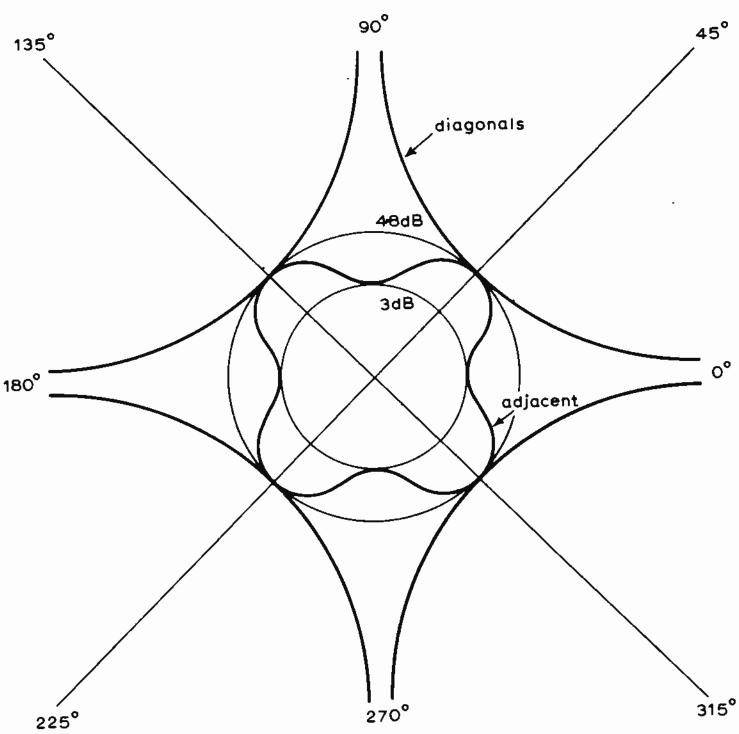


Fig. 7. Characteristics of Scheiber 'high separation' tetraphonic coding used pantophonically. Extra 1.4-dB separation between adjacent speakers is realizable only in the four cardinal directions, and is at the expense of separation across diagonals.

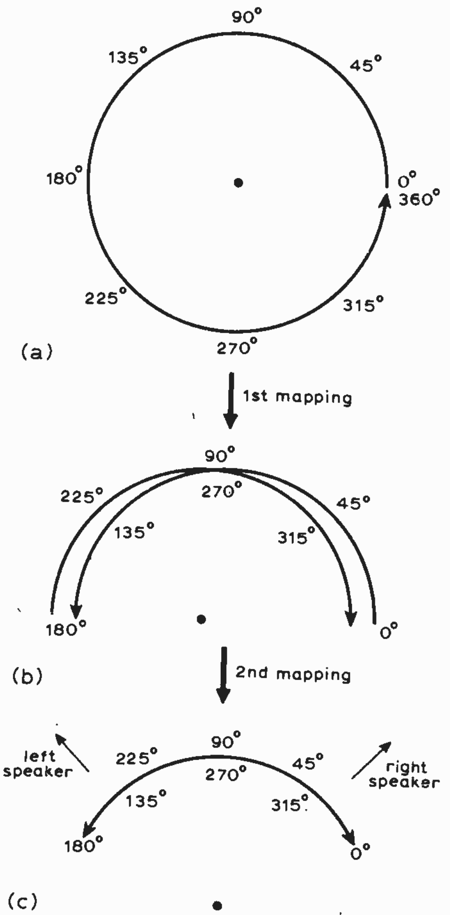


Fig. 8. Way in which a system for reproducing 360° sound stage (a) can be made stereo compatible by first mapping rear semicircle onto front (b) and then compressing to range encompassed by stereo speaker positions (c).

the front 180° of sound-stage should appear between the two stereo speakers (with perhaps some further additional spread from 'stereo enhancement' if used). The rear 180° should be superposed on the front hemisphere by mirror imaging in the diameter joining these hemispheres.

- The system should be capable of being played pantophonically with the, logically, minimum of three decoder outputs feeding three loudspeakers, and there should be progressive subjective improvement as the number of decoded signals is increased to four or more. This requirement follows on from the practice of using an intermediate speaker in conventional stereo; four loudspeakers fed from either three or four decoder outputs may be a convenient compromise between aesthetics and cost, but there is nothing magical about the number four in this connection and the use of six to eight speakers (not all covering the full bass range) may be required.

The fourth requirement is illustrated in Fig. 8. The true 360° sound-stage, illustrated in Fig. 8(a), should be reproduced in pantophonic play-back. In stereo mode, it is inevitable that front-to-back information is lost, and the only practicable mapping seems to be to that shown in Fig. 8(b) in which corresponding front and rear azimuths are superposed. If the loudspeakers are placed in the conventional stereo positions, the hemisphere of Fig. 8(b) is necessarily compressed to the region shown in Fig. 8(c) lying approximately between the left and right speakers. This compression would of course give an undesirably narrow stereo image if a studio technique had been used which confined the sound-stage to the stereo limitation of the front quadrant only. If however a correct pantophonic studio technique has been used, in which instrumentalists may occupy at least the front semi-circle, then this compression will be harmless and will give the stereo listener a good approximation of what he is used to. The inevitability of mapping operations closely approximating those shown in Fig. 8 has been missed by those who have concentrated attention on cardinal directions only, with the possible addition of stereo blending over the front quadrant, instead of considering the whole of the pantophonic locus.

Most, or all, of the theory on which the above discussion is based was afterwards found to have been anticipated in the more general and elegant theoretical analysis developed by Michael Gerzon[3] and incorporated in a series of circulated reports. The author has benefited from these reports, and from helpful discussions with Mr. Gerzon for which acknowledgement is gratefully made. Any errors in this article are however the author's own.

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1. Cooper, D. H. and Shiga, T. 'Discrete Matrix Multichannel Stereo'. Audio Engineering Society, Inc. Munich Convention, 1972, also Los Angeles Convention, 1972.

2. Scheiber, P. 'Analysing Phase-Amplitude Matrices'. *J. Audio Eng. Soc.*, vol. 19, 1971, p. 835.

3. Gerzon, M. 'Matrix Systems for Four-Speaker Stereo I'. Privately circulated report, 1971.

Announcements

The Institution of Electrical Engineers has organized a residential vacation school on *"Lasers and Optical Electronics"* to be held at the University of Southampton from 10th-22nd September. Information from the Divisional Secretary LS(S), I.E.E., Savoy Place, London WC2R OBL.

The Royal Television Society is introducing a series of short training courses the first of which will cover *digital techniques in television engineering.* The nine lectures will be given on consecutive Wednesday evenings beginning on October 25th at University College, Gower Street, London WC1. The fee for non-members is £7.50. Application forms from Mrs. Jill Cousins, R.T.S., 166 Shaftesbury Avenue, London WC2H 8JH.

Three courses to be held at the Polytechnic of North London are: *"High Quality Sound Reproduction"* commencing October 26th at 18.30, fee £6.30; *"Sound Studios and Recording"*, October 26th at 14.30, fee £10.50; *"Television Engineering"*, a post-graduate course, October 3rd at 17.30, fee £6. Further details from the Head of Department of Electronic and Communications Engineering, Polytechnic of North London, Holloway Road, London N7 8DB.

An advanced *course in noise and vibration* is being conducted by the Institute of Sound and Vibration Research at the University of Southampton from 18th-22nd September, the fee, including accommodation, is £50. Registration forms etc. from Mrs. O. G. Hyde, I.S.V.R., University of Southampton, Southampton SO9 5NH.

A post-graduate *evening course on integrated circuit electronics* will be held at North East London Polytechnic commencing November 2nd. Fee £6. Details from The Engineering Faculty Registrar, North East London Polytechnic, Barking Precinct, Longbridge Road, Dagenham, Essex, RM8 2AS.

A 32-week evening course entitled *"Non-destructive Testing"* is to be held at Croydon Technical College starting in the Autumn. Fee £3. further details from Head of Science Department, Croydon Technical College, Fairfield, Croydon, Surrey, CR9 1DX.

A course of ten *lectures on audio techniques* will be held at Norwood Technical College at 18.30 each Tuesday commencing October 3rd. Fee £3. Further information from the Senior Administrative Officer, Norwood Technical College, Knight's Hill, London SE27 OTX.

"Confidence in Measurement" is the title of a 16mm colour film which describes the work and facilities of the British Calibration Service and is available for hire or purchase from the Central Film Library, Government Building, Bromyard Avenue, Acton, London W3 7JB.

A ten-week evening course on *modern electronic techniques* will be held at Portsmouth Polytechnic commencing October 12th. Fee £5. Details from D. Meek, Administrative Assistant, Portsmouth Polytechnic, Department of Electrical and Electronic Engineering, Burnaby Road, Portsmouth PO1 3QL.

Mr. M. R. Lord, 7 Dordells, Basildon, Essex, would be pleased to hear from anyone interested in the formation of an *Amateur Computer Constructors' Society.*

Acoustic Research are to open a new assembly plant in London for their range of audio products on October 2nd. It will supply products for the U.K. market and be the base for service operations. The significant feature of this move is that the price of their range of products will be cut by 17% to 25%. The AR tuner, not previously available in the U.K. will be introduced at £110 together with a new loudspeaker unit designated the AR7 and priced at £26.

The signing of an agreement appointing Celdis as Hewlett-Packard's sole U.K. distributor for their range of *opto-electronic components* has been announced. The existing H-P marketing organization will continue but will concentrate on "the big-order end of the business." The products Celdis will be stocking are light-emitting diode numeric and alpha-numeric displays, solid-state lamps, emitters of visible light, and isolators made up of emitter-detector pairs.

Available in the U.K. from Tranchant Electronics (U.K.) Ltd, Tranchant House, 100a High Street, Hampton, Middlesex, TW12 2ST, is the Intersil series of high slew-rate *operational amplifiers,* electrically and pin-for-pin compatible with the Harris HA2500 Series.

The Plessey Company, Poole, Dorset, has signed a marketing agreement with the Frederick Electronics Corporation, of Maryland, U.S.A., covering the sale and service of *data and telegraph equipment* in the U.K. and a large number of countries overseas.

Burndept Electronics (ER) Ltd, have acquired the *mobile radio-telephone* business previously carried on by Ultra Electronics Ltd, for the sum of £350,000.

The electronic component distributors SDS-WEL Components Ltd, of Gunstore Road, Hilsea Trading Estate, Portsmouth, Hants, have changed the name of the company to *SDS Components Ltd.*

B & W Electronics have moved from Littlehampton Road to a factory at Meadow Road, Worthing, Sussex, BN11 2RX.

From 1st September, *Aveley Electric Ltd*, of South Ockendon, will operate from premises at Roebuck Road, Chessington, Surrey. (Tel: 01-397 8771)

Electronic Building Bricks

27. The channel

by James Franklin

When information is transmitted from point P to point Q (Fig. 1) via some medium, the overall system — which may be a radio link, a cable (perhaps containing repeaters) or a chain of electronic units (Part 11) — can be regarded as a building brick with definable properties; namely a *channel*. The physical distance between P and Q may be half-way across the world or merely from one part of an electronic equipment to another. What mainly concerns the electronics engineer is the *capacity* of the channel to communicate information — that is, the maximum rate at which it will convey information in bits per second (see Part 15). This is the main "definable property" of the channel, mentioned above. Other definable properties are certain physical quantities which determine the rate at which information can be conveyed.

What are these physical quantities? Two of them have been discussed already in this series — the electrical power in the signal (Part 8) and the electrical power in the noise introduced by the transmission medium and apparatus (Part 24). As explained in Part 24, it is the relative values of these two (average) powers — the signal-to-noise ratio so called — which is important in determining the accuracy of the information in the received signal. The presence of noise sets a limit on the number of distinct "levels" in the signal waveform (Part 15) that can be detected. The signal power available depends on what can be economically transmitted.

Another factor determining the accuracy of signal transmission is the ability of the channel to convey the variations in the electrical quantity (say current or voltage) which represent the signal information — more specifically the *rate of change* of these variations. For example Fig. 2 shows parts of two signal waveforms, each being a transition between two steady voltage values. In (a) the rate of change of voltage with time at the transition is higher than the rate of change of voltage in (b). The information channel has physical properties called resistance and reactance (Part 7) which together limit the rate of change of an electrical quantity which can be conveyed from point P to point Q. Consequently if the signal waveform at the source P contains higher rates of change than the channel can transmit, the waveform received at Q will be distorted and hence the channel will introduce errors. For example, the transition (a) of Fig. 2 could be considered as part of a signal generated at P. If the channel is incapable of transmitting this rate of change the transition received at Q will be something like (b).

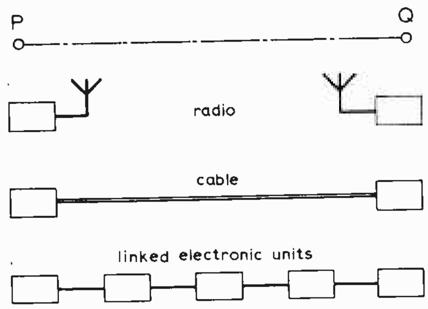

Fig. 1. Three types of information channel.

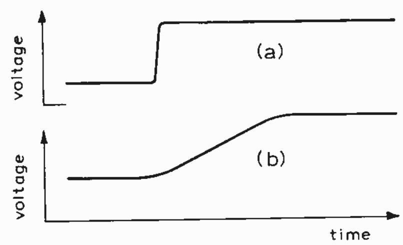

Fig. 2. Parts of signals illustrating rates of change of voltage with time.

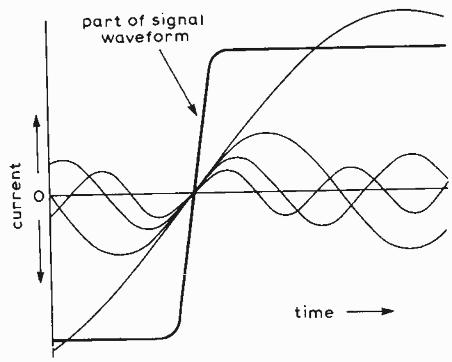

Fig. 3. How a signal wave-form may be analysed into sine-wave components. (Not all the components are shown.)

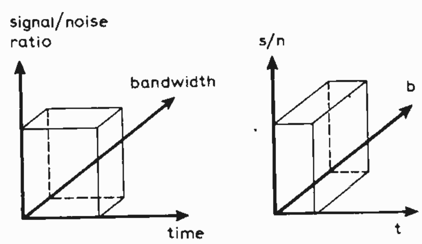

Fig. 4. In transmitting a given amount of information (volume of "box"), the time taken, bandwidth and signal/noise ratio may be "exchanged". Here an exchange of time and bandwidth is illustrated.

The ability of the channel to convey such rates of change of electrical variables is measured in a quantity called *bandwidth*. This term requires some explanation. Any signal waveform can be analysed into a number of component sine-wave oscillations, equivalent to harmonic motion (Part 10), of different magnitudes, frequencies and phases (Part 13)*. This is illustrated in Fig. 3, which shows some — by no means all — of the sine-wave components which, when added together, constitute the signal waveform. Each component has a maximum rate of change. A channel will convey a limited range of these sine-wave components, and the bandwidth of the channel is the extent of the continuous range, in hertz (cycles per second), over which the channel transmits a specified proportion of the original signal power. For example, the bandwidth of a typical telephone channel is about 4000 hertz.

The channel's bandwidth therefore indicates the rates of change that can be transmitted and, as a result, the number of independent "levels" of a waveform that can be conveyed by the channel in a given time. Because of this it is a factor in determining the maximum information rate in bits/second.

So signal power, noise power and bandwidth together determine the highest rate at which a channel can convey information. The exact relationship of the three quantities to this maximum rate is given by a formula†, which enables the channel to be designed to "match" a given source of information, or vice-versa.

In some electronic systems it may be possible to adjust the rate at which information is generated at the source — that is, to alter the length of time available to convey a given number of binary digits (Part 15). This enables alterations to be made to the required bandwidth and/or required signal-to-noise ratio. A graphical illustration of this interdependence, due to G. G. Gouriet, is shown in Fig. 4. The volume of the "box", the triple product of time, bandwidth and signal/noise ratio, is proportional to a given number of bits of information. Obviously it is possible, as shown, to change the time available, the bandwidth and the signal/noise ratio (actually its $\log_2$) in various ways that will maintain constant the volume representing the number of bits of information.

*This process is known as Fourier analysis, because the component sine-waves are expressible mathematically as the terms of a Fourier series (named after J. P. J. Fourier, French mathematician and physicist 1768-1830).

† Derived by C. E. Shannon, $C = W \log_2 [(S^2 + N^2)/N^2]$ bits/second, where C = maximum capacity of channel, W = bandwidth in hertz, S^2 = mean signal power in watts, N^2 = mean noise power in watts.

Synchronous Detection in Radio Reception -1

*by Pat Hawker**, G3VA

Increasing use is being made of various forms of synchronous, or coherent, detection in radio communication, broadcasting and instrumentation. For a decade now, product detectors have been fitted in general purpose communications receivers; synchronous detectors form an essential part of stereo and colour television decoders; there is considerable interest among radio amateurs in 'direct-conversion' receivers as an alternative to the superhet; the availability of complete phase-locked loop detectors in integrated circuit form — these are all examples of this trend.

Again, the advantages of synchronous demodulation when applied to vestigial sideband television signals have led to the development of special synchronous detectors for high-quality re-broadcast receivers. For the future, it seems feasible that the phase-locked loop and related techniques will open the way for much wider use of double-sideband suppressed-carrier transmissions for mobile communications, s.s.b. broadcasting or relatively narrow-band v.h.f. /a.m. broadcasting. The potential of such systems as the 'bi-aural', synchronous, exalted-carrier detector – which will be described in Part 2 – is already being stressed in some quarters. This list could readily be expanded, but in listing what is technically possible, there is the danger of underestimating the inflexibility of broadcasting systems and standards, resulting from the massive investment by the public in existing systems. No matter how many advantages may be claimed for synchronous detection, it would be misleading to suggest that the days of the simple diode envelope detector or, even more, the well-established superhet receiver are now numbered. Nevertheless, the time is ripe to review – in non-mathematical terms – some aspects of this growing interest in synchronous detection and to outline how this may influence receivers for broadcasting and amateur radio communications.

One of the most attractive features of a phase-locked loop synchronous detector is its flexibility: it can be designed to cope with a.m., s.s.b., d.s.b.s.c., f.m., n.b.f.m., c.w. and r.t.t.y. (radioteleprinting). In addition, it has long been recognized that synchronous detection provides much improved signal/noise performance at the very low input levels where the diode envelope detector is notoriously inefficient — Fig. 1. At low s/n ratios the envelope detector distorts or may even lose the intelligence signals. The synchronous detector preserves the s/n ratio and thus makes possible the use of very effective post-detector signal processing, allowing recovery by integration of certain types of signals even when these are buried deep in the noise.

For the broadcasters the attraction of synchronous detection is the flexibility it would give receivers, opening the way to the use of different modes. On the other hand, work[1] by the B.B.C. Research Department, carried out on behalf of the B.B.C. and I.B.A., emphasized the practical problems involved in attempting to adopt synchronous detection in, say, simple portable broadcast receivers. This showed that the marginal benefits on a.m. would hardly compensate the listener for the extra cost and increased power consumption. Yet clearly some form of synchronous detection will be essential if the ordinary listener is ever to be offered such spectrum-saving modes as s.s.b. or relatively narrowband v.h.f. /f.m.

The performance of the diode detector can be improved on weak signals by exalted carrier techniques, which can be regarded as a form of synchronous detection. In this system a locally generated carrier is added to the incoming signal to ensure that the diode detector works at its most efficient level.

Synchronous detection is essentially a linear frequency conversion process. The r.f. or i.f. signal is heterodyned by the original carrier frequency and then passed through a low-pass filter to remove the r.f. components, so that the modulation products are converted back to their original frequencies. To improve dynamic range and to limit the number of unwanted products, it is an advantage if the heterodyne or product detector is balanced.

When the incoming signal at r.f. is applied to the synchronous detector, without first being converted to an intermediate frequency, the arrangement is frequently called 'direct-conversion' – Fig. 2.

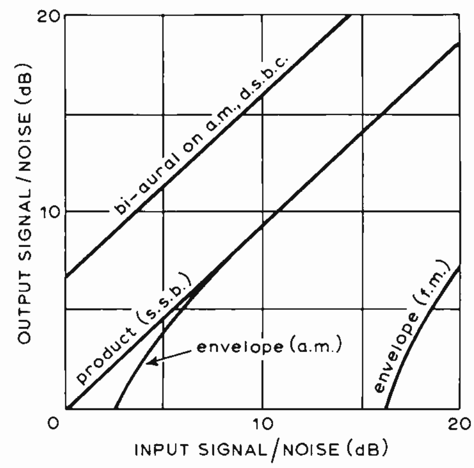

Fig. 1. Effect of demodulators on signal/noise ratios (after Haviland).

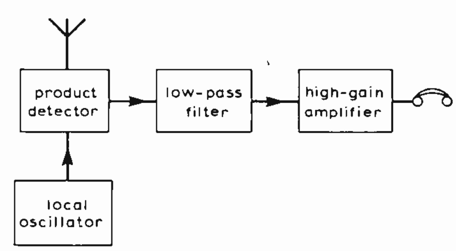

Fig. 2. Basic form of direct-conversion receiver.

A carrier is needed for both envelope and product detection; this carrier can be radiated along with the sidebands, as in a.m., or locally generated and inserted in the receiver for suppressed carrier systems. Any difference in frequency between the inserted carrier and the original carrier results in a frequency shift in the intelligence signal. Investigations have shown that for speech communication, the amount of shift that can be tolerated depends on the direction of the shift and the signal/noise ratio; but typically a shift of up to about 100 to 300 Hz will not seriously degrade speech intelligibility, particularly to an ear attuned to frequency-shifted speech.

Thus for s.s.b. speech it is common practice to use synchronous detection in the simple form of a product detector and

* Independent Broadcasting Authority

free-running carrier insertion oscillator. For the reception of music, much closer agreement between the modulating and demodulating carriers is essential, around 2Hz or better. Even less tolerance is demanded by some specialized forms of s.s.b. transmission such as Lincompex speech or Piccolo telegraphy.

If s.s.b. is ever to be widely used for broadcasting* a conventional product detector of the type used in communications receivers is out of the question. One of the more complex forms of synchronous detection with simple-to-operate means of locking the re-inserted carrier – virtually in phase-coherence to the original modulating carrier – must be used; or alternatively some related form of a.f.c. provided. Provision of a.f.c. on systems with the carrier suppressed to a very low level clearly presents difficulties but a recent suggestion by Villard[2] shows that phase-locked loop and/or zero-crossing techniques might be used, at least for communications applications.

For synchronous demodulation of a.m. or double-sideband suppressed-carrier transmissions there are two main approaches. A modern communications receiver fitted with a good s.s.b. crystal or mechanical filter can resolve d.s.b. by filtering out one sideband and the carrier if present and then seeing the signal as s.s.b. at the conventional product detector. This, however, results in the loss of the potential advantages which arise from coherent demodulation of two sidebands, including greater resistance to narrow-band interfering signals. Haviland[3] has emphasized that if one is to make a true assessment of the "figure of merit" of different modulation systems it is essential to take fully into account the form of demodulation used in the receiver and its performance under conditions of random interference.

To take advantage of the presence of two sidebands, we need to insert a fully phase-coherent carrier, that is to say the locally generated carrier must be within a few degrees of phase of the modulating

*See, for example, G. Wareham's article pp358-63 August issue.

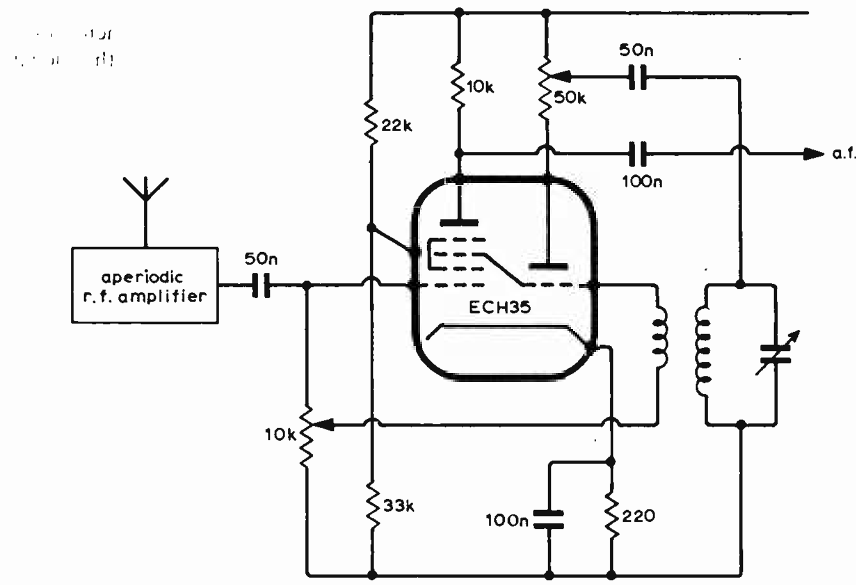

Fig. 3. Synchrodyne mixer detector described by Tucker in 1947.

carrier. As noted earlier with a.m. we obtain similar benefits by using such a carrier to provide enhanced-carrier demodulation. A local tunable or crystal-controlled oscillator cannot be maintained to this degree of accuracy unless some form of synchronizing control is included.

In its simplest form, synchronization may take the form of forcing a phase lock on a free-running oscillator by applying a portion of the incoming carrier to the oscillator, if one is available. (Even with suppressed-carrier systems a weak carrier may be available by careful filtering.) This approach formed the basis of the Tucker "synchrodyne" direct-conversion receivers.[4]

These designs showed that in practice effective phase coherence could usually be achieved by feeding a little of the incoming signal to the oscillator over which it assumes control. This technique had been known for many years, but Tucker showed that it could form a satisfactory basis for broadcast receivers of varying complexity, demodulating the incoming r.f. signal directly to a.f. without intermediate frequency amplification. (Almost all the techniques which in recent years have been applied to direct conversion amateur receivers were foreshadowed either in the original Tucker articles or in the correspondence to which they gave rise.)

The synchrodyne receiver thus consisted of an optional signal-frequency amplifier, a frequency-changer stage (product detector plus synchronized local oscillator), a post-detector audio filter which effectively governed the selectivity of the receiver, followed by a high-gain audio amplifier. It represented a form of "straight" (t.r.f.) receiver but, because of its linear form of demodulation, permitted the selectivity to be governed by the audio filter without the problems of cross-modulation and blocking which arise when attempting to do this with a conventional t.r.f. receiver.

In his articles, Tucker presented receivers using various forms of synchronous demodulators: a simple triode-hexode frequency-changer demodulator (Fig. 3); a double-balanced ring-type

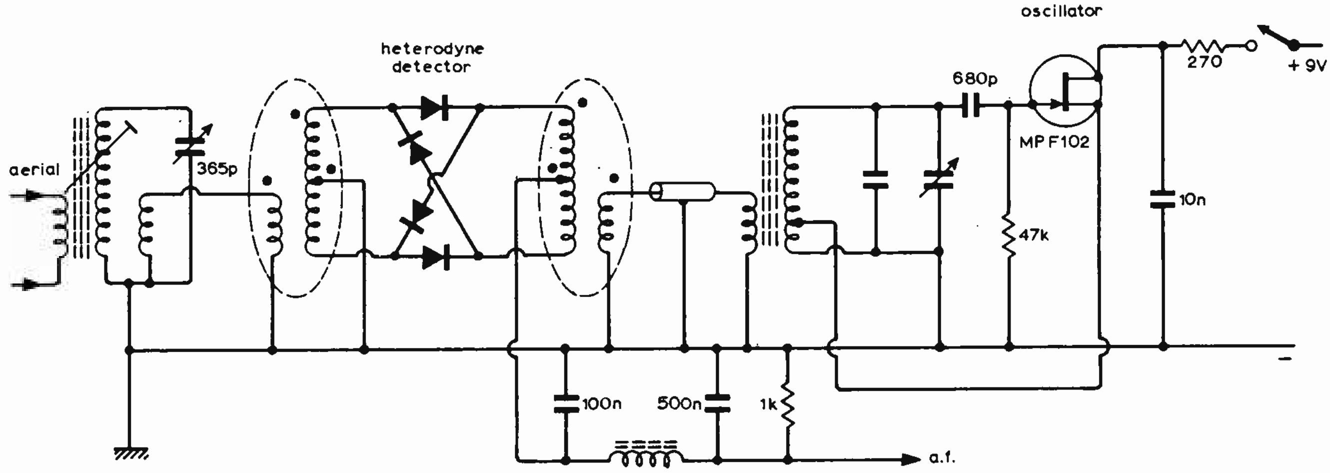

Fig. 4. Double-balanced ring demodulator with f.e.t. local oscillator in the Hayward and Bingham 3.5MHz direct-conversion receiver.

demodulator; and single-balanced Cowan-type four-diode demodulator. At least one of these receivers was demonstrated at one of the early post-war London Radio Shows, and a number were built by home constructors; I vividly recall dabbling with one. However some constructors experienced difficulty in ensuring that the local oscillator synchronized effectively; another common criticism was the heterodyne whistles generated during tuning. As far as I have been able to discover, no commercial models were produced.

Among the correspondence generated at the time was the suggestion by Apthrope[5] that it should be possible to lock the oscillator by using a frequency twice that of the carrier. Such a signal can be derived by full-wave rectification of a proportion of the incoming signal. This technique has recently been revived, as an alternative to the phase-locked loop, by Macario[6] in his investigation of synchronous demodulation for h.f. and v.h.f./d.s.b.s.c. applications.

The synchrodyne era also produced another suggestion: that one sideband could be phased-out by the use of quadrature two-phase techniques. This system was later used by Costas (see below) and is being applied by a number of amateurs for high-performance direct-conversion receivers. Undoubtedly, the synchrodyne was yet another example of techniques a little ahead of the technology; widespread use was to await the development of semiconductor devices.

The development was also influenced by the coming of amateur s.s.b. and the greater use of phasing techniques for s.s.b. generators and add-on demodulator units. Villard[7] pointed to the use of balanced product detectors to allow much more effective use to be made of post-detector audio filtering, and his ideas formed the basis of the first simple direct-conversion receiver presented by White[8] specifically for amateur reception of c.w. and s.s.b. signals. Phasing-type single sideband demodulators never achieved wide use, largely owing to the development of effective mechanical and crystal s.s.b. filters, but a number were described, including several by General Electric (U.S.A.) engineers such as the Signal Slicer[9].

But the most powerful advocate of synchronous systems and direct-conversion receivers during the 1950s was undoubtedly J. P. Costas, also of General Electric. In the issue of December 1956 of *Proc. I.R.E.*, devoted almost entirely to s.s.b., he struck an "odd-man-out" attitude in showing that the main arguments in favour of s.s.b. were based on conventional demodulation, and would not apply if receivers fully utilized synchronous demodulation[10]. He outlined, as Tucker had done, the advantages of direct conversion and gave some details of an experimental high-performance (and clearly very complex) synchronous receiver — the AN/FRR-48 (XW-1). This complexity was largely because of the use of a frequency synthesizer of that period; it also used two-phase synchronous demodulation, phase-locking the local oscillator by the use of an a.f. phase discriminator.

Costas pointed out that the direct-conversion receiver eliminates the basic superhet problem of image response as well as providing the opportunity to use economical post-detector filtering to achieve extreme selectivity and be readily switchable. Despite a later blast at s.s.b.[11], Costas' advocacy of d.s.b.s.c. and direct-conversion phase-locked receivers had little immediate effect on professional communications. Even today s.s.b. is often credited with the higher communications efficiency and more economical use of the

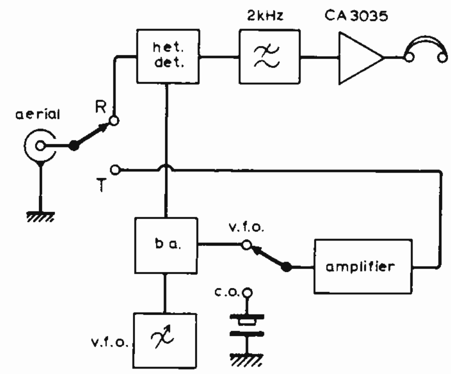

Fig. 5. Simple transceiver using the same oscillator for synchronous detection and for the transmitter v.f.o.

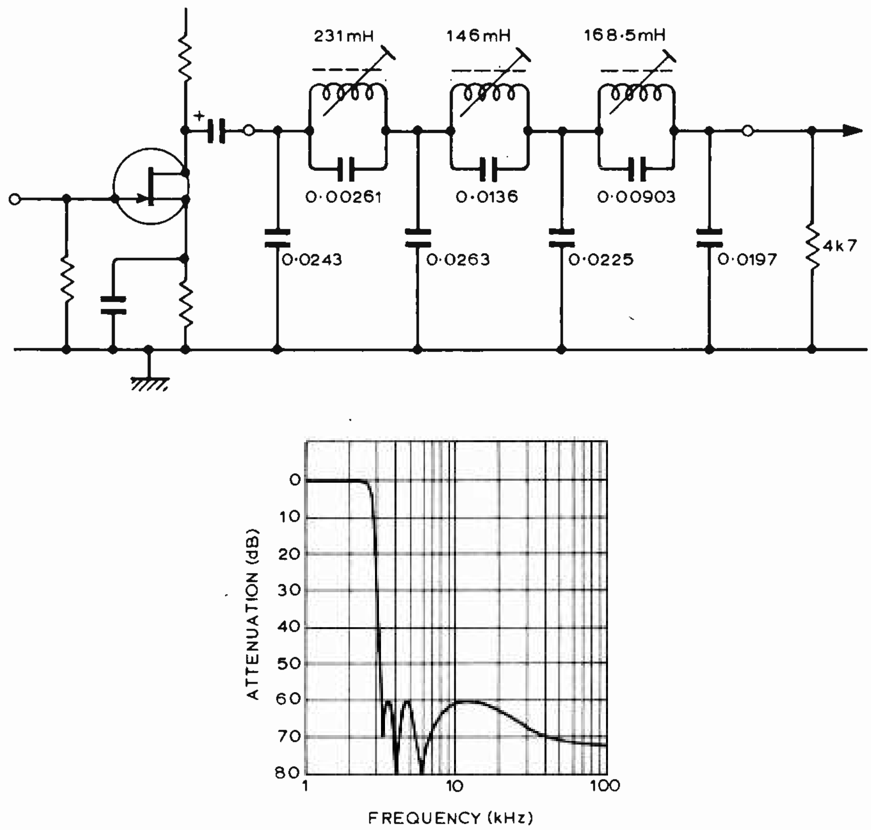

Fig. 6. Dual-gate mosfet heterodyne detector in the Ten Tec transceiver.

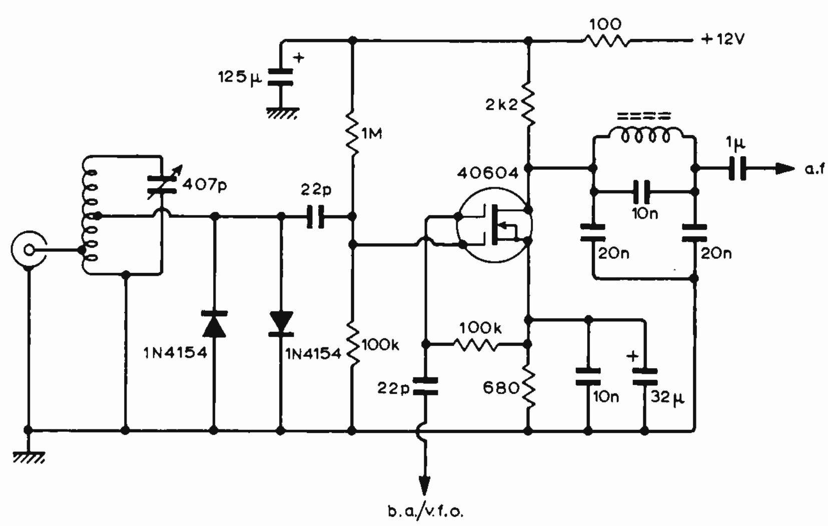

Fig. 7. High performance audio filter designed by P. G. Martin for use in direct-conversion receiver.

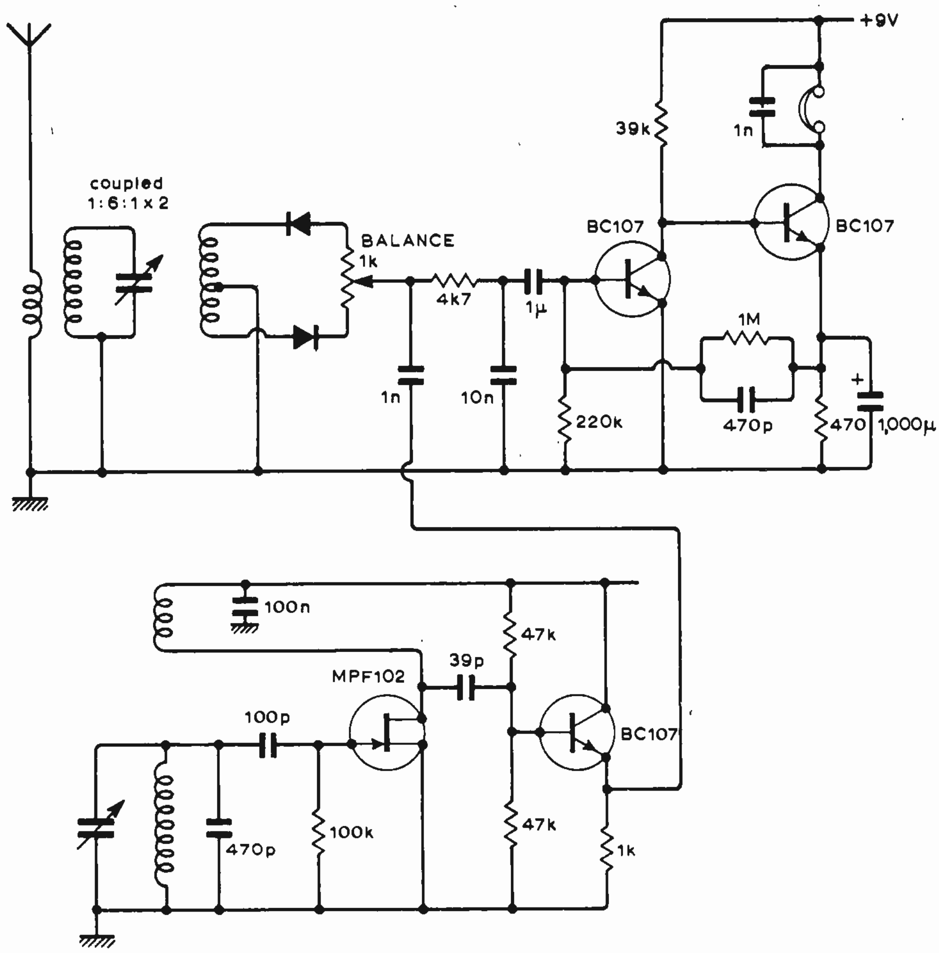

Fig. 8. How simple can you get? A direct-conversion receiver for 3.5MHz designed by K. Spaargaren.

spectrum — both claims are open to debate.

But meanwhile the simple direct-conversion receiver, which makes no attempt to achieve phase coherence, began to attract the interest of home-construction-minded amateurs concerned at the soaring prices of communications receivers suitable for s.s.b. reception. K. Spaargaren (PA0KSB) led the way by describing[12] a simple all-semiconductor receiver for 3.5MHz using five bipolar transistors and single-balanced demodulator. This design was reprinted in the U.K. and attracted considerable interest. The following year, two American amateurs Hayward and Bingham presented a design[13] using four hot-carrier diodes as a double balanced ring demodulator with f.e.t. local oscillator, Fig. 4. Meanwhile Charles Bryant (GW3SB) had pointed out[14] that, for amateur operation, the direct-conversion receiver could form a useful basis for simple transceivers because, unlike the superhet, the local oscillator was virtually at signal frequency and could be used as the transmitter v.f.o. This approach has been used by a number of amateurs for home-built portable transceivers and also forms the basis of a low-cost transceiver marketed by the American company 'Ten Tec': see Figs. 5 and 6.

Many amateurs have already found that a simple direct-conversion receiver can provide performance fully comparable to that of a medium-cost superhet, particularly where balanced heterodyne detectors are used and where the local oscillator has good stability and a low tuning rate. Selectivity of a good direct-conversion receiver is governed by the design of the post-detector low-pass filter: Fig. 7 shows an s.s.b. filter designed by P. G. Martin, G3PDM[15] with a slope factor (6 to 60dB) of 1.18, cut-off frequency 3kHz, and ultimate attenuation 75dB.

Theoretically there is no requirement for high-selectivity tuned circuits or r.f. amplification in front of the mixer provided it is of a low-noise type such as those using hot-carrier Schottky diodes. In practice it may be advisable to incorporate a reasonable degree of signal-frequency selectivity and a low-gain amplifier stage to prevent overloading the detector by strong local broadcasts or other signals and also to eliminate spurious responses that can result from harmonics of the local oscillator. Provided the detector is truly linear this is virtually the only form of spurious response, representing a marked advantage over simple superhets. Many designs, both with semiconductor devices and valves, have appeared in the past few years in the amateur press. Although such designs are usually presented as suitable only for s.s.b. and c.w. reception, some a.m. capability is usually achieved with stable oscillators, allowing the detector to work in the enhanced carrier mode.

Altogether these developments have underlined the usefulness in this specialized application of synchronous direct-conversion receivers, even when these are of extreme simplicity. Part 2 discusses how performance can be improved by the use of two-phase quadrature techniques and indicates how synchronous detection can now be extended to normal broadcast reception by phase-locked loop demodulators, and outlines the operation and advantages of bi-aural synchronous detection.

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2. O. G. Villard, "Sideband-operated a.f.c. of suppressed-carrier s.s.b. voice signals", *IEEE Trans vol. com-19*, October 1971.
3. R. P. Haviland, "Comparative Study of Communications systems using different modulation-demodulation techniques" *EBU Review* — Part A no. 115, June 1969.
4. D. G. Tucker, "The Synchrodyne", *Electronic Engineering*, March, August and September 1947.
5. Apthrope, "Correspondence concerning homodyne reception". *Electronic Engineering*, July 1947.
6. R. C. V. Macario, "An experimental diminished carrier receiver for v.h.f. radio telephony" IEE Conference Publication No. 64.
7. O. G. Villard, "Selectivity in s.s.s.c. reception", *QST*, April 1948.
8. J. R. White, "Balanced detector in a t.r.f. receiver" *QST*, May 1961.
9. "Signal Slicer", *G.E. Ham News*, July-August, 1951.
10. J. P. Costas, "Synchronous communications", *Proc I.R.E.*, December 1956.
11. J. P. Costas, "Poisson, Shannon and the Radio Amateur", *Proc I.R.E.*, December 1959.
12. K. Spaargaren, "Eenvoudige super voor 80 m EZB en CW", *Electron*, January 1967 (see also Technical Topics, *R.S.G.B. Bulletin*, March 1967).
13. Hayward and Bingham, "Direct conversion — a neglected technique" *QST*, November 1968
14. Pat Hawker, "Technical Topics", *R.S.G.B. Bulletin*, July 1967.
15. P. G. Martin, correspondence, *Radio Communication*, July 1970.

(to be concluded)

Sixty Years Ago

September 1912. In this month's issue of *The Marconigraph* Dr. W. H. Eccles contributed a short article "as a kind of supplement" to an earlier article in which Dr. J. A. Fleming had suggested that the incidence or non-incidence of sunshine on an antenna affected the intensity of the signals received by the antenna. Dr. Eccles wrote "The proposition does not appear to receive support from any known physical fact. Of course the possibility that light, especially ultra-violet light, might affect the waves *emitted from* an antenna is well known . . . but this possibility has nothing in common with the impossibility of the illumination of an antenna to affect the signals received."

Letters to the Editor

The Editor does not necessarily endorse opinions expressed by his correspondents

Sinclair pocket calculator

I was very disappointed by the mention of our calculator in the News pages of your August issue. It seems to me that this was an entirely negative piece of reporting which did nothing but criticize inaccurately what we feel to be a considerable technical achievement.

As you may be aware, a great number of companies have announced that they are making pocket electronic calculators in this country but to the best of our knowledge we are the only people actually doing so. Furthermore, our calculator has been the most enormous immediate success both in England and abroad. Far from being behind our competition we are very far ahead of it. Several very large companies abroad have asked for licences for our design and circuitry, these are the people who have seen and used our machine.

The criticisms made by your correspondent are so ephemeral and at times meaningless that they are hard to deal with individually, nevertheless I will try. The switch contacts which he describes as being rather crude are gold plated beryllium copper and if anyone knows a better way to make a switch than the one we are employing we would be delighted to hear about it. The keyboard on virtually all competing machines can, if hit fast enough, fail to make an entry; this is not the case with our machines for reasons which are clearly rather too subtle to have been obvious to your writer. I cannot understand how he found the fixed/floating point selector switch difficult to use, he is certainly the only one to complain about it. The on/off switch is definitely stiff and it is intended to be. This prevents it from being accidentally switched on when the machine is slipped into a pocket.

The Sinclair calculator, some features of which were criticized by our reporter last month.

The reference to Texas Instruments not guaranteeing the chip in our circuit is really rather irrelevant. We have never asked Texas Instruments to do so any more than we ask them to guarantee the operations of their transistors in our other circuits. As a matter of interest, however, Texas Instruments have complimented us on the ingenuity of our application and have stated that they see no objection to it whatsoever. Your correspondent also says that we ought to do a temperature cycling test on every machine. We have done a great many tests on individual samples but to do this test on every single machine would be uneconomical and totally unnecessary. We have left our machines immersed for several days in water and they work perfectly well afterwards but we do not propose to carry out this test on every machine just in case somebody drops theirs in the bath.

C. M. Sinclair,
Sinclair Radionics,
St. Ives, Hunts.

Bootstrapping

I was a little puzzled by Mr Cathles' letter (May issue p. 225) on bootstrapping. If, as he suggests, there is microphony and hash produced by the high-impedance (and therefore high gain) at the first transistor collector, then this is a noise source within the feedback loop and should be correspondingly reduced. I have not experienced this trouble, even in microphone amplifiers (e.g. Figs. 7 and 8 in "Stereo Mixer," part 1, May '71), and connecting and disconnecting the bootstrap capacitor does not cause any audible change in noise level. I have, however, found that a defective bootstrap capacitor can be the source of a large amount of low-frequency noise.

While designing the circuits for my mixer, I investigated the effect of bootstrapping as it was almost universally used in power output stages but rarely in preamplifier circuits. It is easily shown that for the circuit similar to Mr Cathles' Fig. 3 the voltage gain is approximately $A_{vo} = g_{m1} \cdot h_{fe2} \cdot R_E$, where R_E is the total a.c. load resistance at the emitter of Tr_2. The output impedance is theoretically R_E, since the emitter follower is current driven, but in practice is lower due to the finite output impedance of the first stage. The bootstrap capacitor, by simulating a constant-current load for Tr_1, ensures that signal current enters the base of Tr_2 rather than the collector load. Bootstrapping is a form of positive feedback with a loop-gain less than unity.

At the time, I made measurements on a circuit similar to Mr Cathles' Fig. 3 but with a collector load of 68kΩ + 33kΩ instead of 10kΩ + 10kΩ. Without bootstrapping, the gain (open-loop) was, as predicted, about 350 and harmonic distortion 8% at 1kHz for a 3V r.m.s. output, whereas with bootstrapping, the gain was 2500 to 3000 (slightly less than predicted) and distortion 0.8% for a 3V r.m.s. output. It is the improved linearity that is advantageous rather than the increased gain.

The necessity for the 470pF from collector to ground (reducing the open-loop high-frequency gain) in Mr Cathles' circuit suggests that high-frequency instability or a large frequency response peak was amplifying noise picked up on the long leads used, or that the leads were themselves responsible for the instability. Alternatively the emitter follower may have been unstable when driving a capacitive load. On this latter point I strongly disagree with Mr Burrows in his letter (also in the May issue) concerning the ASP circuit when he states that emitter followers do not usually oscillate on their own. In my experience it is the most unstable transistor configuration, particularly when driving a capacitive load. It can be shown by using the single-pole approximation for the current gain, that these circumstances create a negative input resistance and indeed such a circuit is used for v.h.f. oscillators. As a result I always include a damping resistor (50 — 1000Ω) in series with base, collector or load impedance or connect a 2000pF from base to emitter (to reduce h.f. response) particularly when long screened leads are involved.

H. P. Walker,
South Queensferry,
West Lothian.

Doppler effect

I write to attempt to clear a point brought to the surface by Cathode Ray both in his "Doppler Effect" article in the May issue, and again in his reply to the first letter concerning doppler effect in the August issue.

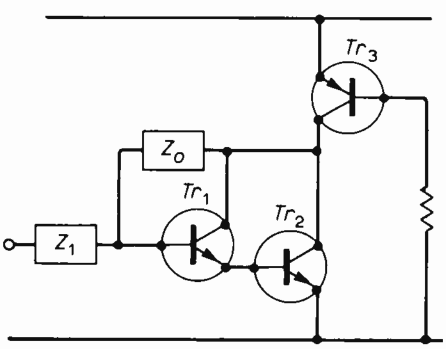
Taking the latter point first, Cathode Ray suggests a practical experiment whereby a loudspeaker mounted at one end of a tube is fed by two sine waves, one of 50Hz and one of 1000Hz, both of equal strength, so the amplitude of the 50Hz. wave will be greater than that of the 1000Hz. He argues that the 1kHz wave will be frequency modulated by the 50Hz, and suggests that this be detected by a microphone coupled to a wave analyser to show the presence of sidebands around the 1kHz "carrier". However, assuming both microphone and loudspeaker to be ideal, would not the 50Hz wave from the speaker cause the microphone diaphragm to vibrate and alter the frequency of the higher frequency signal — by Doppler Effect — in such a way as to neutralize the doppler distortion introduced by the loudspeaker? Thus the output from the microphone would consist only of frequencies of 50Hz and 1000Hz.

By a similar argument, when recordings are made for public distribution, or for any other purpose, the microphone must introduce Doppler distortion into the recording, but on replay, the speakers neutralize this by a "negative Doppler distortion". The fact that multiple diaphragm arrangements are used for hi-fi reproduction surely stems from the difficulty of manufacturing a speaker with a single diaphragm which can reproduce the entire sound spectrum within a few decibels, and not from the need to reduce Doppler distortion. In fact, with a multiple speaker system, the bass unit would cause the mass of air in front of the speaker to move, and upon this moving air mass would be superimposed the higher frequencies, so although the individual units are not creating "negative" Doppler distortion, the combined effect of all units is. Although with loudspeakers, multiple unit systems are used, very few manufacturers of headphones use multiple units, and since headphones are generally superior to loudspeakers and are preferred by some hi-fi enthusiasts, I feel this shows that Doppler distortion is not reduced by multiple diaphragm systems, if it exists at all. The one case where Doppler distortion might be said to be significant is that of electronic musical instruments where there is no microphone to impart "positive" distortion to the signal.

Finally, in his reply to Mr. P. J. Unwin's letter, Cathode Ray states: "Let us first of all dispose of the electromagnetic wave analogy . . . the waves are transverse, not longitudinal, so the Doppler question does not arise." If this is so, what was the second article by Cathode Ray about? It was entitled: "Doppler Effect in radio and other electromagnetic waves"!
John E. Foggitt,
Chesterfield, Derby.

Simple d.c. amplifier

A simple but effective improvement to the d.c. amplifier circuit on p. 239 of the May issue is to replace Tr_1, by a Darlington pair, as shown in the accompanying diagram. This has two effects: it increases



the input impedance to a high value, and increases the open-loop gain of the circuit.

With this circuit, low leakage silicon transistors should be used for best effect.
P. D. Bailey.
Markfield.
Leicester.

History repeats?

Substituting "transistor" for "valve", and "world" for "German", in the following suggests a contemporary need. Extract from *Wireless World*, August 31st 1934, Page 197, column 3.

"Valves

"Germany in the course of this year has had a so-called valve holiday, lasting from last autumn up to the present Exhibition. In this period the German valve industry agreed to sell no new types of valve, since the German public, the German trade, and also the German radio *industry, needed a rest.*"

(The italics are mine.)
R. E. Smith,
Combined Electronic Services Ltd,
Croydon.

Quadraphony news

Earlier this year we reported that RCA were tending to favour the CD-4 four-channel system (developed by the Victor Company of Japan) for their discs, rather than adopt a two-channel matrix technique. Now, RCA seem to have firmly committed themselves to this, announcing that 15 discs cut to the system will be introduced in 1972 and that all new material (no re-issues or singles) will be CD-4 by the end of 1973.

The technical achievement is remarkable and the list of techniques that make the disc possible is impressive in its length – frequency and phase modulation of the carriers, tracing-error compensation, adaptive carrier level, noise reduction operative on low-level signals in the difference channels, matrixing for compatibility, reduced-speed cutting, Shibata stylus increasing contact area, and specially developed disc material.

There are still drawbacks however, like reduced playing time – 22 min as opposed to 30 min for an ordinary l.p. disc – and disc cutter system problems, so the announcement may seem a little premature, but RCA feel confident they will be ironed out before the end of the year. The shorter playing time is necessitated because of the tracing distortion which worsens as the groove radius gets smaller. The effects are aggravated in CD-4 because of the presence of the carrier. Waveform compensation during recording helps, reducing intermodulation distortion in a carrier channel in the presence of an unmodulated carrier and 3kHz baseband signal to –27dB (–12dB without compensation). A property of the disc is the reduced level of the baseband signal, a 1kHz signal being recorded at 2.23cm/s.

A drawback until recently was playing lifetime of the discs, but as a result of incorporating a phase-locked loop to produce a "more sensitive" demodulator and using a harder record compound (which includes a resin mix containing anti-static, lubricating stabilizer and other lubricants) it is claimed that discs can be played 100 times in an inexpensive stereo player with conical stylus and 5g tracking weight while maintaining satisfactory separation and signal-to-noise ratio (no figures available yet).

In the CD-4 system the left front and left back signals are summed and used as baseband modulation for the inner groove wall — similarly for the right channels. To enable the two baseband signals on each wall to be separated, an angle-modulated carrier is modulated with the difference between the two. (Actually the 30kHz carrier is frequency modulated up to a turnover frequency of 800Hz, pre-emphasized from 800Hz to 6kHz – i.e. phase modulated – and frequency modulated over 6kHz.) Thus the system is stereo and mono compatible.

Following hard on the heels of the RCA announcement in London came another CBS demonstration of the SQ system. At this demonstration – the most effective we have heard – Benjamin Bauer from CBS Labs and Joseph Dash from CBS Records told *Wireless World* that there were now 33 hardware licensees of the SQ system, including such well-known American names as AR, Fisher, Harman-Kardon, KLH, Marantz, Sherwood, H. H. Scott, Electro-Voice, Lafayette, Radio Shack, Pilot, EICO, Metrotec/BSR, Morse and so on. In Japan Trio/Kenwood, Pioneer, Aiwa and Sanyo have joined Sony in incorporating SQ decoders in their equipment. First European licensee is Connaught Equipment who are selling modules to U.K. and Continental equipment manufacturers.

The effectiveness of the demonstration in terms of entertainment value – from our seat anyway – may well be to the credit of the full "logic" system fitted to the Sony SD2000 decoder (not available in the U.K.), although we have not done experiments to determine when this "logic" circuitry becomes defeated, and many find its action unacceptable. (The long-awaited Sony decoder/amplifier type SQA200 fitted with front-back "logic" will be available in the U.K. from September priced at £57.75.)

A joint licencing agreement between Electro-Voice and CBS Records was announced a little earlier. This gives CBS and licensees rights to the basic Scheiber/E-V patent covering matrix techniques. E-V are promoting their decoder chip which can be used to decode SQ as well as other matrices and CBS now have a matrix chip – with two chips for "logic" use expected later this year.

Meanwhile in Japan, it is reported that agreement has been reached to "standardize" on three systems, the "regular matrix" – a Sansui-type matrix – SQ matrix, and the CD-4 system.

Corrections

Hand-portable Transceiver. Inductors 1 to 9 for D. A. Tong's design in the April issue should be close wound with 22 s.w.g. wire. Inductors 16, 18, 28 and 22-26 in the transmitter section may be wound with 18 or 20 s.w.g. wire, spacing not critical. (Inductors 10-15, 17, 19-21, 24, 27 and 31-33 are all r.f. chokes). A capacitor of 10μF, 10V working should be connected to Tr_{38} emitter (Fig. 6) and earth. Resistors 91 and 95 (Figs. 7 & 8) are 10-kΩ linear-law types; 96 should be 8.2kΩ and not 82kΩ. Diode types such as the 1N914 and 1N4148 may be used in place of the FD101 for D_{23} and D_{24}. In Fig. 2 we regret that capacitors 12, 14 and 16 were shown connected incorrectly. They should all be in parallel across L_8 primary.

Current-limited Power Supply. In Fig. 7 of A. Royston's article, February issue, the centre-tap of the transformer should be wired to connection 9 and not 5. The lead from the junction of R_{13} and C_6 should be taken to connection 5.

Experiments with operational amplifiers

5. Using an op-amp as a differentiator

by G. B. Clayton,* B.Sc., F.Inst.P.

An operational amplifier with negative feedback applied via a resistor connected between output terminal and phase inverting input terminal performs the operation of differentiation on a signal applied to the phase inverting terminal through a capacitor.

A simple differentiator circuit is illustrated in Fig. 5.1. The input current through the capacitor is proportional to the rate of change of the input voltage. The output voltage of the amplifier causes the input current to flow through the feedback resistor and the output voltage thus takes on a value proportional to the rate of change of the input voltage. If the performance of the amplifier is assumed to be ideal the response of the circuit is described by the equation

$$e_o = -CR\frac{de_i}{dt} \qquad (5.1)$$

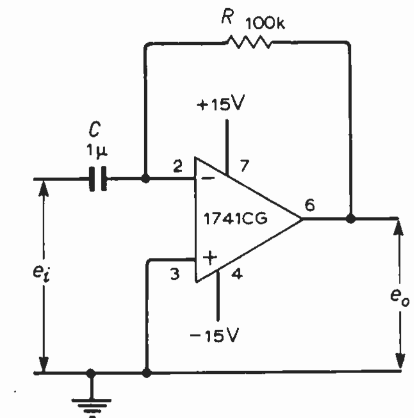


Fig. 5.1. Simple differentiator made with an op-amp.

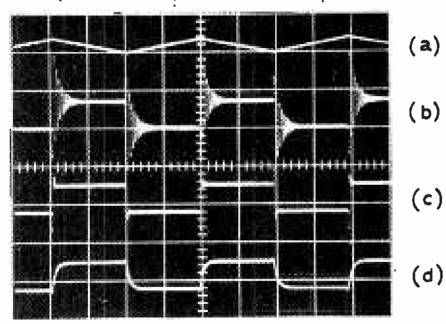


Fig. 5.2. Showing frequency compensation effects: (a) input signal, 1V/div.; (b) output without frequency compensation, 5V/div.; (c) output with addition of R_i in Fig. 5.3, 5V/div.; (d) output with addition of R_i and C_f in Fig. 5.3, 5V/div. Horizontal scale, 10ms/div.

Frequency compensation

A low-frequency linear sawtooth is a convenient input test signal for examining the action of the simple differentiator. Typical input (a) and output waveforms are shown in Fig. 5.2. In trace (b) the output voltage is seen to overshoot and ring in response to a sudden change in the slope of the input wave. This is because signal frequencies outside the 3dB bandwidth limit of the amplifier undergo a phase change approaching 180° when fed back to the input. The input capacitor causes a phase lag of 90° and at frequencies outside its 3dB bandwidth limit the amplifier contributes an additional phase lag approaching 90°.

Overshoot and ringing can be prevented by adding either a resistor R_i in series with the capacitor C or a capacitor C_f in parallel with resistor R, or both R_i and C_f may be added to the circuit. The circuit in Fig. 5.3 shows the simple differentiator with the addition of frequency compensating components, and traces (c) and (d) of Fig. 5.2 show their effects on the output waveform.

The response equation, (5.1), for the frequency compensated differentiator may be verified by varying the amplitude and frequency of the input sawtooth voltage and measuring the output voltage for each input waveform.

It is instructive to observe the response of the differentiator to a variety of input wave-

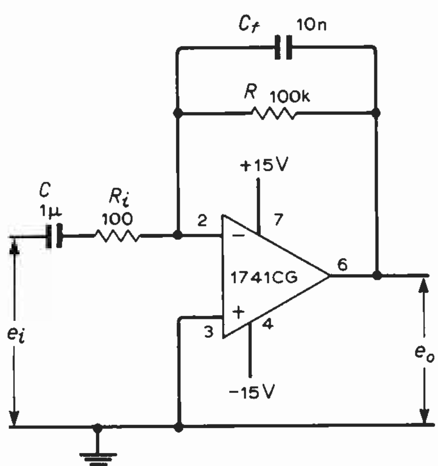


Fig. 5.3. Differentiator with frequency compensation components added.

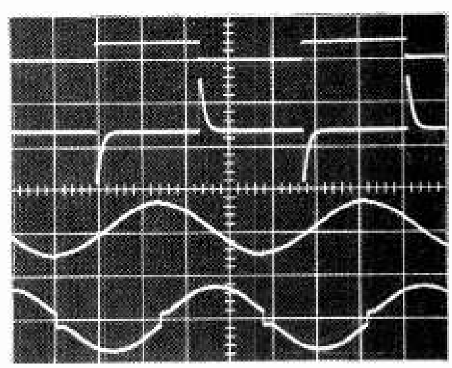


Fig. 5.4. Response of differentiator to square wave input (top trace) and to sine wave input (third trace down).

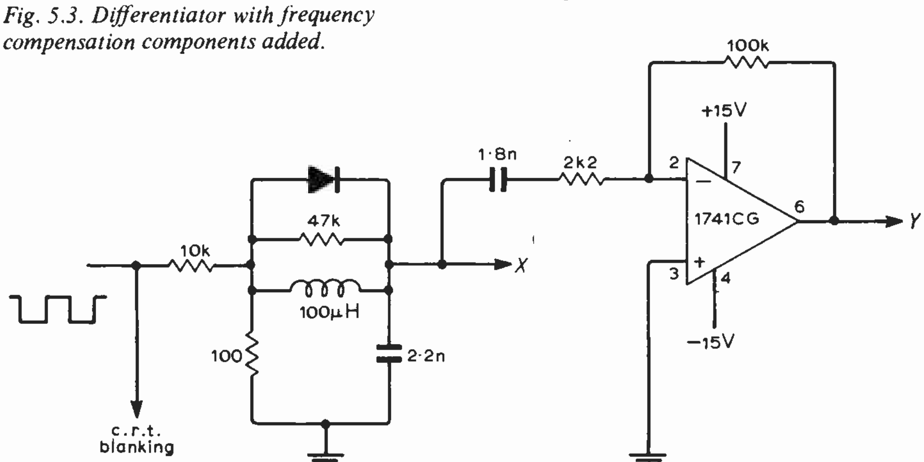


Fig. 5.5. System for obtaining a plot of de/dt *against* e *for a transient.*

*Department of Physics, Liverpool Polytechnic.

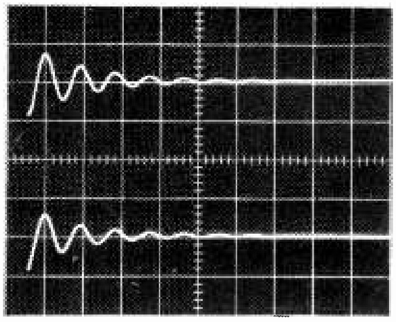


Fig. 5.6. Transient response of an LCR circuit with (a) resistive damping, (b) diode damping.

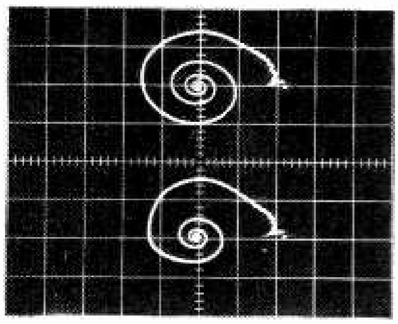


Fig. 5.7. Plots of de/dt *against* e *for an LCR circuit with (a) resistive damping, (b) diode damping.*

forms. The traces in Fig. 5.4 show input and output waveforms for a square wave input (top) and for a sinusoidal input. In fact the input sinusoid (third from top) has a slight discontinuity at its peak values; the presence of this discontinuity is far more apparent in the differentiated waveform (bottom).

Application of a differentiator

The transient response of a control or feedback system is often observed in order to investigate the stability of the system. In such observations it is difficult to distinguish effects due to non-linear elements. If the transient signal is plotted against the differentiated transient, the non-linear effects are more readily observed.

A circuit to illustrate the principle of this type of application is shown in Fig. 5.5. The effects of resistive (a) and diode (b) clamping on an *LCR* circuit is first compared in a transient response display (Fig. 5.6). Values are chosen so that the diode does not conduct heavily and it is difficult to distinguish the effect of diode non-linearity. The transient is now used to produce the horizontal deflection and the differentiated transient the vertical deflection of an oscilloscope. The traces obtained are shown in Fig. 5.7, and the effect of diode non-linearity in (b) is clearly apparent. The input square wave used to excite the *LCR* circuit is applied to the c.r.t. cathode so as to blank off the display of the negative transient.

(Next month: op-amps with defined non-linear response.)

Conferences and Exhibitions

Further details are obtainable from the addresses in parentheses

LONDON

Sept. 4-8 Grosvenor House
International Broadcasting Convention
(I.B.C., c/o I.E.E., Savoy Pl., London WC2R OBL)

Sept. 11-15 Savoy Pl.
Gas Discharges Conference
(I.E.E., Savoy Pl., London WC2R OBL)

Sept. 13-15 Polytechnic of Central London
Instruments — Measurements, Specifications and Certification
(Lisa Spaducci, Polytechnic of Central London, 115 New Cavendish St., London W1M 8JS)

Sept. 18-22 U.S. Trade Center
Exhibition of Materials for the Electronics Industry
(U.S. Trade Center, 57 St. James's St, London SW1)

BRIGHTON

Sept. 18-20 The University
Point Defects and their Aggregates in Metals
(Inst. Physics, 47 Belgrave Sq., London SW1X 8QX)

Sept. 26-28 Metropole Convention Centre
Power Sources Symposium
(Int. Power Sources Symp., P.O. Box 136, 26 Wellesley Rd, Croydon CR9 2EG, Surrey)

EDINBURGH

Sept. 6-8 Heriot-Watt University
Tunable Lasers Conference
(Inst. Phys., 47 Belgrave Sq., London SW1X 8QX)

Sept. 26-28 The University
Electronic Equipment Design and Manufacture
(G. K. P. Ayrton, Post Office Telecommunications Headquarters, 19 Canning Street, Edinburgh EH3 8TH)

FARNBOROUGH, Hants.

Sept. 4-10 R.A.E.
Electronics in Aviation Exhibition
(S.B.A.C., 29 King St, St. James', London SW1)

HARROGATE

Sept. 1-3 Majestic Hotel
Audio '72
(Exhibition & Conference Services, Claremont House, Victoria Ave., Harrogate)

KEELE

Sept. 20-22 The University
Automation of Testing Conference and Exhibition
(I.E.E., Savoy Pl., London WC2R OBL)

LANCASTER

Sept. 12-15 The University
ESSDERC — European Solid State Device Research Conference
(Inst. Phys., 47 Belgrave Sq., London SW1X 8QX)

LEEDS

Sept. 15-17 The University
Social Responsibility and Education in Physics Conference
(Inst. Phys., 47 Belgrave Sq., London SW1X 8QX)

MANCHESTER

Sept. 6-12 U.M.I.S.T.
Electron Microscopy Congress
(Inst. Phys., 47 Belgrave Sq., London SW1X 8QX)

OXFORD

Sept. 25-28 New College
Quality Assurance Conference
(Inst. Eng. Inspection, 146 Cromwell Rd, London SW7 4EF)

SOUTHAMPTON

Sept. 12-14 The University
Elementary Particle Physics Conference
(Inst. Phys., 47 Belgrave Sq., London SW1X 8QX)

Sept. 25 & 26 The University
Electro-optic Systems in Flow Measurement Conference
(Inst. Phys., 47 Belgrave Sq., London SW1X 8QX)

UXBRIDGE

Sept. 4-7 Brunel University
Online 72 — Computing Conference
(Brunel University, Uxbridge, Middlesex)

OVERSEAS

Sept. 3-10 Leipzig
Autumn Fair
(Leipziger Messeamt, DDR-701 Leipzig, Postfach 720)

Sept. 12-14 San Francisco
Compcon 72
(I.E.E.E. Computer Society, 8949 Reseda Boulevard, Suite 202, Northridge, California 91324)

Sept. 13-15 Newport, Rhode Island
Engineering in the Ocean Environment
(Ocean 72, The Newport Harbor Treadway Inn, Newport, Rhode Island 02840)

Sept. 13-15 Geneva
Electro-Optics Exhibition and Conference
(Mack Brooks Exhibitions Ltd., 7 London Rd, St. Albans, Herts)

Sept. 20 & 21 New York City
Optical Instrumentation in Security, Surveillance and Law Enforcement
(Society of Photo-Optical Instrumentation Engineers, P.O. Box 288, Redondo Beach, California 90277)

Sept. 21-27 Tokyo
Japan Electronics Show
(Electronic Industries Association of Japan, 14 Marunouchi 3-chome, Chiyoda-ku, Tokyo)

Sept. 25-29 Amsterdam
Fiarex 72 — Electronics Trade Exhibition
(R.A.I., Europaplein 8, Amsterdam)

Sept. 27-Oct. 1 Los Angeles
Expo Electronex
(Expo Electronex, 3600 Wilshire Blvd, Los Angeles, Ca. 90010)

Sept. 30-Oct. 6 Copenhagen
Elektronik 1972
(Erhvervenes Udstillingsselskab Bella-Centret A/S, Hvidkildevej 64, 2400 Kobenhavn NV)

Meetings

CARDIFF
26th UWIST — ICL lectures in technical communication: "Lively messages for alert management" by Stafford Beer; "Technology, communication and social life" by Colin Cherry; and "Are there rules for writing English?" by Peter Wason at 11.30 at University of Wales Institute of Science and Technology.

COLCHESTER
12th IERE — "High power ultrasonic transducers" by A. E. Crawford at 18.30 at The University of Essex, Wivenhoe Park.

EDINBURGH
22nd Inst.P. — One-day meeting "Non-linear ultrasonic effects in solids" at 10.00 at The Engineering Department, University of Edinburgh.

GLOUCESTER
20th IERE — "Advanced aircraft display systems" at 19.30 at the Technical College.

SOUTHAMPTON
28th Inst.P. — "Electron beams" symposium at 10.15 at the University.

STANSTED
26th IERE — "The civil aviation flying unit and its functions" by J. H. Gidman at 18.30 at Stansted Airport. (Advance registration necessary: I. Jefferies, 3 Tees Road, Springfield, Chelmsford.)

Graphical Analysis of Pulses on Lines

by B. L. Hart, B.Sc., M.I.E.R.E.*

When transistor-transistor logic elements are connected together to form a digital system, spurious pulses can be introduced as a result of reflection effects on the interconnecting wires. The standard reflection chart approach is not appropriate to the analysis of waveshapes on transmission lines with non-linear terminations. The shape of the reflected pulses and, hence, the trouble they are likely to cause can be calculated by the graphical technique described using a new idea of a 'sliding load-line'.

The conventional approach to the treatment of pulse reflections on uniform lossless transmission lines is based on the assumption of linear source and load impedance. The concept of reflection coefficient and the use of a suitable 'reflection chart' reduces many line problems to simple arithmetic calculations. In the practical construction of high-speed pulse generators capable of delivering waveforms with nanosecond and sub-nanosecond edges, precision coaxial cable (e.g. RG213U) and coaxial terminations (e.g. BNC, GR-874, etc.) are employed and the agreement between waveforms observed experimentally and predicted theoretically with the aid of a reflection chart is precise to a degree which is acceptable in the majority of applications likely to be encountered in practice. There are, however, some instances—specifically, those associated with the use of transmission lines with transistor-transistor logic—where the source and load impedances are distinctly non-linear. In these cases a reflection chart approach would be of limited use. Furthermore, a purely mathematical solution would present formidable difficulties.

However, a graphical approach, which avoids the computational problems, yields rapid results and enables us to see simply the effect of changes in system parameters. It is the background to the graphical technique which is explored in some detail in this article. First, though, it is necessary to review our basic ideas concerning pulses on lines.

Line pulse fundamentals

Consider the simple arrangement shown in Fig. 1, in which a uniform, lossless, transmission line having characteristic impedance Z_o and terminated by an impedance Z_T is connected, at time $t = 0$, by a perfect switch to a battery in series with an impedance Z_G. (Lest it be thought that we are

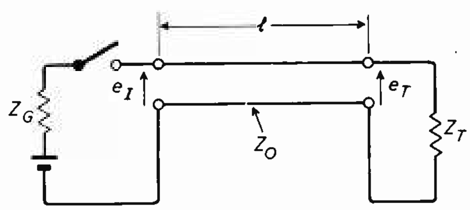


Fig. 1. Transmission line with voltage step function drive

already completely divorced from reality it must be borne in mind that a good practical approximation to a perfect switch is a specially-constructed mercury-wetted relay in a coaxial housing: the rise time of pulse edges using such a component is typically <2ns.)

A straightforward, though tedious, mathematical approach (see for example ref. 1) to the problem of finding the time function of current and voltage at any point, X, on the line is to solve the differential equations describing line behaviour in terms of its intrinsic properties for specified boundary conditions, viz. nature of driving function, particular values of Z_G,Z_T.

A practical engineering approach is different. The starting point is the appreciation that the appropriate differential equations are linear; thus, any two possible solutions can be a combined and the result is also an equally-valid solution. A physical interpretation of this is the possibility for the simultaneous existence of two waves travelling in opposite directions along the line. The voltage (or current) at a given x,t is given by the algebraic sum of these two waves, the amplitudes of which are calculable for specified Z_G,Z_T. For each wave the voltage is related to the current by the impedance Z_o.

Thus, if in Fig. 2(a) the voltage and current components for a forward wave, i.e. travelling to the right in the direction of increasing x, are e_f, i_f respectively, then

$$e_f/i_f = Z_o$$

Similarly, for a reverse wave (i.e. one travelling from right to left in the direction of decreasing x) characterized by voltage e_r and current i_r (Fig. 2(b))

$$e_r/i_r = Z_o$$

Now for any further quantitative treatment we must decide on, and adhere to, a consistent sign convention, and a long established one is this. A positive line voltage is one for which the 'top' conductor in Fig. 2 is positive with respect to the 'bottom', and a positive current is one flowing from left to right in the top conductor (and hence right to left in the bottom). For top and bottom read inner and outer respectively for a coaxial line system; a parallel line system is used here for each of drawing. On this basis e_f, i_f are both positive. Write $e_f = e_+$, $i_f = i_+$ where '+' denotes the forward wave. Then the expression for the forward wave becomes

$$(e_+/i_+) = Z_o$$

In describing the reverse wave by a '−' subscript we have $e_r = e_-$ but now $i_- = -i_r$. Thus the expression for the reverse wave becomes

$$(e_-/i_-) = -Z_o$$

The emphasis on signs at this stage pays dividends later. The reason for the minus sign in this equation is not always well explained in textbooks and it is essential that it be clearly understood for future discussion.

Let us return to our specific problem:

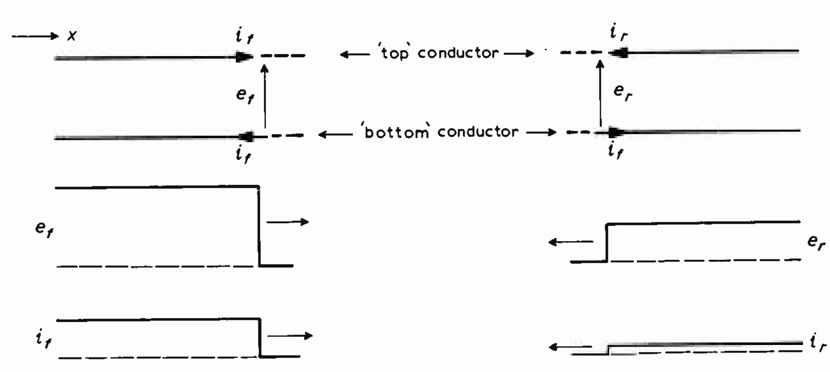


Fig. 2. Sign conventions for (a) forward wave, and (b) reverse wave.

*North-East London Polytechnic.

when the switch in Fig. 1 is closed at $t = 0$ the line appears as an impedance Z_o irrespective of the nature of Z_T. (It is for this reason that Z_o is sometimes referred to as 'surge impedance'.) For the low-loss cable assumed, this means a pure resistance, typically 50Ω. A forward wave thus starts down the line. No reverse wave is yet possible: the forward wavefront has got to discover first what exists at the other end of the line. Fig. 3 thus shows the simple equivalent circuit for calculating e_+. Clearly,

$$e_+ = e_I(0+) = Z_oE/(Z_o+Z_G)$$

and this voltage waveform travels towards Z_T at a velocity v dependent on the properties of the line and typically 0.3m/ns. As the wave progresses the electric and magnetic fields are in step and the energy is shared equally between the magnetic and electric fields. The current i_f at x charges up the line between x and $(x+\delta x)$ to e_+ in a time $\delta t = (\delta x/v)$.

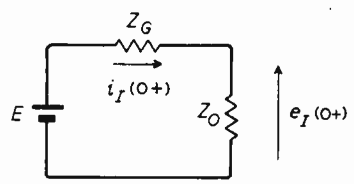


Fig. 3. Equivalent circuit at line input at $t = 0+$.

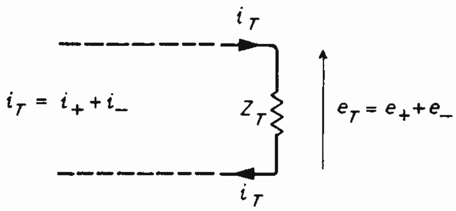


Fig. 4. Circuit conditions at termination at $t = t_d$.

At the termination all the incident energy in the forward wave is absorbed if the relationship that the line establishes between e_+ and i_+ is the same as the termination demands between the voltage across it and current in it at $t = t_d$, i.e. $e_T(t_d), i_T(t_d)$ respectively. For this condition

$$\frac{e_+}{i_+} = Z_o = \frac{e_T}{i_T} = Z_T$$

If $Z_T \neq Z_o$ all the energy in the forward wave cannot be absorbed at the termination. The law of conservation of energy thus demands a reflection phenomenon which is characteristic of all physical systems in which all the incident energy cannot be accepted. Thus light is reflected from the boundary between two media with different optical properties. Actually the optical analogy is a good one and it is profitable to regard the termination as a partially silvered mirror—but this goes beyond the scope of the present treatment.

The reflection of energy causes a reverse wave, characterized by e_- and i_- which must be such that, at the termination (see Fig. 4)

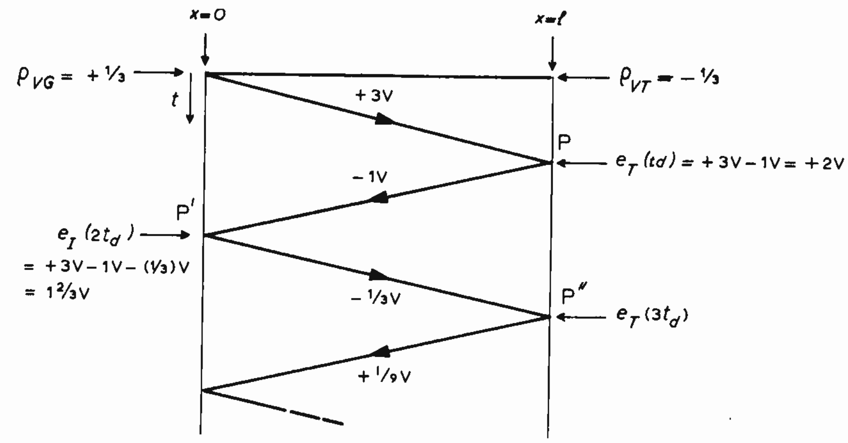


Fig. 5. Reflection chart solution to problem of a mismatched line having linear source and terminating resistances.

$$e_T = e_+ + e_-$$
$$i_T = i_+ + i_-$$

The plus sign is dictated by our sign convention. From these

$$\frac{e_T}{i_T} = \frac{(e_+ + e_-)}{(i_+ + i_-)}$$

$$\frac{e_T}{i_T} = \frac{(1+\rho_{VT})}{(1-\rho_{VT})} Z_o$$

where, by definition, $\rho_{VT} = (e_-/e_+)$ = voltage reflection coefficient at the termination. Rearranging this last equation gives

$$\rho_{VT} = \frac{(Z_T-Z_o)}{(Z_T+Z_o)} \quad (1)$$

If, as is often the case, $Z_T \neq f(i_T)$, then ρ_{VT} = constant. As it travels along the line e_- causes the line to assume a voltage corresponding to that at the termination. When e_- reaches the battery or source end there is complete absorption of energy in the e_- wave if $Z_G = Z_o$. For $Z_G \neq Z_o$ a reflection occurs characterized by a voltage reflection coefficient, ρ_{VG}, where by analogy with equation 1

$$\rho_{VG} = \frac{(Z_G-Z_o)}{(Z_G+Z_o)} \quad (2)$$

Again, if and only if $Z_G \neq f(i)$, then ρ_{VG} = constant.

To illustrate the simplicity of a typical practical calculation assume $Z_G = R_G = 100\Omega$; $Z_o = R_o = 50\Omega$; $Z_T = R_T = 25\Omega$; $E = +9V$. From equations 1 and 2 $\rho_{VG} = (+1/3)$; $\rho_{VT} = (-1/3)$. Furthermore, $e_I(0+) = e_+ = \{(9\times150)/50\}V = 3V$.

The line behaviour can be represented by a reflection chart (Fig. 5) which is a convenient two-dimensional plot aiding and summarizing our calculations. By convention, distance along the line increases horizontally to the right on this chart and time vertically downwards. From the chart we can find the line conditions at an instant 'frozen' in time or the time function of voltage (or current) at any specified point. The motion down the line of the initial +3V step is indicated by the line OP: the magnitude of the slope of OP is the wave velocity, v. The progress of the −1V reflected wave is shown by the line PP′ sloping equally in the opposite direction.

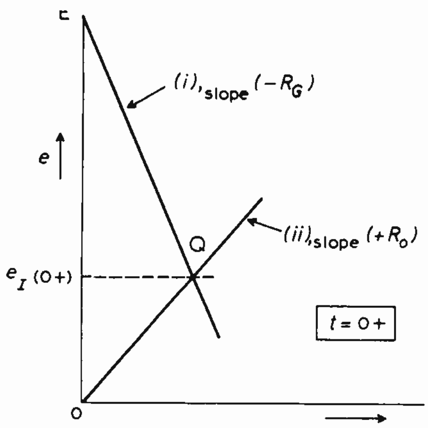


Fig. 6. Graphical representation of Fig. 3. (Abscissa should be labelled i.)

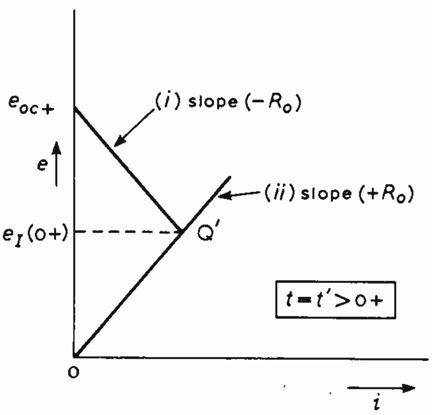


Fig. 7. Construction giving line behaviour at point X reached by forward wavefront at t' where $0+ < t' < t_d$.

No fundamental importance attaches to the signs of the slopes: they have the polarities shown through the arbitrary (but convenient) choice of t increasing vertically downwards. The manner of finding $e_I(t)$, $e_T(t)$ is clear from Fig. 5: we just sum the existing voltage, the incident wave voltage and the reflected wave voltage. If all Z_G,Z_T were linear then what we have discussed is, arguably, all that one need to know in the majority of practical circuits in which line pulse reflections occurred.

The fundamental implication of non-linearity in source and termination impedances is that ρ_{VG}, ρ_{VT} are $f(i)$ and are thus not constant: thus a simple reflection chart approach—with all its appealing elegance—is not possible. How then to deal with a practical non-linear termination such as a semiconductor diode? We will see that a graphical approach is appropriate. To develop this approach and in so doing to smooth the transition from familiar to less familiar concepts we will re-solve graphically the problem just considered. This will enable us to compare and check our analysis as we proceed.

Graphical approach: the 'sliding Thévenin source' concept

For purely resistive impedances, i.e. $Z_G = R_G$, $Z_o = R_o$, $Z_T = R_T$, Fig. 6—a graphical interpretation of Fig. 3—is a 'snapshot' of the initial conditions at the input to the line. (We have chosen to plot $e = f(i)$ but the alternative choice $i = f(e)$ is equally valid.) In Fig. 6 curves (*i*) and (*ii*) are both straight lines: (*i*) represents the characteristic of the battery E and source resistance R_G while (*ii*) gives the instantaneous relationship between $e_I(t)$ and $i_I(t)$ at $t = 0$; (*ii*) is, of course, a straight line of slope $+R_o$ passing through the origin.

The intersection point Q, gives $e_I(0+) = ER_o/(R_o+R_G)$. Suppose we now wish to represent line behaviour at a point X, situated at a distance x along the line, at a time $t'(=x/v)$ by an instantaneous time plot similar to Fig. 6. Such a plot is shown, correspondingly labelled, in Fig. 7. The dynamic resistance seen looking in either direction from the selected point is the characteristic resistance R_o. Thus curve (*ii*) represents, as before, the line to the right of X while (*i*) with a slope $-R_o$ (rather than $-R_G$) now represents the resistance looking backwards towards the battery. The point of intersection, Q', of (*i*) and (*ii*) is the line voltage: but this is equal to $e_I(0+)$ because in travelling down the line the wave e_+ causes successive points to assume this value. It is thus evident that the apparent battery voltage seen from X is $e_{oc+} = 2e_I(0+)$.

Clearly, as far as points to the right of X are concerned we can imagine the line to be cut at X and the section to the left to be replaced by a Thévenin source circuit which 'slides' along the line. This process is illustrated in Fig. 8: the equivalent source components in the chain-dotted rectangle move to the right at the velocity v of the e_+ wavefront. If we 'stack-up', one behind the other, a series of plots such as Figs. 6 and 7, we obtain the three-dimensional e,i,t plot in Fig. 9, and this provides a good visualization of wavefront progress. Thus the projection on a plane drawn perpendicular to the i axis and passing through Q' gives $e = f(t)$, while the projection on a plane drawn perpendicular to the e axis yields $i = f(t)$.

The conditions obtaining at the termination at $t = t_d$ are found by considering the e,i curve there—see Fig. 10(a): $e_T(t_d)$ can be calculated from this or the associated circuit diagram in Fig. 10(b). Obviously,

$$e_T(t_d) = e_{oc+}R_T/(R_T+R_o)$$

$$\text{or, } e_T(t_d) = 2e_I(0+)R_T/(R_T+R_o)$$

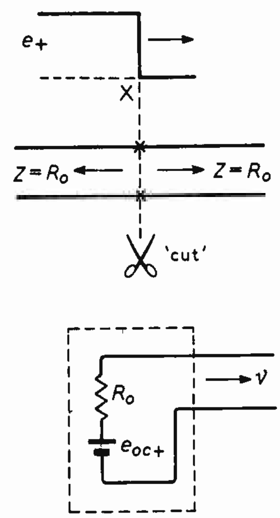


Fig. 8. Illustrating the 'sliding Thévenin source' concept for forward wave.

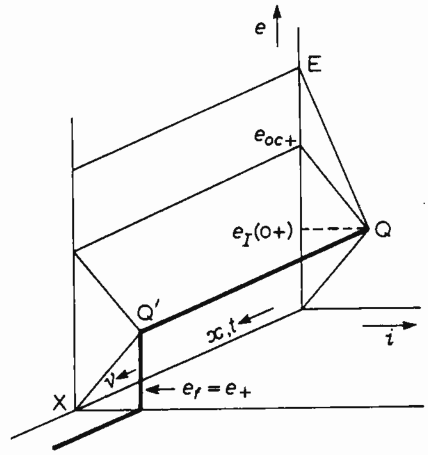


Fig. 9. An e,i,t plot of line behaviour for $0+ < t < t_d$.

Using the numerical data for our example,

$$e_I(0+) = \{(9\times 50/150)\}\text{V} = 3\text{V}$$

$$e_T(t_d) = \{(2\times 3\times 25)/75\}\text{V} = 2\text{V}$$

The value of $e_T(t_d)$ agrees, of course, with that indicated in Fig. 5. Now $R_T \neq R_o$ so there is a reflected wave e_- which proceeds from the termination to the battery and causes successive points on the line to assume the voltage $e_T(t_d)$ as it reaches them. The line voltage e at any point due to e_- is given by

$$e = iR_o + e_{oc-}$$

This equation is represented in Fig. 11 by a line of slope $+R_o$ passing through a point, Q'', having co-ordinates $e_T(t_d), e_T(t_d)/R_T$. The plus sign for R_o arises through our choice of sign for i, discussed earlier: we wish to draw the characteristic of the reflected wave on the same diagram as we have used for the forward wave in which the positive direction of current flow is taken to the right in the top conductor. To the left of the e_- wavefront the line appears as a battery of magnitude e_{oc+} in series with a resistance R_o. Fig. 11 should be compared with Fig. 7: Fig. 12 shows a sliding source equivalent representation of e_- and should be compared with Fig. 8. We can, if we wish, draw for e_- a diagram similar to Fig. 9,

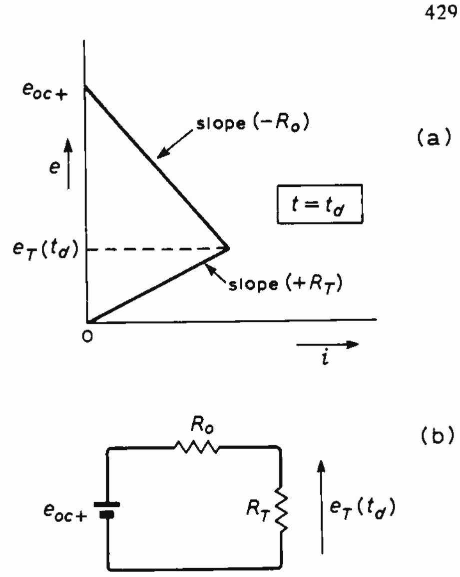


Fig. 10. Line conditions at termination at $t = t_d$ *(a) graphical construction and (b) equivalent circuit.*

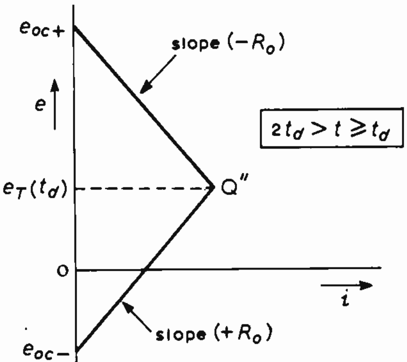


Fig. 11. Counterpart of Fig. 7 for reverse wave for $t_d < t < 2t_d$.

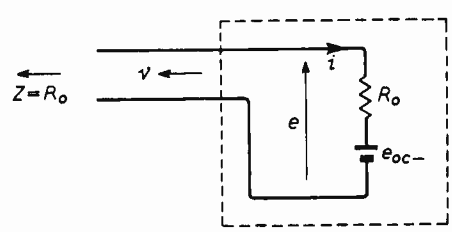


Fig. 12. 'Sliding Thévenin source' representation of reverse wave (c.f. Fig. 8).

but this has not been done as no new principle is involved.

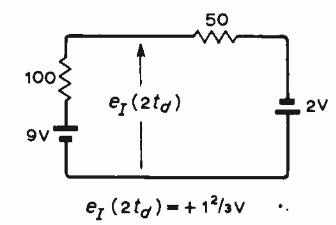
At the battery, Figs 13(a), 13(b) apply when the reflected wave arrives. Thus $e_I(2t_d) = \{9-(11\times 100/150)\} = 1.67\text{V}$, as predicted by the chart of Fig. 5. The graphical construction can now be repeated for further forward and reverse waves. The intersections of successive lines of slope $+R_o$ with the source characteristic (in this case a battery) yield, respectively, $e_I(0+\leqslant t<2t_d)$, $e_I(2t_d\leqslant t<4t_d)$, etc., while the intersections of the lines of slope $-R_o$ with the termination characteristic give, respectively, $e_T(t_d\leqslant t<3t_d)$, $e_T(3t_d\leqslant t<5t_d)$, etc. Theoretically, reflections occur until $t=\infty$ at which time $e_I(\infty) = e_T(\infty) = 1.8\text{V}$; this represents the intersection of the d.c.

output characteristic of the resistance at the end of the, supposedly lossless, cable.

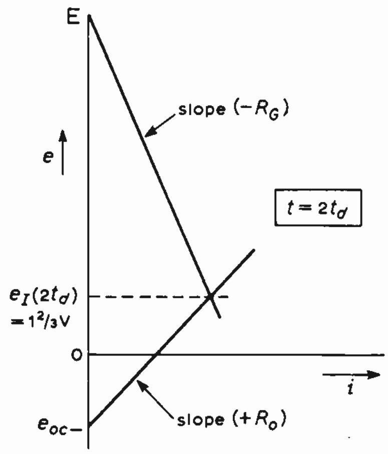
The lines with slope $\pm R_o$, used to describe the forward and reverse waves, are sometimes called Bergeron lines in honour of the Frenchman who appears to have been the first to use this graphical technique for describing the motion of waves. To describe the terminal behaviour of the line up till, say, $t = 2t_d$ we can superpose the individual time pictures to get on a single e,i graph the composite time picture shown in Fig. 14(a). The terminal voltage waveform plots are readily obtained as in Fig. 14(b).

Though it offers some physical insight, the graphical technique would not normally be used for linear source and termination resistances since the reflection chart method is so quick and easy to apply. However, graphical analysis comes into its own with non-linear source and/or load resistances, as will be seen in the sections which follow.

Fig. 13. (a) Diagram for finding $e_I(2t_d)$. (b) Graphical equivalent of (a).



(a)

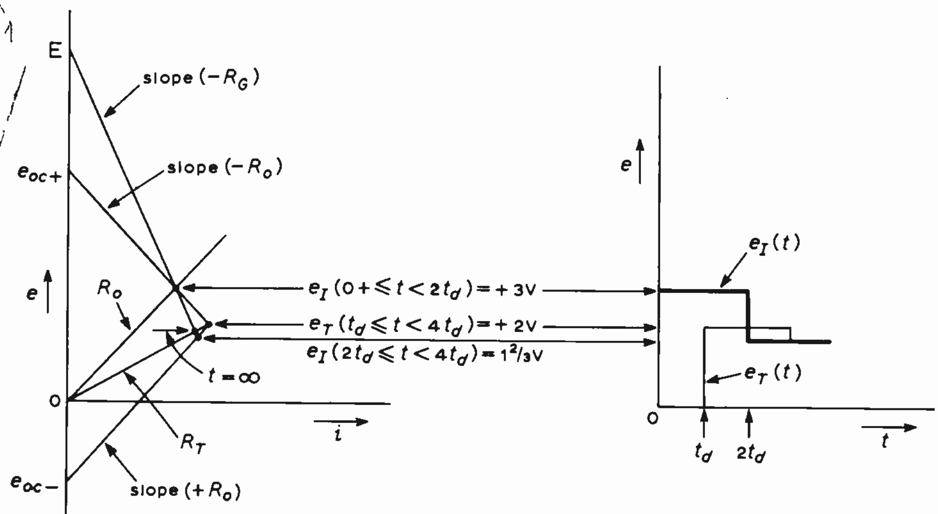


(b)

Graphical approach: non-linear termination

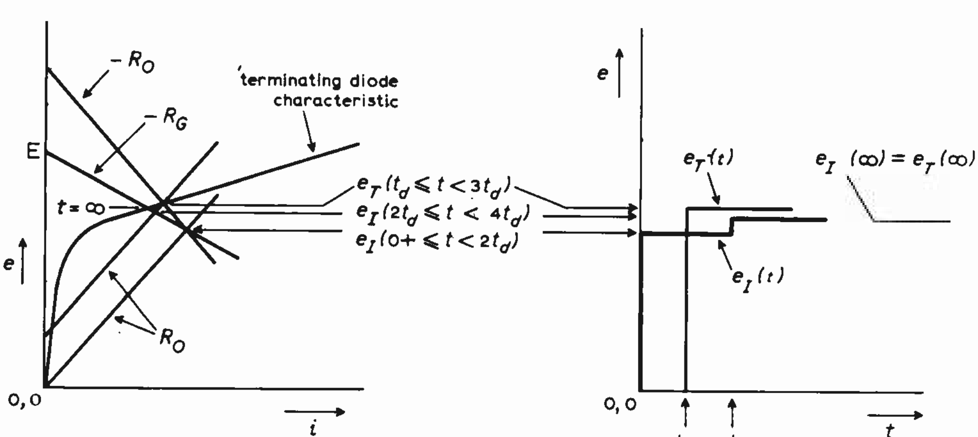
Suppose we replace the impedance Z_T in Fig. 1 by a semiconductor diode with its anode connected to the top conductor and its cathode to the bottom. None of the theory behind the graphical construction so far discussed is altered—we do not change the mechanism of wave propagation on the line—only the boundary conditions are different. Thus, Fig. 15(a) shows a composite picture of terminal behaviour for $0+ \leqslant t < 2t_d$, for an arbitrary diode. Reflections occur and continue till the point $e_T(\infty)$ is reached. The photographs in Fig. 16 show the reflection behaviour in a practical case. Fig. 16(a) shows the open-circuit output voltage of a tunnel-diode pulse generator (H.P. type 213B). Figs 16(b), 16(c) respectively show $e_I(t)$, $e_T(t)$, when the generator is used to drive a length of precision 10-ns, 50-Ω delay cable terminated by a Schottky barrier diode (H.P. type 5082-2301).

Depending on the conditions of the problem it is possible to choose a diode which correctly terminates the line. This is easily seen graphically. The use of a diode to suppress overshoots has been known for many years; as far as the author is aware, no



(a) (b)

Fig. 14(a) Composite time picture for $e_I(t)$, $e_T(t)$ for $0+ \leqslant t \leqslant 2t_d$ and $t = \infty$. (b) Waveforms deduced from (a).



(a) (b)

Fig. 15(a) Composite time picture when terminating component is a semiconductor diode. (b) Waveforms for (a).

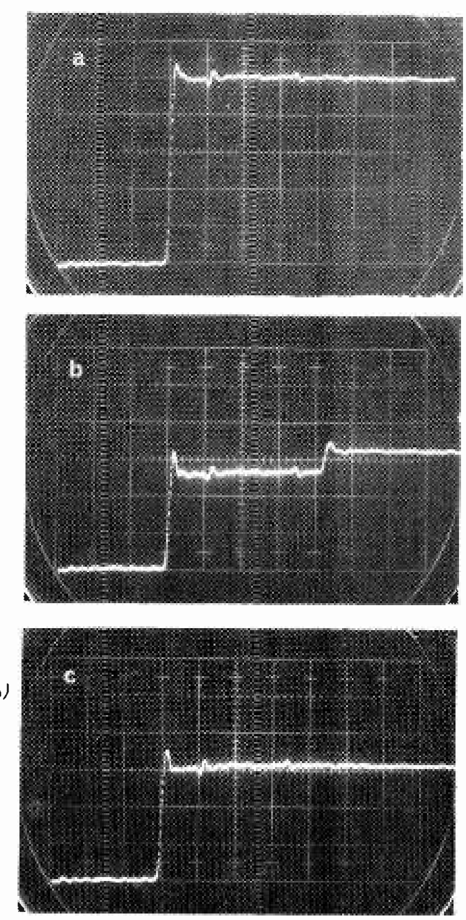


Fig. 16. Experimentally observed waveforms using matched source impedance and Schottky diode as terminating component. Horizontal scale: 5ns/div., vertical scale: 100mV/div. (a) Open-circuit output voltage of pulse generator, (b) $e_I(t)$, (c) $e_T(t)$.

exact mathematical solution of the problem has yet been worked out.

Reflections with t.t.l. interconnections

For the preservation of pulse edges and the reduction in crosstalk effects it is common practice to interconnect high-speed digital integrated-circuit logic elements with some form of transmission line; this usually means strip-line ($R_o \approx 100\Omega$) on a printed circuit board and twisted-pair between board assemblies. Now, by their very nature, switching devices and circuits have pronounced non-linear impedances associated with them.

Thus the problem reduces to investigating reflection effects which occur and seeing if we can live with them. This will be accomplished if any reflections cause no voltage overstress in the driving and driven gates and no logic malfunction, through spurious triggering, of other gates to which they are connected.

The d.c. characteristics of a t.t.l. gate are shown in Fig. 17. (The interested reader is referred to ref. 2 for a good discussion of their origin). The positive logic convention is assumed, i.e. the more positive of the two discrete voltage levels in the system is taken as '1'. A single continuous curve—(*i*) in Fig. 17(a)—gives $e_I = f(i_I)$. For each value of e_I there is a single value of e_o, but we are only interested in the e,i characteristics corresponding to $e_o \equiv$ '0' and $e_o \equiv$ '1'. These are shown respectively as curves (*ii*) and (*iii*) in Fig. 17(b), 17(c). Bearing in mind our e,i sign conventions we now superpose (*i*), (*ii*), (*iii*) to obtain the t.t.l. composite d.c. characteristics (Fig. 19) relevant to the two cascaded gates, A and B in Fig. 18, connected via a length of transmission line.

We consider only the $1 \rightarrow 0$ transition at the line input; the case of the $0 \rightarrow 1$ transition follows by analogous treatment. Thus, the output of A is initially 1 and has been so long enough for any past line transients, which may have occurred, to have died away: then $e_I = e_T = e_a$ and this corresponds to '*a*', the point of intersection of (*i*) and (*iii*). The dynamic representation of the line at $t = 0$, when e_I switches to the 0 condition, is simply a resistance R_o in series with a battery e_a; this is given by a straight line of slope $+R_o$ passing through '*a*'. Clearly, $e_I(0+) = e_I(0+ \leqslant t < 2t_d) = e_b$, where '*b*' is the intersection of this load line, extended back, with (*ii*). The sliding-source circuit for the wavefront which starts down the line is a battery e_{oc+} in series with a resistance R_o. This is depicted graphically by a line with slope $-R_o$ passing through '*b*'. Where this intersects (*i*) gives $e_T(t_d) = e_T(t_d \leqslant t < 3t_d) = e_c$. We can continue to draw our Bergeron lines and thus determine $e_I(t)$, $e_T(t)$ for whatever time scale we choose. Incidentally, we have enough information to plot $e_x(t)$, x being our general point on the line should this be required. Plots of $e_I(t), e_T(t)$ for $0+ \leqslant t \leqslant 2t_d$ are simply derived and are given in Fig. 20.

The first negative excursion in $e_I(t)$ can, if large enough, lead to a subsequent overshoot in e_T sufficient to switch B. Various methods of reducing the likelihood of this happening have been described. They include control of the input characteristic[3] designated (*i*) in circuit design and the use of some form[4] of built-in clamping diode.

Validity of graphical approach

The main assumptions of the graphical technique are; a well specified, effectively lossless, transmission path; transmission line pulses with step edges; terminating components which are completely specified by their d.c. characteristics. The inevitable departure from these ideal requirements, in practice, does not invalidate the approach. The twisted-pair is fairly reproducible in its characteristics and the lengths usually employed are unlikely to give any significant loss. Provided the pulse transition times are small compared with the electrical length of the line the effect of finite rise times and the presence of stray capacitance at terminations is to cause rounding of waveforms without having a serious effect on the amplitude of reflected pulses.

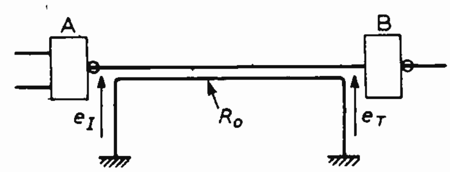


Fig. 18. Interconnected t.t.l. gates.

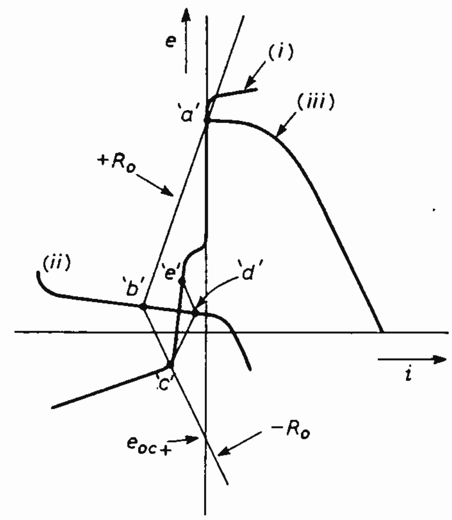


Fig. 19. Composite d.c. characteristics of t.t.l. gate with 'Bergeron' lines.

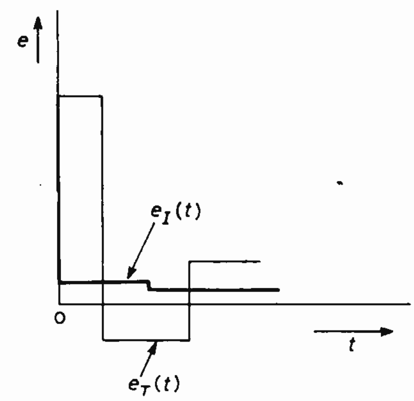


Fig. 20. Waveforms for $0+ \leqslant t \leqslant 2t_d$ deduced from Fig. 19.

It might be thought that minority carrier storage effects in the semiconductor devices constituting the t.t.l. gates might nullify the approach. However, the close agreement[5] between experimentally observed waveshapes and those predicted on the basis of device d.c. characteristics suggests that storage effects can be safely ignored—for t.t.l. at any rate.

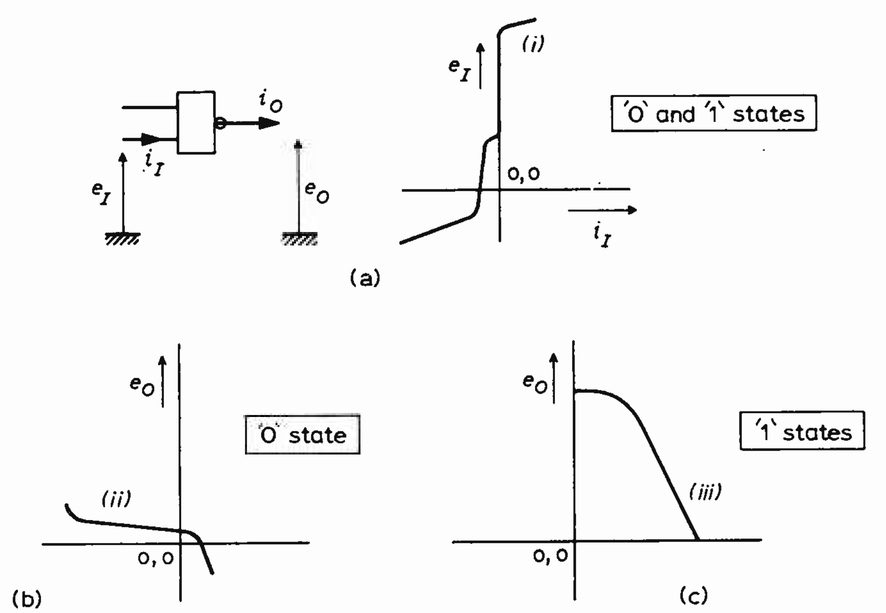


Fig. 17(a) T.T.L. gate showing sign conventions and input characteristic. (b) '0' level output characteristic. (c) '1' level output characteristic.

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Digital Stereo Sound Recorder

Elimination of timing errors: a step on the way to digital television recording

by A. H. Jones, B.Sc., and F. A. Bellis*, B.Sc., M.I.E.E.*

Sound and television signal processing is beginning to undergo radical changes with the introduction of digital techniques. Once expressed as a series of binary numbers, the signal resembles those handled by computers, and important advantages associated with computer processing can be realized. For example, provided that digital errors are avoided the original signal quality can be preserved, however complicated the subsequent operations, with a mathematical precision. Moreover, the circuits used are not susceptible to drift and so there is no need for frequent adjustment. These advantages would be particularly valuable in recording, and work is therefore on hand at the B.B.C's Engineering Research Department to investigate the possibilities and potentialities of digital sound and television recording.

To fully describe a high-quality television signal in digital form, one requires an extremely high data rate. If, for example, one samples the signal at a frequency of three times the colour subcarrier frequency and then uses 8 bits to describe each sample, the resulting serial data rate is over 100 megabits per second. The recording of the 200 gigabits of data generated by this method during a half-hour programme presents formidable problems. It seems very likely, however, that a satisfactory solution to these problems will be found and that digital methods will in due course be applied to television recording.

* B.B.C. Research Department

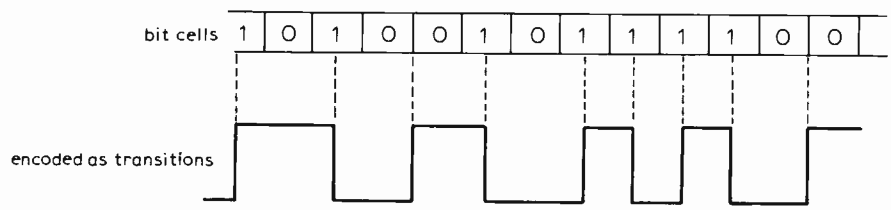


Fig. 2. Typical stream of digital data (top) encoded using "delay modulation" method.

It is as yet too early to say what form the digital television recorder of the future will take. A number of new recording technologies are being developed for the high-speed storage of large quantities of data, and maybe one of these will ultimately be used for television. They employ laser or electron beams directed at a variety of recording media, and offer the prospect of very high packing densities, but most are still at a relatively early stage of development. Meanwhile, multi-track magnetic recorders can cope with the data rate required for digital television, though at present they require a high consumption of medium as compared with high-quality analogue television recorders. Medium consumption is not an overriding factor, however, in a research programme which is devoted to the investigation of basic signal processing techniques. This article describes a digital stereo sound recorder, built as an introductory project, which contains a number of interesting features, notably timing correction, and could have a direct application in future digital television recorders.

The chief parameters of the stereo recorder are:

Tape speed	15in/sec
Tape width	$\frac{1}{2}$in
Tape	3M type 951 instrumentation tape
No. of tracks	16 (13 for data, 2 parity, 1 stereo switch)
Audio bandwidth	14.6kHz
R.m.s. signal to r.m.s. weighted noise ratio	72dB
Crosstalk between stereo channels	—45dB (occurs within the analogue input and output circuits)

Fig. 1 is a block diagram of the recorder. The two audio input signals are sampled at 32kHz, the samples being interleaved in the multiplexer to give a 64kHz sample rate at the input to the 13-bit analogue-to-digital converter. The 13 outputs from the a.d.c. are made up to 16 by the inclusion of two parity bits and a channel identification signal. After further processing the signal is recorded on 16 longitudinal tracks on $\frac{1}{2}$-inch tape.

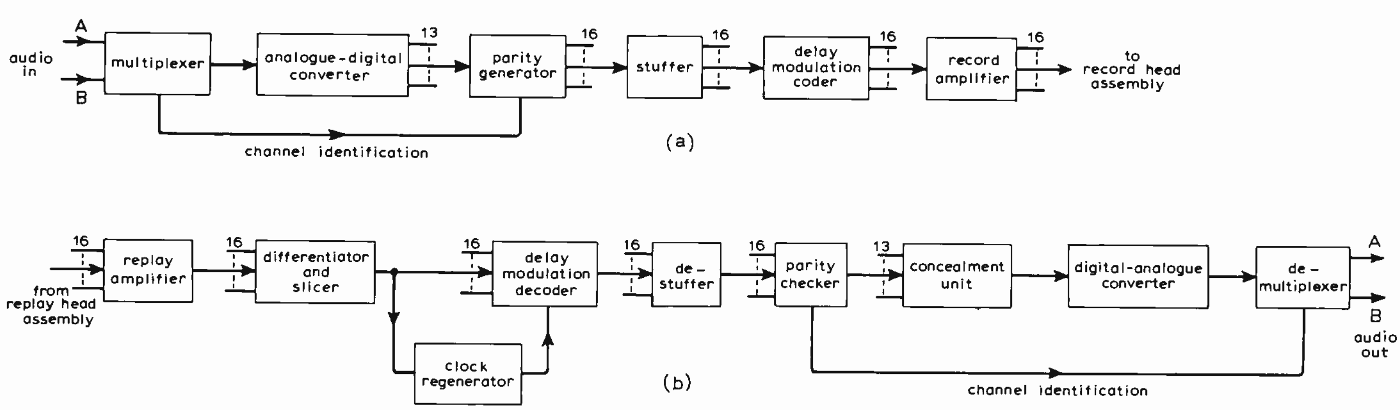


Fig. 1. Block diagram of the recorder: (a) signal processing system for recording, (b) for replay.

Thus the 13 bits corresponding to each sample of the signal are recorded simultaneously, "in parallel" across the tape.

The signal recovered on replay is processed to remove timing errors and after parity checking and error concealment is passed to a digital-to-analogue converter. The interleaved analogue samples emerging from the d.a.c. are then separated and filtered to provide the two output signals.

The recording process

One of the advantages of digital recording is that it is not necessary to linearize the recording process by the use of bias. This leads to a considerable simplification, as there is no need to provide a bias oscillator and the recording amplifier need not be linear. Indeed, all that is necessary is an electronic switch, controlled by the data to be recorded, which changes the direction of a fixed current through the head to magnetize the tape in one direction or the other.

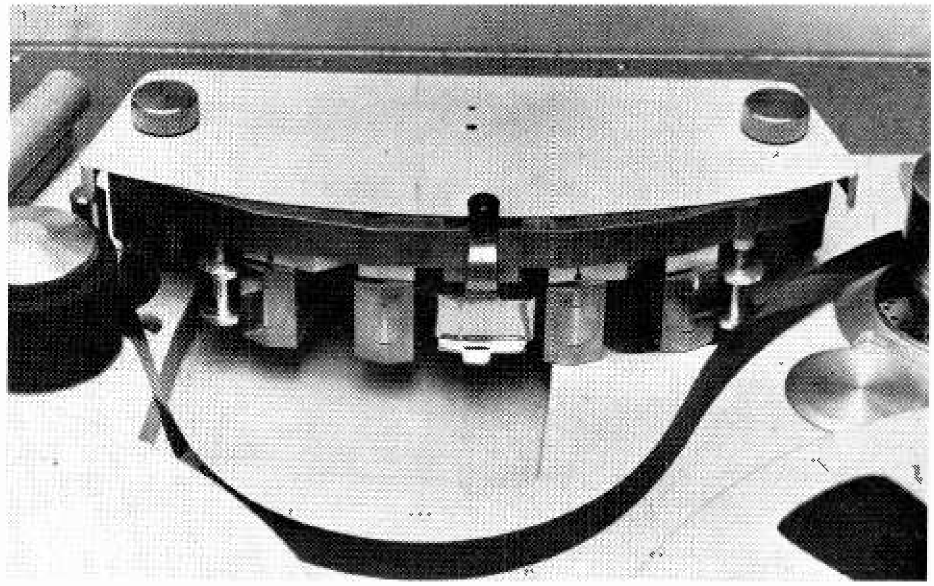
Digital data are commonly represented by two voltage levels, one for a digital "0" and the other for "1", and it might at first sight be thought that all that is necessary to record this information on magnetic tape would be to represent a 0 or a 1 by opposite directions of magnetization. However, the data may well contain long strings of noughts or ones, and as the replayed signal is proportional to the rate of change of recorded flux, long periods could then occur in which there was no output from the replay head. This is an undesirable feature and the signal must therefore be re-encoded in such a way that frequent changes of recorded flux take place even when the data stream does not change. A number of coding techniques have been proposed; the one chosen for this application is known as "delay modulation" (sometimes also called Miller code).

For delay modulation the rules are as follows: A transition (polarity unspecified) occurs in the centre of bit cells in which 1s are being conveyed. Transitions also occur between adjacent 0-bit cells.

Fig. 2 shows the signal obtained by applying this coding method to a typical string of data. Note that polarity is unimportant, thus no special precautions need be taken to specify such things as direction of head windings, etc. This is a useful advantage when many channels are being considered.

Signal regeneration on replay

Fig. 3 shows at (a) a two-level signal applied to a recording head. The replayed voltage, given at (b), is formed from positive and negative pulses whose peaks correspond to the transitions in the recorded signal. The width of these pulses depends on the magnetic properties of the tape and the resolution of the replay head. When high packing densities are being attempted, the pulses overlap, but the data can still be recovered provided that inter-pulse crosstalk is not too severe. Fig. 3(c) shows a commonly used method of data recovery, which is



The experimental digital recorder uses a modified audio deck. The 16 recording heads are arranged in two staggered 8-head assemblies; likewise the 16 replay heads. Signal processing functions are carried out largely by integrated circuits mounted on printed circuit boards.

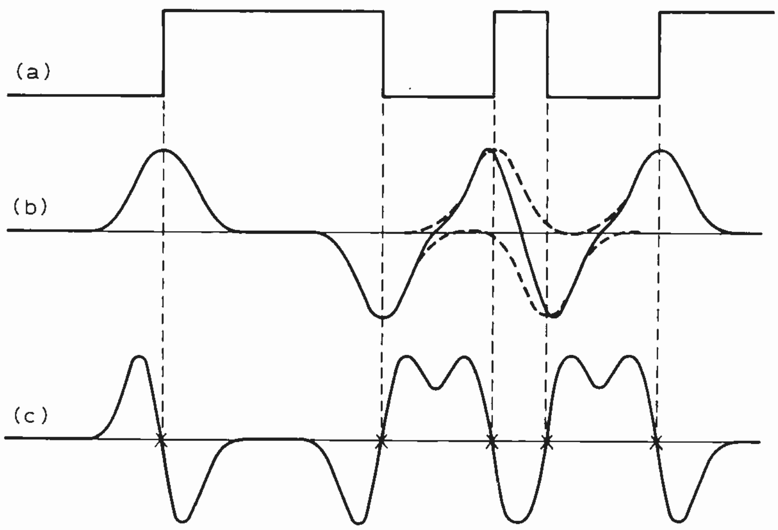


Fig. 3. Basic waveforms: (a) two-level signal applied to recording head; (b) replayed voltage; the broken lines show individual replay pulses and the solid line the composite waveform; (c) waveform resulting from differentiating (b); from this the recorded signal can be obtained using the information of the zero crossings.

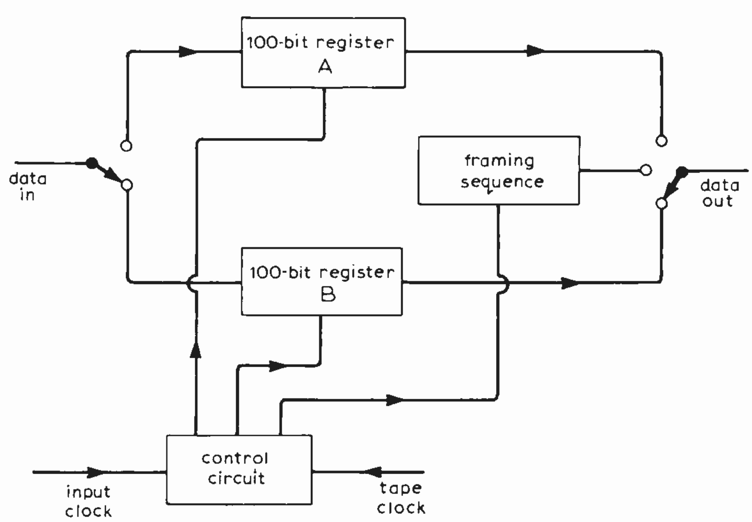


Fig. 4. "Stuffer" used for insertion of framing pulses for timing correction.

employed in the sound recorder. The replayed signal is passed through a differentiating circuit, the output of which has zero-crossings at times corresponding to the input peaks. The recorded signal may now be recovered by slicing. As Fig. 3 indicates, this method relies on some degree of inter-pulse crosstalk because the output of the differentiator returns to zero if the transitions are widely spaced. This is one reason for using the modulation technique described above.

Timing correction

The signal recovered from the tape is subject to two types of timing error, static and dynamic. Dynamic errors are caused by irregularities in the transport system and are usually evident as "wow" and "flutter". Static errors are those caused by errors of head alignment; when working at high packing densities only a small error in the alignment of the replay heads relative to the recording heads can cause the bit streams obtained from different tracks to be incorrectly associated. However, digital techniques allow for the removal of both types of timing error, and with the appropriate circuits installed, the sound recorder is unaffected by gross skew errors and timing fluctuations.

In broad terms, the timing correction is carried out as follows. Time markers, or framing pulses, are inserted into the data in each channel at regular intervals before recording. On replay, the data stream is clocked into temporary stores at times governed by the arrival of the associated framing pulses. The signal is arranged in the stores in such a way that when clocked out by crystal-controlled "station clocks" (i.e. without timing errors) the data that immediately follow the framing pulses emerge simultaneously in each channel.

In choosing the form of the framing sequence, several factors had to be considered. The sequence must not be so long as to occupy a disproportionate amount of time, nor must it occur so frequently as to raise the data rate by an unnecessarily large amount. It was decided to have a sequence 8 bits long, repeated after every 100 input data bits, the actual sequence being 101011*xx* where *xx* indicate two additional bits comprising a "label" used to convey elapsed time and a tape reference number, i.e. they do not form part of the framing sequence proper. It is desirable to include the sequence 101 in the framing sequence to facilitate the correct decoding of delay modulation.

The insertion of framing pulses is carried out in the circuit marked "stuffer" in Fig. 1; this is shown in greater detail in Fig. 4. It has two main sections, that containing two 100-bit shift registers, A and B, and that which generates the necessary pulses to control them. Although the block diagram shows only one channel, shift registers and input and output switching are provided for each of the sixteen channels, while the clock generation and "housekeeping" portion controls all sixteen channels.

Data is clocked at the input rate into the two 100-bit shift registers alternately, under the control of the input switch. While the data stream flows into one of the registers, it is clocked out of the other, but at a slightly faster rate (108/100 × input rate). Thus each register is emptied before the other one is completely filled, leaving time to insert the 8 framing digits. Thus the data string is broken into 100-bit blocks, with synchronizing information between them. It is important to note that this is happening simultaneously in all sixteen channels and that the framing pulses occur at the same instant in all of them; thus synchronizing information is provided for removing static as well as dynamic timing errors.

The delay modulated data contains a strong component at half the clock frequency which can conveniently be extracted on replay and used to control the circuits which clock the replayed data. Because of skew errors, separate clock regenerator circuits are provided for all of the sixteen channels. Their outputs are, of course, subject to the same timing fluctuations as is the data stream itself.

Timing errors are now removed within "de-stuffer" circuits, one of which is shown in greater detail in Fig. 5. Its operation is similar to that of the stuffer, in that the data is clocked at different rates into and out of an array of shift registers. In the de-stuffers, however, the steering of data between the shift registers is controlled by the framing sequence.

Referring to Fig. 5, data is first clocked through a short shift register in which the presence of the framing sequence is detected. The arrival of this sequence operates the input selector switch and the next 108 bits of data are clocked into, say, shift register X, under the control of the tape derived clock (the last-mentioned having the same timing errors as the data). As the register is only 100 bits long, the 8 bits of the framing sequence are lost, and register X is left containing the same block of data which was originally in one of the stuffer shift registers. By the time register X is full, the next framing sequence arrives and switches the data into register Y which is in turn filled, and so on. This process occurs in each of the sixteen channels independently under the control of its own clocks and framing sequences.

Clocking out of the registers may now take place simultaneously in each of the channels, under the control of the slightly slower station clocks formerly used to generate the data. Thus both static timing errors, due to skew, and dynamic ones, due to wow and flutter in the transport mechanism, are removed. It will be noted that because the de-stuffer has 3 × 100-bit shift registers, the delay which it introduces into the data can vary from 100 to 200 bits without violating the condition that reading and writing must always be taking place in different registers. Thus timing fluctuations of $\pm\frac{3}{4}$ ms can be accommodated. A low bandwidth servo control of the tape capstan keeps the timing errors within these limits.

In addition to steering the data between the shift registers the arrival of the framing sequence in the short 8-bit shift register initiates the decoding of the last two bits which constitute the label.

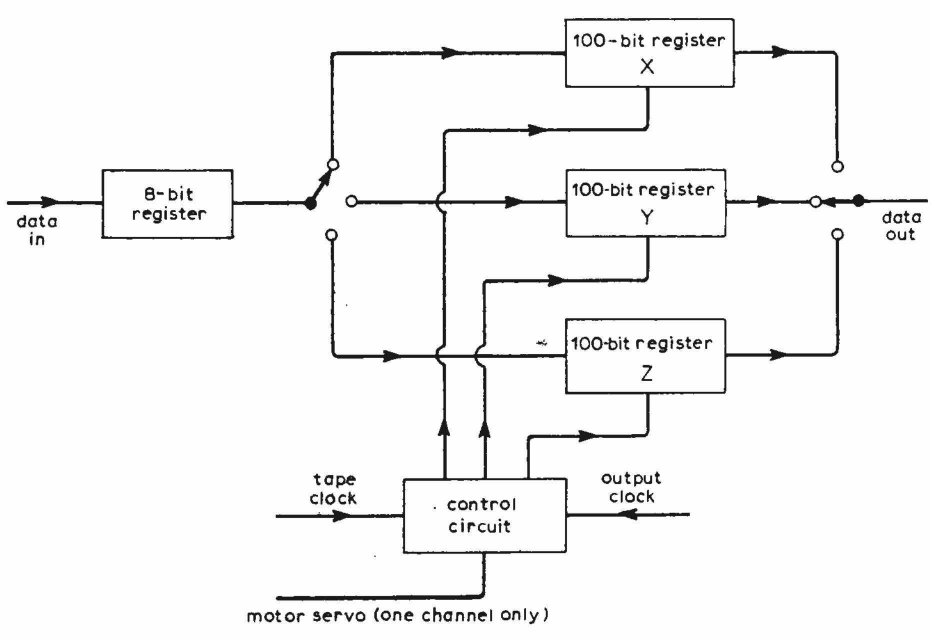


Fig. 5. "De-stuffer" for removal of timing errors.

Error concealment

The effect of errors in a digital system is to cause gross disturbances to the decoded analogue signal; in the case of sound signals these disturbances are heard as loud clicks and crackles. The magnitude of the interference is dependent on the significance of the digits affected; if digits of high significance are in error, then some action to conceal the errors is essential. In the case of randomly occurring single digit errors, it is generally sufficient to omit each erroneous word, repeating instead the previous word and thus arresting the analogue signal until the next word arrives.

Errors in a magnetic recording system are generally caused by "drop-outs" due to imperfections in the tape or temporary separation between the tape and the heads. When high packing densities are used, the duration of drop-outs can be such as to cause a prolonged burst of digital errors. The above technique of retaining the last good word until the end of the disturbance continues to give some improvement, however, even under these conditions, and this form of error concealment is therefore employed in the sound recorder.

Two of the sixteen tracks are provided for parity bits generated to facilitate the detection of digital errors. One of the parity bits is associated with digits 1 (most significant bit), 3 and 5, the other is associated with digits 2 and 4. This assignment of parity bits was chosen because the location of the tracks carrying digits 1 to 5 is such that there is only a very small risk of simultaneous errors in two or more of the associated tracks. A simple parity technique can therefore be used.

Performance

The performance figures mentioned above as chief parameters of the recorder are those which result when the a.d.c. is connected directly to the d.a.c. and are not altered when the rest of the circuitry including the magnetic tape system is inserted. The a.d.c. and d.a.c. do not give the optimum 13-bit performance; had they done so, the signal-to noise ratio achieved would have been about 3dB better, namely 75dB.

The packing density of about 5k bits per inch per track at which the machine operates represents a very modest requirement in this application. It is possible to improve packing density and therefore reduce tape consumption by at least 2:1, at the cost of more expensive heads.

The timing correction circuitry is, as mentioned above, capable of removing "wow" and "flutter" up to a maximum timing error of $\pm \frac{3}{4}$ ms.

The simple error concealment technique works reasonably well, though some impairment to sound quality is still apparent when lengthy drop-outs are encountered. Further work is needed to provide improved error protection in these circumstances.

The experimental stereo recorder has established the feasibility of digital sound recording and has provided a convenient test bed for the development of a number of useful processing techniques, notably those relating to timing correction. The application of these techniques to the digital recording of high-quality television signals is now being investigated.

Acknowledgements. The authors wish to acknowledge the contributions and help received from their colleagues in the B.B.C. Research Department; they also wish to thank the Director of Engineering of the B.B.C. for permission to publish this article.

The Domestic Video Recorder

Magnetic-tape cassette machines available at £300

The first British radio and television set manufacturer to produce a video cassette recorder at a "domestic" price level put the machine on view to the public at the Internavex 72 show (International Audio-Visual Aids Conference and Exhibition) at Olympia, London, from 25th to 28th July. The machine is the British Radio Corporation's model 8200 Colour Video Cassette Recorder, which is selling at "about £285" (without cassette) and is available for renting at "about the same payment as for a 26in colour television set, approximately £96 per year". At present the video recorder is available only to organizations — schools, colleges, industrial training establishments, hospitals etc. — through an associated company in the Thorn group, Radio Rentals Contracts Ltd (1-15 Clyde Road, Tottenham, London N.15). There is no commercial machinery for retail buying or renting. In fact the recorder is being presented, for the moment, as basically a low-priced professional machine, but B.R.C. are making no secret of the fact that they see it as their first domestic video cassette recorder when the domestic market gets under way.

What makes this machine the probable archetype of future domestic recorders is the fact that it is a cassette equipment,* allowing simpler operation than with open-reel recorders, and makes use of the well tried and inexpensive Philips deck. The Philips VCR equipment, using $\frac{1}{2}$-inch tape, has been commercially available direct from Philips Electrical Ltd for some time (see July 1970 issue, p.340), and recently Philips have been licensing receiver manufacturers (mainly in Europe) to use their recording "format" — which in practice means using the deck made at this company's tape recorder factory in Austria. For example, Blaupunkt, Bosch, Grundig, Loewe-Opta and Telefunken are among the German licencees. In Britain, Philips Electrical are selling the latest machine — the type N1500, which also was on show at Olympia — again only to organizations. Their original projected price of £230 (July 1970) has not been adhered to, and the machine now costs £315. This, however, includes a television tuner, whereas the lower-priced B.R.C. recorder does not have a tuner, so, if the B.R.C. machine is needed to record programmes "off the air", it must be used in conjunction with a television receiver. A 26-inch schools colour TV receiver, the model 8733, and a monochrome receiver, were shown at Olympia with the B.R.C. recorder.

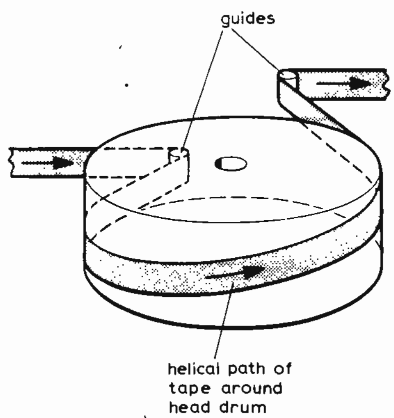


Fig. 1. Helical-scan principle: the tape passes round the head drum in a helical path; also the rotating heads trace helical magnetic paths obliquely across the tape.

Another cassette video tape recorder, put on show by E.M.I. as distributors, was the Sony VO-1700, working in conjunction with a Trinitron monitor. This, however, is priced at "about £600", uses a $\frac{3}{4}$ inch tape and has been designed to operate on the N. T. S. C. 525-line colour television standard, whereas the Philips and B.R.C. machines shown work on the European 625-line PAL standard. A video cassette recording made on this Sony machine can only be played back on the same machine or on two or three other Japanese made cassette recorders. There is no interchangeability of cassettes possible, or "compatibility", to use the current phrase, between the Sony machine and the Philips and B.R.C. machines. Even when a version of the Sony cassette recorder designed to work on European television standards becomes available, the Sony and Philips systems will not be compatible.

The question of compatibility, and of video tape recording standards generally, is a very vexed one. A video tape standard includes not only the standard of the

*The term "cassette" is used in this report to include what is called a "cartridge". There are basically three types of cassette: (a) a rectangular box containing two tape spools in the same plane; (b) a rectangular box containing two spools stacked vertically on the same axis; (c) a flat cylindrical or rectangular container holding a single spool (cartridge), from which the tape is drawn while recording or playing back and on to which the tape is re-wound afterwards.

television broadcasting system providing the signal to be recorded and afterwards played back into a television receiver, but a whole host of mechanical quantities, such as tape width, tape speed, number of recording/playback heads, the head drum diameter and the geometrical layout of the magnetic tracks carrying vision and sound information formed by this machinery on the tape. First of all there is no world standard recommended by an international organization. Several bodies, such as the C.C.I.R., the E.B.U. and the S.M.P.T.E., are studying temporarily established commercial "standards"—resulting from several recorder manufacturers agreeing to use the same basic design of machine — possibly with a view to recommending official standards. But behind these technical investigations fierce commercial battles are being fought to see who can sell the greatest numbers of machines of particular designs, for no doubt the winners will establish the ultimate standards.

The possible standards for domestic video recorders are automatically divided into two groups by the principal colour television broadcasting standards — the 625-line, 50 fields/s PAL system in Europe and the 525-line, 60 fields/s N.T.S.C. system on the American continent and in other countries. There is a further division, into broadcasting quality standards, which means large machines costing in the region of £20,000, and the standards used for small semi-professional recording, which at present means machines in the price region of £300 to £1000. It is the last-mentioned category, for both 625-line and 525-line television standards, which includes cassette and open-reel operation, that will almost certainly provide the basis of the domestic video recorder of the future.

Removing a cassette from the British Radio Corporation's colour video cassette recorder, model 8200.

The Sony type VO-1700 video cassette recorder, using $\frac{3}{4}$ inch tape, with a Trinitron colour receiver acting as a monitor.

The main reason for this is that these machines use a technique for obtaining adequate head-to-tape "writing" and "reading" speed (in the region 500-1000 inches per second), called helical scan recording, which allows them to be mass-produced at the lower prices mentioned. Helical-scan recorders are so called because a narrow tape ($\frac{1}{4}$in, $\frac{1}{2}$in $\frac{3}{4}$in or 1in) is wrapped round a drum, incorporating rotating recording heads, in what amounts to part of a helix (Fig.1). In addition, if the tape is considered as a helical "slice" belonging to an imaginary cylinder of magnetic material rotating round the head drum of the machine it can be seen from a vector diagram that any point on the cylinder has a vertical component of velocity, parallel with the cylinder axis, as well as an angular velocity due to the rotation. Consequently the path traced by a rotating recording head on the inside of the imaginary cylinder, and therefore obliquely across the tape, is also helical.

There are at present four important and contending standards for helical-scan recording — "important" in the sense of having been extant for several years and having established substantial markets for recorders: (1) the Philips $\frac{1}{2}$-inch tape standard, as seen in that company's video cassette recorder and those of licensed manufacturers; (2) the Sony $\frac{3}{4}$-inch tape video cassette standard, also being used by other Japanese manufacturers; (3) the Electrical Industries Association of Japan (E.I.A.J.) standard for $\frac{1}{2}$-inch tape, to which a group of Japanese manufacturers (e.g. Ikegami, Matsushita, National, Nivico, Sanyo, Shibaden and Sony) are supplying open-reel machines, and Ampex a cassette machine; and (4) the Cartrivision standard, of the American company Cartridge Television Inc., which uses $\frac{1}{2}$-inch tape in cassettes. Since the various manufacturers see the possibility of export markets in each other's countries these "standards" must not be linked too closely with their names. For example, as mentioned above, Sony are going to produce in Japan a version of their $\frac{3}{4}$-in video cassette machine for the European 625-line television standard, while Philips are intending to make in Austria a version of their video cassette recorder adapted to work on the 525-line N.T.S.C. television standard and conforming to the Japanese E.I.A.J. standard.

In a forthcoming issue we shall publish a more detailed analysis of techniques and standards used in the lower priced helical-scan video recording machines.

10 – 80 Metre Amateur Transceiver

4: Power supplies, alignment and performance evaluation

by D. R. Bowman, G3LUB

The power supply has to provide 750V at 120mA for the p.a. anode; 200V at 15mA stabilized for the p.a. screen grid; 350V at 50mA for the driver stage; −40 to −60V (variable) p.a. stage bias; 11V at 200mA stabilized for the transistor sections; 8V at 7mA stabilized for the v.f.o.; and 6.3V a.c. at 1.85A for the valve heaters (or 12.6V at 0.925A for mobile use).

Heater supplies

The p.a. and driver valves require heater supplies which are most conveniently provided by connecting the two valve heaters in parallel to a 6.3V winding mounted on the main transformer (Fig. 15). The author in his efforts to make the transceiver suitable for mobile use connected the heaters of the 12BY7A in series and then in parallel with a 12V heater (special) 6146, thus the heater requirements become 12V at 1A suitable for either car battery or transformer operation. If such a 6146 valve is not available then a 6W 5Ω ballast resistor should be connected in series with the p.a. valve heater. Please note that neither side of the heater winding is connected to earth although a number of 10nF bypass capacitors are used to reduce the r.f. feedback that could occur along the heater lines. The reason for leaving the heaters floating will become obvious when we deal with the low voltage d.c. supplies.

Low-voltage stabilized supply

The semiconductor sections of the transceiver are supplied from a positive 11V stabilized rail with the critical variable frequency oscillator obtaining its supply from a zener stabilized line at 8V. To minimize the number of transformers required, various ways of using the heater secondary to provide the 11V d.c. were investigated. Most mains transformers have at least two 6.3V secondaries and it was found that if they were connected in series and used to feed a bridge rectifier then enough voltage and current became available to supply the low voltage circuits. If one examines the circuitry (Fig. 16) it will be seen that the series stabilizer is connected in series with the negative-to-earth line. This allows the use of n-p-n power and p-n-p low-power transistors forming the most economical arrangement. It is also important to use silicon devices throughout. The stabilizer is short-circuit proof and has a considerably higher capacity than is required. In view of this excess current capacity a 12V transmit/receive changeover relay is used. Almost any n-p-n power transistor could be used to replace the BD123 and any signal level n-p-n device the BFY50. The author recommends that type 2N3702 be used in the differential pair but, if it is necessary to substitute, then, due regard must be taken of the collector voltage rating of Tr_3. This transistor must have a collector emitter voltage rating in excess of the input d.c. appearing across C_5 minus the zener voltage (D_5). Almost any rectifier diodes may be used in the D_1–D_4 bridge as the maximum current drain is only 200mA.

Resistor R_2 was included in the circuit because in rather special circumstances the power supply could remain paralysed upon switch-on when very large loads were permanently connected. If it is intended to experiment with loads of only a few mA then the value of R_2 should be increased to at least 1kΩ.

The ultra stable rail for the v.f.o. is obtained using an 8V zener diode D_6. This simple technique has proved quite adequate and there are no traces of f.m. present on either the received or transmitted signal.

This description has been rather superficial as the complete stabilized power circuit was dealt with in the August 1970 issue of *Wireless World*.

> **WARNING**
>
> During the alignment and testing procedure the constructor is asked to make several measurements using a high impedance meter fitted with an r.f. probe. The probe will contain a capacitor of around a 1nF in value that can charge up to about 1kV during measurements on the valve sections of the transceiver. If the next measurement is on the transistor sections the high voltage charge will destroy any semiconductor that gets between it and earth; even the gate protected devices will not stand up to this sort of treatment.
>
> WHEN USING THE R.F. PROBE DISCHARGE THE PROBE TO EARTH AFTER EVERY MEASUREMENT.
>
> It is very easy to forget and mistakes could prove very expensive.
>
> **The voltages associated with the high voltage power supply and p.a. stage could quite easily prove lethal, so please be very careful.**

High-voltage power supply

The high-voltage power unit is called upon to supply 750V at 125mA maximum for the p.a. anode, 200V stabilized at 15mA maximum for the p.a. screen grid, −40 to −60V (variable) at negligible current for the grid bias, together with 50mA at 350V for the driver stage. Considerable thought went into the circuit design as it was realised that a large weight saving could be made if only one transformer were to be used for all the transceiver's needs. A 300–0–300V at 120mA transformer was chosen. This may seem to greatly limit the power available but it must be realised that the p.a. anode supply does not require good d.c. regulation but only dynamic stability. The average anode current "kicks-up" in sympathy with the microphone waveform and it is found that a capacitance of 50μF produces more than adequate dynamic voltage regulation. To produce 750V supply the whole secondary is employed, the centre tap is then used to provide the 350V line. Neon stabilizers connected via a ballast resistor produce the 200V stabilized screen potential from the 350V line. The centre tap is used to obtain the negative grid potential via two diodes. This convenient use of the transformer was paid for by having to tolerate half-wave rectification.

In order to avoid p.a. white noise being picked up by the receiver and to reduce valve dissipation it was decided to bias the output valve off during receive periods. This is achieved by using a normally open pair of contacts RL_{1d} to short-circuit a resistor in the negative line potential divider changing the bias line to about −20V. This reduces the p.a. standing current to zero.

In use the complete power supply has been found to be reliable and cool in operation. The only components requiring special note are the high-voltage smoothing capacitors which must have a rating of at least 450V. Other components are safe with 350V ratings.

It is unlikely that a prospective constructor will have available an identical meter to that used by the author and therefore all current shunts noted on the circuit diagram will have to be changed in value. In view of the competence required to build the rest of

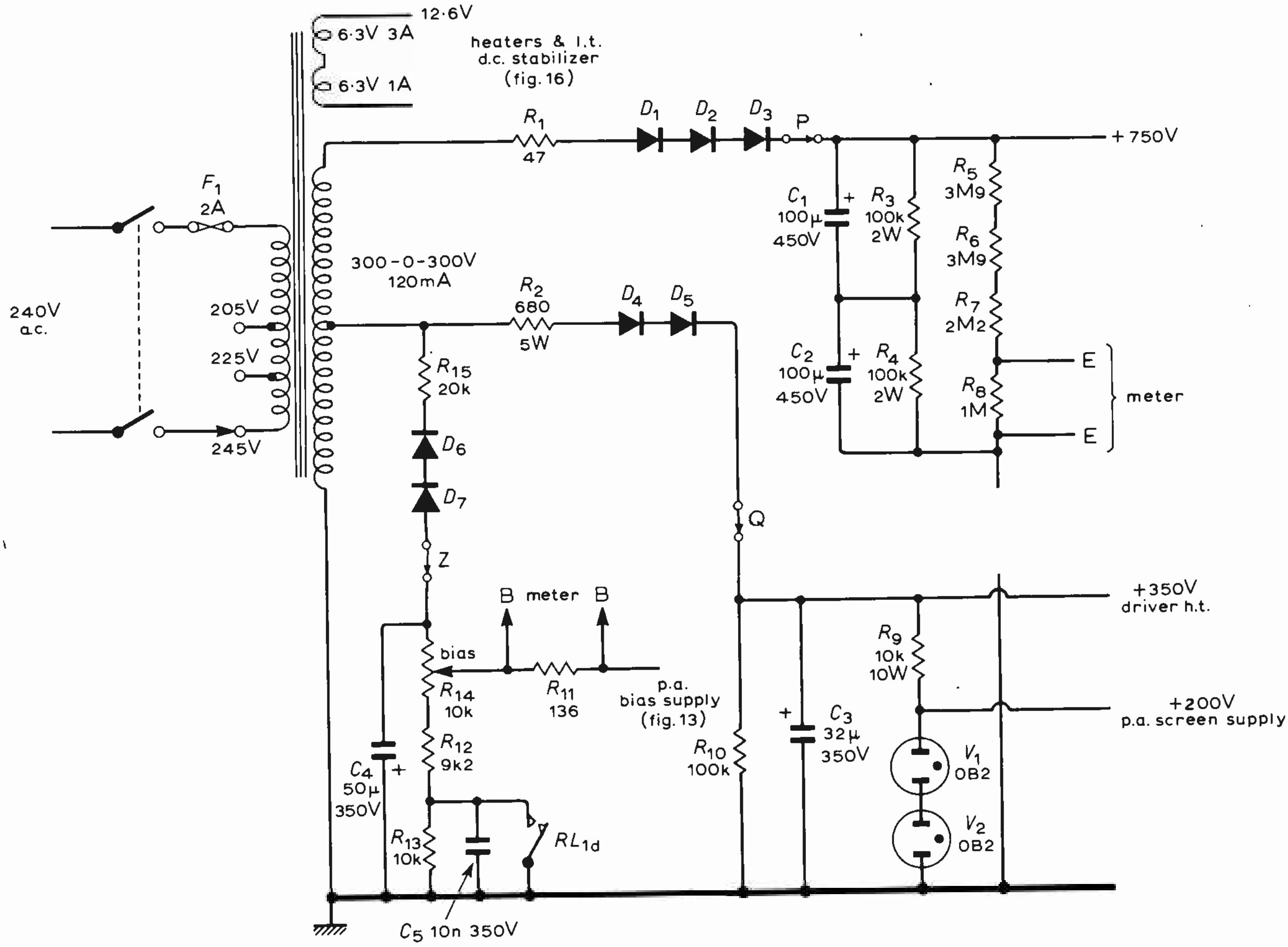


Fig. 15. High-voltage power supply unit. The contact across C_5 is one on the transmit/receive relay. The lettered links shown in this drawing (and in Fig. 16) are shorting links in a plug fitted to the rear of the transceiver so that the unit can easily be supplied from a mobile power supply.

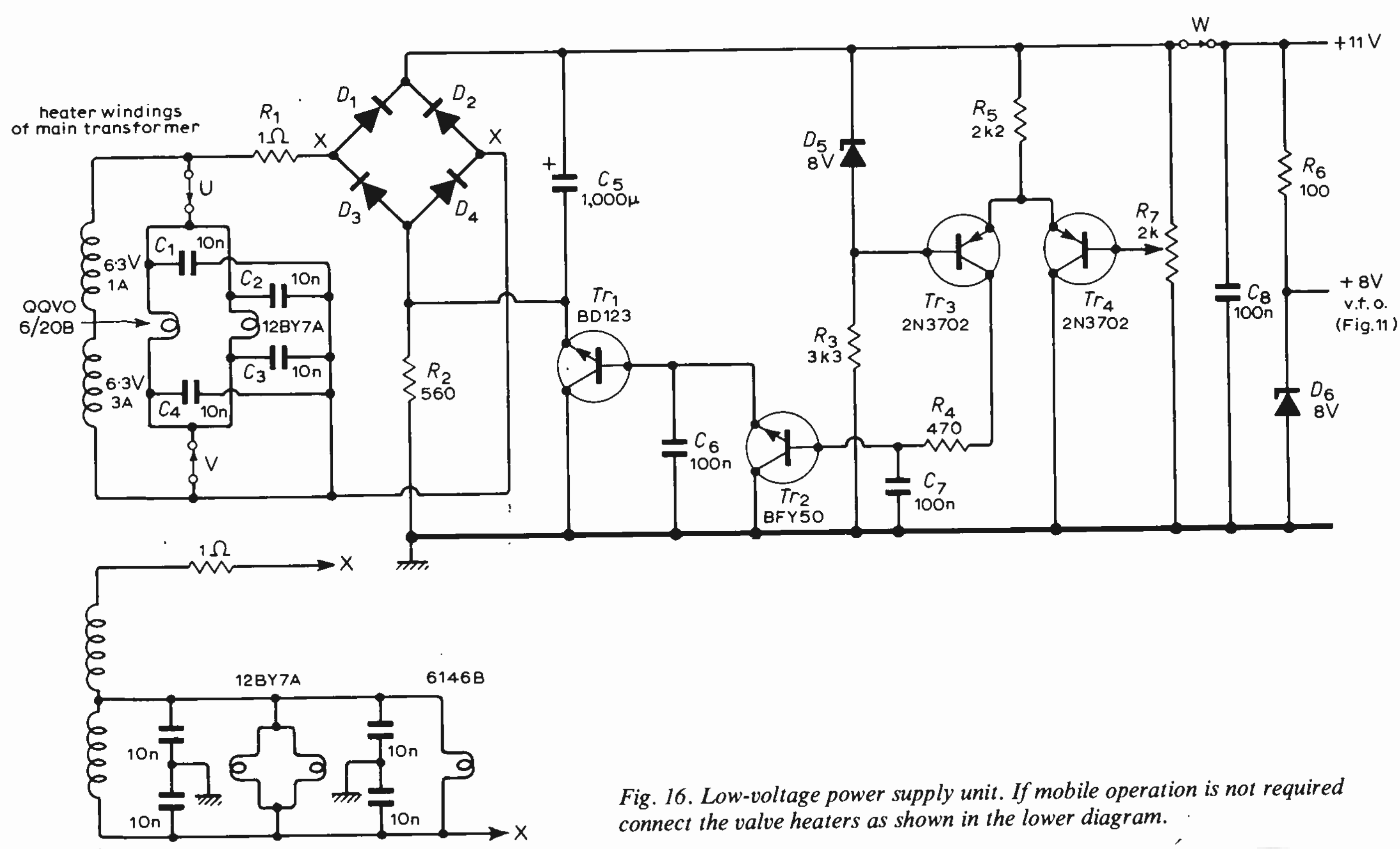


Fig. 16. Low-voltage power supply unit. If mobile operation is not required connect the valve heaters as shown in the lower diagram.

the transceiver it has been decided to leave the individual calculations for the meter shunts to the constructor.

Alignment procedure

We are now nearing the end of this series and some details of suggested alignment and commissioning procedures are in order. Once again the author would suggest a careful plan for the commissioning process. The first rule is never to build a piece of equipment completely and then on some fateful day to switch on the unit. In the author's experience, unless the unit is very simple this procedure is doomed to failure. All complex electronic equipment suffers from what are colloquially termed bugs and there seems no reason to suppose that this transceiver should be different. A constructor's enthusiasm and drive has practical limits and therefore by far the best procedure is to test and align individual stages as they are completed.

Now we come to the need of test equipment. The less test equipment available the more the improvization and aptitude of the constructor is exercised. It is possible to construct this transceiver while having access only to a good high impedance voltmeter equipped with an r.f. probe. Some form of h.f. signal generator is almost essentail as is an all-band receiver of at least AR88 standard. An amateur-band-only receiver could possibly be substituted with some extra difficulty.

During the alignment description further equipment will be mentioned; do not despair if such is not available as there are usually other ways of doing each job.

Receiver

It is suggested that the alignment is subdivided; those sections leading to quick results being dealt with first. The aim is to commission the receiver section placing most effort in the 80m band. When this is working well then the emphasis is transferred to the transmitter section again limiting work to the 80m band. This technique allows the use of the transceiver on the air at the earliest possible time thus helping to maintain the constructor's interest. Next the 20m band is commissioned followed by the most critical pre-mix range namely 10m, the other frequencies then following with few problems.

Audio amplifier: Throughout these instructions reference will be made to the various circuit diagrams which were published in the second and third parts of this series.

Fig. 10 (part 2) shows the audio amplifier which consists of Tr_{4-7}. A multimeter switched to read about 50mA f.s.d. should be connected in series with the 11V supply to the amplifier. With no input signal and R_{24} at minimum resistance a careful note of the current should be made. Potentiometer R_{24} should be carefully increased in value until the current consumed, as indicated on the meter, has increased by about 1 to 3mA. It is essential not to increase the standing current excessively. A resistive load of about 3Ω instead of the loudspeaker should be connected across the output terminal and if an audio signal generator is available it should be connected across R_{25}. With the audio gain R_{25} at maximum and about 10mV r.m.s. supplied from the generator an output of about 200mV should be measured across the load. Obviously if no test equipment is available then this test can be ignored. The various oscilloscope and voltmeter measurements listed in the tables are all r.m.s. values (35% of the peak-to-peak voltages).

Beat frequency oscillator: The b.f.o. should be switched on (Fig. 10) and about 1.5V oscillator voltage should be present at the collector connection of Tr_1. Later C_{22} and C_{23} will have to be set so that the b.f.o. frequency is about 20dB down either side of the filter response.

Product detector: With the b.f.o. disabled all the d.c. potentials noted in the tables should be checked and once these are approximately correct one can proceed. A 9MHz signal generator is connected via a coupling capacitor (10nF) to the base of Tr_3 (Fig. 10) then as the generator is swept in the vicinity of 9MHz a whistle going through zero beat should be heard from the loudspeaker. Do take care not to drive the audio output stage continuously or Tr_6 and Tr_7 may overheat. The signal generator should be set for about 50mV output.

Intermediate frequency amplifier: Throughout this description it will be assumed that the constructor has made d.c. measurement checks as outlined in the various tables. Very great care is necessary when making measurements on high-gain i.f. amplifiers. The amplifier may be perfectly stable when input and output cables are connected, but become uncontrollable when test clips are haphazardly connected to the circuit. The r.f. probe should be connected to the base of Tr_7 (Fig. 7). A 9MHz generator should be connected via a miniature 10nF capacitor to the base of Tr_6. With the generator set to about 9MHz the core of L_{25} may be adjusted for maximum r.f. reading. Having completed this change the injection point to the base of Tr_4 and continue by adjusting both L_{24} and L_{25}. If difficulty is experienced it may help to disconnect the a.g.c. line from the collector of Tr_7 and manually control the gain from the centre tap of a potentiometer whose ends are connected from h.t. to earth; the centre tap going to the a.g.c. line. Almost any value of potentiometer between 1 and 25kΩ could be used. By doing this a completely stable amplifier should be obtained by turning the potentiometer towards the zero potential end. Lastly, connect the generator to the base of Tr_2 and adjust all three i.f. transformers L_{23}, L_{24} and L_{25}. It should be possible once the r.f. voltmeter is removed to hear microvolt signals being supplied by the generator producing whistles from the loudspeaker. The final maximum gain for the i.f. amplifier is determined by the value of R_{20}. This setting is best carried out at the end but will be described here so as to avoid excessive hopping about within the description. When the receiver is complete it will be found that with no input signal the collector of Tr_7 will settle at about 7V (measured with a high-impedance voltmeter). At maximum r.f. gain this voltage should be reduced by adjusting the value of R_{20} until the front-end noise, even on 10m, is just audible. If it is not reduced enough the large signal handling will be degraded and if the gain is too low the sensitivity will deteriorate. When the receiver is driven with a 50mV signal no distortion should be heard and the d.c. voltage at the collector of Tr_7 should be of the order of 2V.

Crystal filter: At this point the filter should be connected into circuit. The signal generator may be connected to the drain of Tr_2 (Fig. 6) via a 10nF capacitor. If whistles can be heard then move the generator injection point to gate 1 of Tr_2. Once again if signals can be heard, remember the generator must be set very close in frequency to 9MHz; trim L_{11} for maximum signal.

Signal strength meter: This circuit is simply a high-impedance voltmeter where the zero (i.e. no signal in) is adjusted with R_{32} (Fig. 7) and the sensitivity by R_{31}. The two trimmers will be found to be interdependent. Having commissioned the *S* meter, the signal generator should be fed once again to gate 1 of Tr_2 (Fig. 6) using a fairly low signal strength (say 50μV) for the final adjustment of the b.f.o. crystal frequencies. Trimmers C_{22} and C_{23} (Fig. 10) should be adjusted so that the zero beat points occur about 20dB down either side of the filter response as measured using the *S* meter. One trimmer capacitor adjusts one crystal frequency while the other trimmer adjusts the other crystal. All i.f. transformers should have been peaked at the centre frequency of the crystal filter which is nominally 9MHz.

Variable frequency oscillator: The alignment of the v.f.o. should commence by disconnecting one end of C_{16} (Fig. 11) thus disabling the independent receive/transmit control. This should be followed by connecting +8V from either the power unit or from an auxiliary battery to the circuit. With the r.f. probe about 800mV should be measured across R_6. As C_1 is rotated the frequency generated will sweep from approximately 5 to 5.5MHz. The output circuit of Tr_2 consists of a wideband coupler and should be aligned using the following technique. Set C_1 to about 120kHz from the high-frequency end of its travel. Reduce C_{12} to minimum capacitance and adjust L_2 and L_{13} for maximum r.f. as measured on either gate 1 of Tr_4 or Tr_5. This completed, slowly rotate C_1 (to lower the frequency) until the r.f. measured drops to half the previous level. This 3dB down point will be found to move further and further from the peak point as the capacitance of C_{12} is increased. With care it is possible to achieve a double humped response where the output voltage drops approximately 3dB at 5, 5.25 and 5.5MHz. This adjustment should be finalized when the receiver is complete.

High-frequency amplifier/mixer: Probably the easiest way to align the balanced mixer (Tr_4, Tr_5 Fig. 11) is to complete the wiring of the front end mixer and r.f. stages. With this completed a large signal, say 50mV, from either a signal generator or the aerial may be fed to the receiver's input terminals. With the receiver and signal generator set to about 3.75MHz it should be possible to hear a tell tale hetrodyne. Once this is heard L_{18} should be adjusted to give maximum signal as measured on the S meter. Following this action L_3 and L_7 Fig. 6 should also be adjusted to maximize the response. Now it should be possible to try the receiver on an aerial where good readable signals should be heard. Remember to try changing the sideband switch if difficulty is experienced in resolving s.s.b. signals. It should also be possible to receive signals on the 20m band if S_1 is set appropriately, but of course L_4 and L_8 will have to be resonated using the signal generator.

Now the author suggests that the transmitter construction and alignment is started, but for the sake of clarity the alignment of the remaining premix bands will be outlined. The bandswitch should be set to one of the sections of 10m and if an r.f. measurement is made across R_{18} (Fig. 11) it should be possible to maximize the reading while adjusting the appropriate coil (L_{16} or L_{17}). It will be noticed that the output on either side of the coil adjustment peak will not be identical, i.e. the tuning curve is not symmetrical. The final adjustment should be set to about 0.25 down the gentle slopeside of the output curve.

There are many methods for adjusting the resonance of the balance mixer's output coils, but the most practical is to drive the receiver with an input of at least 50mV of 10m signal from a signal generator and adjust L_{21}, L_{22} for maximum overall receiver sensitivity. This technique reduces the possibility of peaking the circuits to any of the many spuriae present. Finally, adjust L_6 and L_{10} at an appropriate low capacitance point of the tuning capacitors C_{11} and C_{12} for maximum sensitivity. Potentiometer R_{22} is intended for use if breakthrough caused by the crystal oscillator frequency ± 9 MHz responses should prove troublesome. To adjust the 40 and 15m bands the previous procedure should be adopted by trimming the appropriate coils. To adjust the 9MHz i.f. breakthrough traps the signal generator should be connected to the aerial socket and set to 9MHz thus producing a detectable signal which can be nulled out by adjusting L_1 and C_2 consecutively (Fig. 6).

Independent receive/transmit control: The first action needed is to reconnect C_{16} (Fig. 11) thus bringing the variable capacitance diode back into circuit. Next set R_{21}, the i.r.t. control, to about mid-position and zero beat a steady carrier or the signal generator. Now, while switching the i.r.t. switch on and off, adjust R_{20} so that the zero beat occurs coincidently with the i.r.t. switch both on and off. With the present components R_{21} should give a frequency variation of about ± 3 kHz. If a greater deviation is required then one method of obtaining it would be to increase the value of C_{16}.

Linearity of the v.f.o.

It is desirable that the v.f.o. frequency change be linear with capacitor rotation. After careful consideration it was decided to use a two-ball-race capacitor of the straight line capacitance type. Jackson Brothers produce such a unit, type no. 101, with a range of 6 to 75pF. For linear frequency change the percentage CL ratio should not change, but of course it inevitably will. The large parallel value of C tends to mask this change and if small sections of the moving vanes are removed at the minimum C point the overall calibration can be brought to within 1kHz throughout the 500kHz range. This is aided by using only 80% of the available capacitance swing as the variable becomes somewhat non-linear near maximum capacitance. This concludes the alignment of the receiver and the local oscillator synthesizer.

As already stated the initial transmitter calibration should be carried out once the receiver is functioning on 80m and before completing the other ranges.

Transmitter

Microphone amplifier: The amplifier (consisting of $Tr_{5,7,8}$ Fig. 13) has more gain than is likely to be required. Potentiometer R_{32} is pre-set and should be set for the particular microphone in use and in the author's case is set to about one quarter of the way along the track from the h.t. end. The oscilloscope measurements across R_{37} with normal speech shows a peak-to-peak voltage of 200mV which in turn produces about 1.5V peak-to-peak across R_{18}. The tone generator supplies a somewhat distorted square-wave with a frequency of 1kHz to the emitter follower (Tr_5). There might be a slight operating advantage in increasing the value of R_{22} as the resistor used in the diagram does tend to make the tone injection rather stronger than the normal speech waveform. The tone is used extensively during tune-up and the alignment periods as there are no facilities for artificial carrier injection.

Balanced modulator: Before proceeding to the balanced modulator, the transmit/receive switching system must be tested.

With the press-to-talk switch in the transmit position and the R_{32} audio control set at its h.t. end r.f. measurements should be made at the collector of Tr_1. Adjust L_{27} for maximum reading and then transfer the

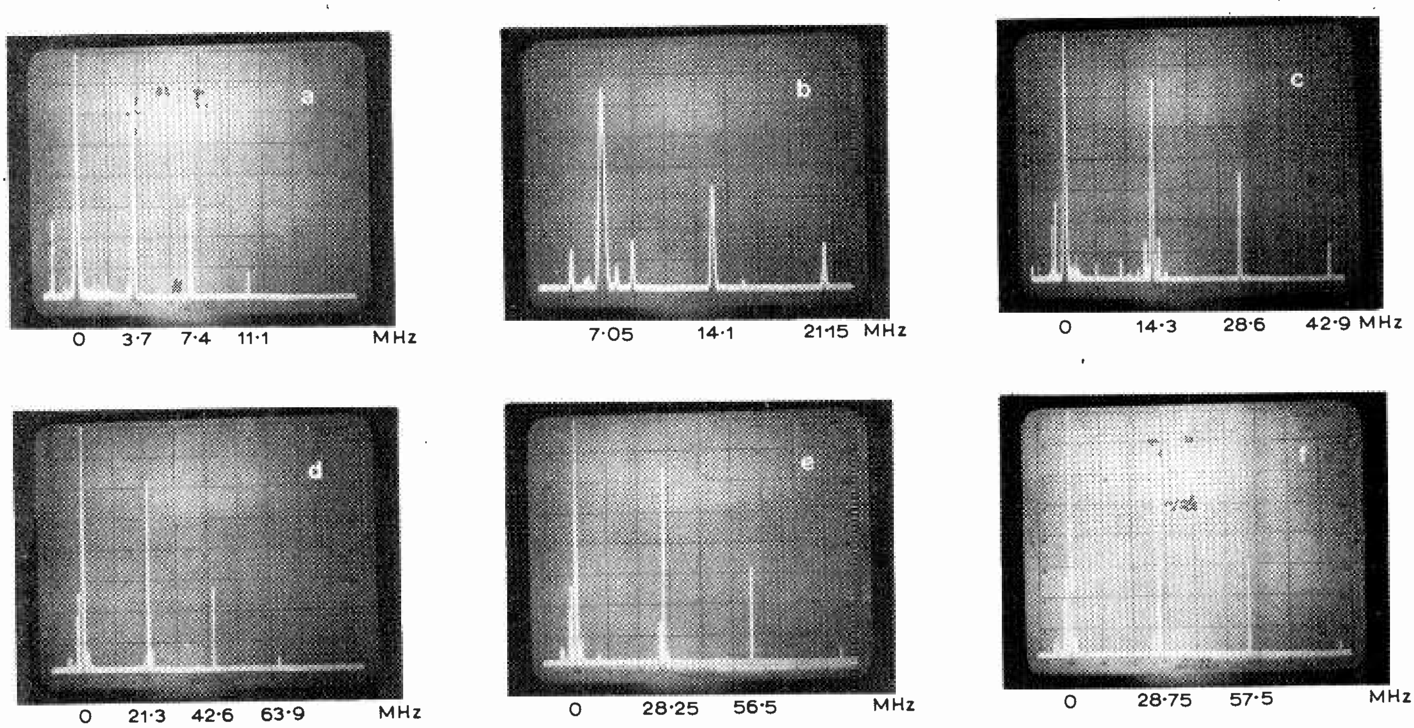


Fig. 17. Transmitter output using single-tone displayed on a Hewlett Packard spectrum analyzer fitted with an 8555A plug-in. (a) 80m band 45W output, (b) 40m band 45W output, (c) 20m band 35W output, (d) 15m band 30W output, (e) and (f) 10m band 27W output.

probe to one end of the secondary and once again resonate L_{27}. If the r.f. probe is now connected to the junction of D_1 and D_2 it should be possible to zero the r.f. present by alternately adjusting R_8 and C_7 remembering to try C_7 connected to the other side of the modulator and finalizing on that side which assists in reducing the carrier. Greater sensitivity while carrying out this adjustment can be had by transferring the probe to the collector of Tr_2. This adjustment having been completed, audio may be introduced by adjusting R_{32} to the point on its track that gave the correct audio previously. It should be possible to see the r.f. "kicking up" to about 800mV when the constructor speaks into the microphone.

Post filter i.f. amplifier: It is advisable to disable the automatic loading control of Tr_6 by temporarily shorting G_1 to earth (Fig. 13). The r.f. probe should now be positioned so as to measure the voltage appearing on the gate 1 electrode of either Tr_9 or Tr_{10}. With the tone generator switched on C_{27} should be adjusted for the maximum r.f. indicated at which point a maximum r.f. voltage of 100mV should be measured if the tone is replaced by ordinary speech via the microphone amplifier.

Transmitter balanced mixer: The alignment of the frequency synthesizer has already been dealt with and so the details concerning the transmitter balanced mixer should be straightforward. The r.f. probe may be connected across R_{42} and the various coils (L_{29}–L_{34}) can be resonated. These circuits should be adjusted for maximum r.f. indicated when the main transceiver tuning control is set to the mid-frequency of each amateur band. For example 40m should be set to 7.05MHz, as in the U.K. this particular band only covers from 7 to 7.1MHz.

Driver stage: Once again the alignment technique is quite straightforward. The r.f. probe is connected in this case to the grid (pin five of V_2) of the p.a. valve

Two notes are in order. If the r.f. probe uses a semiconductor diode care must be taken not to use excessive drive levels or the voltage rating of the diode may be exceeded. Also take heed of the warning note given earlier.

Each coil should be adjusted for maximum r.f. measured with the front end pretuned where the r.f. receive front end was optimized. In other words the tracking between the r.f. receive and transmit circuits must be correct. Each of the coils L_{35} to L_{39} should be resonated at the centre of the appropriate range.

The only out-of-the-ordinary adjustment is that of neutralization. The simplest way to adjust this is to connect the transmitter's output terminal to the input of an auxillary receiver or alternatively to a very sensitive r.f. indicator. The h.t. and screen supplies should be removed from V_1, the driver stage, and with the bandswitch set to 10m the transmitter should be carefully driven using the internal tone. The neutralizing capacitor is adjusted until a very sharp null is discerned in the signal received by the auxillary r.f. indicator. Capacitor C_{33} should be set to the centre of the null. *Take care to disconnect the auxillary receiver before reconnecting the supplies to the driver stage.*

Power output stage: The final stage in the alignment procedure is quite straightforward. The first move is to set the standing current of the p.a., with no drive, to 225mA. This is done by operating the press-to-talk button and adjusting R_{14} (Fig. 15) until the correct current is indicated. Swinging the pre-tune ganged capacitors should have no effect upon the magnitude of this current. This completed the only alignment left is to neutralize the stage and reintroduce the automatic loading control. The neutralizing is carried out in a similar manner to that of the driver stage. Remove both h.t. and screen supplies while adjusting C_{53} for a null in the leakage signal. Do take care in all work connected with the p.a. as the 750V h.t. could quite easily prove lethal. It is just possible that the constructor might find it necessary to reduce the value of C_{42} (Fig. 13) if the range of C_{53} proves to be limited.

To reintroduce the a.l.c. all that is required is to remove the temporarily connected earth on gate 1 of Tr_6 which was introduced earlier. With R_{51} (Fig. 13) adjusted to the R_{52} end, the a.l.c. is inoperative which, in the author's opinion, is the best position for it except under extremely adverse propagation conditions. As R_{51} is wound in the direction of R_{50} the a.l.c. control is brought into operation. The author suggests that if a linear amplifier is being used then its grid current should be monitored. The drive to the transceiver should be progressively increased until the grid current of the linear amplifier just flows on speech peaks. The a.l.c. should then be adjusted until no grid current spikes can be detected. This is the optimum setting for this control.

Correct p.a. output circuit operation

The correct adjustment of C_{51}, L_{43}, L_{44} and C_{50} is of paramount importance. These components make up the so-called pi tank circuit, the function of which is to transform the anode load impedance to that of the aerial impedance with as little loss as is feasible, while at the same time acting as a harmonic trap (i.e., a ringing circuit to regenerate the missing half cycle caused by the class AB1 biasing of the output stage). There is only one correct setting and, contrary to a widely held idea, it is not at all simple to find this setting. If the loading is too heavy then the Q of the filter becomes very low and also the power output drops and conversely if the loading is too light and the Q is high then there is a greater tendency for the p.a. to overload. This overloading will cause excessive harmonic and intermodulation distortion, excessive valve dissipation and of course a reduced output power.

The best method of setting the pi tank is to monitor the output voltage using an oscilloscope and to adjust for maximum undistorted output signal, using a dummy load of course. Some amateurs may not have access to an oscilloscope and therefore the following procedure should be adopted. Using the internal tone first at low level and progressively at higher levels the operator should monitor first the p.a. current and then the screen current. One word of warning, at high drive levels the on-period of the p.a. must be kept short or there is a real danger of destroying the 6146 valve. The correct setting of the tune and load controls is that which produces about 10mA of screen current for 120mA of anode current. If the screen current is too low then the loading is excessive and C_{50} must be decreased in value. Take care not to run the p.a. with excessive screen current as it is quite easy to exceed the screen dissipation rating. The maximum anode current can be taken from the characteristics, but whatever this indicates it is essential that no control grid current flows even on speech peaks. Assuming the procedure that has just been outlined is followed carefully the r.f. drive control should be advanced, while at the same time the control grid current is monitored. With plain speech the drive should be adjusted until not a flicker of current can be detected. This procedure will go a long way to ensuring a clean extremely readable signal.

Performance

Very considerable care was taken to evaluate the performance of the prototype transceiver and the results of these tests are outlined below.

Receive mode: The receiver exceeded the performance requirements outlined in Part 1. The image rejection varied between 60 and 80dB with the i.f. breakthrough rejection just meeting the required 60dB. The second channel rejection is in excess of 60dB and the a.g.c. performance also meets the 6dB output audio change for 80dB change of input. The linearity of the v.f.o. is well within $\pm$1kHz except at one end where the deviation approaches 3kHz.

Transmit mode: To assess the transmitter performance a Hewlett-Packard spectrum analyser was used for most of the tests, but photographs of only the interesting analyses are included here. The carrier suppression can be as great as 60dB below the peak signal output and can be guaranteed to remain 50dB down over many months even when the unit is transported.

Sideband suppression: This depends to a very great extent upon the particular filter used. In the author's case the suppression at 1kHz was always better than 40dB, probably much better, but as distortion products are rarely greater than 35dB down any greater sideband suppression is pointless.

The v.f.o. drift, after very careful adjustment, was repeatedly found to be less than 500Hz on switch on and once the unit was warm it became so difficult to measure that no figures are quoted here.

In view of the author's intention to use the transceiver as an exciter together with a 2m transverter extensive spuriae tests were carried out. The acceptable level of spuriae being much lower on 2m than on the h.f. bands. The photographs (Fig. 17) were taken without any special adjustments being made to the transceiver and are therefore representative of what the unit can be expected to produce. In most of the spectra the spuriae to the left and around zero frequency are a function of the spectrum analyser and can be ignored. It will become obvious that the harmonic suppression, in common with all pi tank output transmitters, is not adequate and it is essential that a television interference suppression low-pass filter is used between the transceiver and the aerial. Contrary to the generally held view a single pi tank cannot be expected to perform much better than the curves show. The second harmonic in all cases is about 30dB below the required output signal with all other spuriae being at least 50dB down. This is quite a respectable performance. The 40m spectrum is of considerable academic interest in that the v.f.o. frequency is visible below the 7.05MHz output signal and the next sizeable spuriae above being the 9MHz i.f.

Inter-modulation tests were conducted using a two-tone generator technique. One tone at 1kHz and the other at 2kHz. The graph (Fig. 18) shows the results which were as good as could be expected for a small power valve such as a 6146.

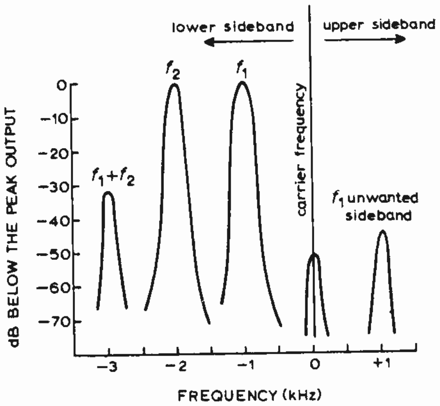


Fig. 18. Two-tone (1 and 2kHz) performance of the transmitter.

Power output measurements show a sharp decline in efficiency towards the h.f. end, but as the transceiver is intended to drive an outboard linear amplifier this is of little consequence. This drop in power is caused by the physical design of the p.a. compartment stray switch inductance, etc.

Little has been said concerning components. It is advisable to use brand new components everywhere and Solid State Modules, of 63, Woodhead Road, Solid, Lockwood, Huddersfield HD4 6ER, have indicated their intentions to stock most parts and in particular the crystal filter which at the time of writing costs £15 with carrier crystals £1.50 each. The h.f. crystals can be supplied by Senator Crystals, Dept. Q.C., 36, Valleyfield Road, London SW16 2HR, and the gate-protected f.e.ts by Henry's Radio Ltd., 303, Edgware Road, London W2.

It is hoped that many experienced amateurs will tackle the project and many more will find the articles useful reading.

Reference

1. D. R. Bowman, "100MHz Frequency Divider", *Wireless World*, August 1970, pp. 389-393.

Table of d.c. voltages

All voltages (r.m.s.) measured with reference to the earthed chassis unless otherwise stated. All gain controls set to minimum and the unit in the receive mode except where transmit voltages are tabulated. Most measurements were made using a Heathkit V-7A valve voltmeter and where required an r.f. probe. No aerial input signal was used.

low-voltage power supply (Fig. 16)

voltage across	C_5	16V
voltage across	C_6	−4V
voltage across	C_7	−3.2V

stabilized output nominally 11.5V (11 to 12V) with 20mV ripple when on receive and 40mV on transmit

emitter Tr_3	4V
zener output to v.f.o.	8.2V, 20mV ripple

high-voltage power unit (Fig. 15)

p.a. h.t. (no load)	840V, 2V ripple
p.a. h.t. (100mA load)	700V, 35V ripple
driver h.t.	340V, 7V ripple
p.a. screen supply	205V, 30mV ripple
p.a. bias supply	−50V, 800mV ripple (transmit)

receiver r.f. amplifier and mixer (Fig. 6)

	source	gate 1	gate 2	drain
Tr_1	4.7V	4.2V	5.4V	10.6V
Tr_2	290mV		800mV	10.6V
junction of R_2 and R_3			3.7V	

i.f. amplifier, a.g.c. amplifier and S meter circuit (Fig. 7)

	base	emitter	collector
Tr_2	2.6V	1.8V	3V
Tr_1	3.7V	—	9.6V
Tr_4	2.6V	1.8V	3.8V
Tr_3	4.4V	—	10V
Tr_6	2.6V	1.8V	6V
Tr_5	6.8V	—	10.2V
Tr_7	660mV	—	7.1V
Tr_8	620mV	—	100mV
junction of R_{23} and C_{19}			5.2V

b.f.o., detector and a.f. amplifier (Fig. 10)

	base	emitter	collector
Tr_1	2.3V	1.6V	7.8V*
Tr_2	1V	400mV	6.7V*
Tr_3	1V	400mV	6.7V*
Tr_4	1.9V	1.2V	5.8V
Tr_5	1.4V	700mV	5.4V
Tr_6	6.7V	6V	11.5V
Tr_7	—	—	0V
junction of R_{15} and R_{16}			6.05V

*no crystals fitted

local oscillator synthesizer (Fig. 11)

	base	emitter	collector
Tr_1	3V	2.2V	5.6V*
Tr_2	2.2V	1.5V	8V*
Tr_3	2.6V	2V	11.4V†
	gate 2	source	drain
Tr_4	500mV	100mV	10.5V
Tr_5	500mV	100mV	10.5V
junction of R_5 and R_3		6.5V	
junction of R_5 and R_8		8V	

*coil short circuited
†base connected to earth via 10nF

transmitter (Fig. 13)

	base	emitter	collector
Tr_7	1.2V	600mV	5.9V
Tr_5	2.2V	1.6V	10.3V
Tr_{3-4}	no relevant measurements		
Tr_1	2.2V	2.9V	10.8V
Tr_2	1.4V	2V	5.4V
	gate 2	source	drain
Tr_8	—	1V	7.6V
Tr_6	4V	700mV	11.6V
Tr_9	600mV	800mV	11.3V
Tr_{10}	600mV	600mV	11.3V

V_1 cathode 15V, screen 335V, anode 350V (cathode gain control at maximum resistance).
V_2 as per high-voltage power supply

Table of r.f. voltages

Tr_1 collector (Fig. 10)	1.5V
Tr_7 base (Fig. 7)	100mV with 50mV aerial input
Tr_2 base (Fig. 11)	0.4 to 6.8V varies between
Tr_4 G1 (Fig. 11)	0.7 to 0.35V these limits as
Tr_5 G1 (Fig. 11)	0.7 to 0.35V C_1 is rotated
Tr_4 G2 (Fig. 11)	2.5 to 4V changes with h.f. crystal selected
Tr_5 drain (Fig. 11)	3 to 5V
L_{27} (Fig. 13), both sides	800mV
Tr_2 collector (Fig. 13)	200mV peaking to this value when driven by normal speech
Tr_9 G1 (Fig. 13), Tr_{10} G1 (Fig. 13)	200mV (synthesizer disabled)
V_1 grid 1, pin 2 (Fig. 13)	1 2V

Power output

band	power	efficiency
80m	45W	60%
40m	42W	57%
20m	38W	51%
15m	36W	49%
10m	34W	46%

Components list

High voltage power supply (Fig. 15)

C_1	100μ, 450V, E	C_4	50μ, 350V, E
C_2	100μ, 450V, E	C_5	10n, 350V, DC
C_3	32μ, 350V, E		

E—electrolytic DC—disc ceramic

R_1	47	R_9	10k, 10W, WW
R_2	680, 5W, WW	R_{10}	100k
R_3	100k, 2W	R_{11}	136*
R_4	100k, 2W	R_{12}	9.2k
R_5	3.9M	R_{13}	10k
R_6	3.9M	R_{14}	10k, WW, P
R_7	2.2M	R_{15}	20k
R_8	1M		

WW—wire wound P—potentiometer
*Adjust so that meter reads 1mA f.s.d. (grid current)

D_1 to D_7	800 p.i.v. rectifier diodes (BY100, REC51A, etc.)
V_1 and V_2	OB2 105V stabilizer

Two pole make on/off switch
Transformer: primary standard mains: secondary 300–0–300V, 120mA; 6.3V, 3A; 6.3V, 1A.
2A fuse

Low-voltage power supply (Fig. 16)

C_1	10n, 350V, DC	C_5	1,000μ, 25V, E
C_2	10n, 350V, DC	C_6	100n, DC
C_3	10n, 350V, DC	C_7	100n, DC
C_4	10n, 350V, DC	C_8	100n, DC

E—electrolytic D3—disc ceramic

R_1	1	R_5	2.2k
R_2	560	R_6	100
R_3	3.3k	R_7	2k, P
R_4	470		

P—potentiometer

Tr_1	BD123 or 2N3055 or 2N3054
Tr_2	BFY50 or 2N696
Tr_3 and Tr_4	2N3702 or BCY72
D_1 to D_4	50 p.i.v., 250mA rectifier diodes
D_5 and D_6	8V zener diodes, BZY88 8.2, ZEP 8.2

Circuit Ideas

If the circuit is made up on a printed circuit board, surface leakage between the supply rails and the "high" input terminal may give rise to large standing errors. This can be avoided by surrounding the track between the "high" input terminal and pin 3 of the 709 with a guardstrip which should be connected to the 0-V rail.

Pin numbers are for TO-99 or 8-lead in-line packages. The 11-Ω resistor from pin 2 and the 1-kΩ input resistor should both be 2% tolerance.

H. MacDonald,
Ashford,
Middx.

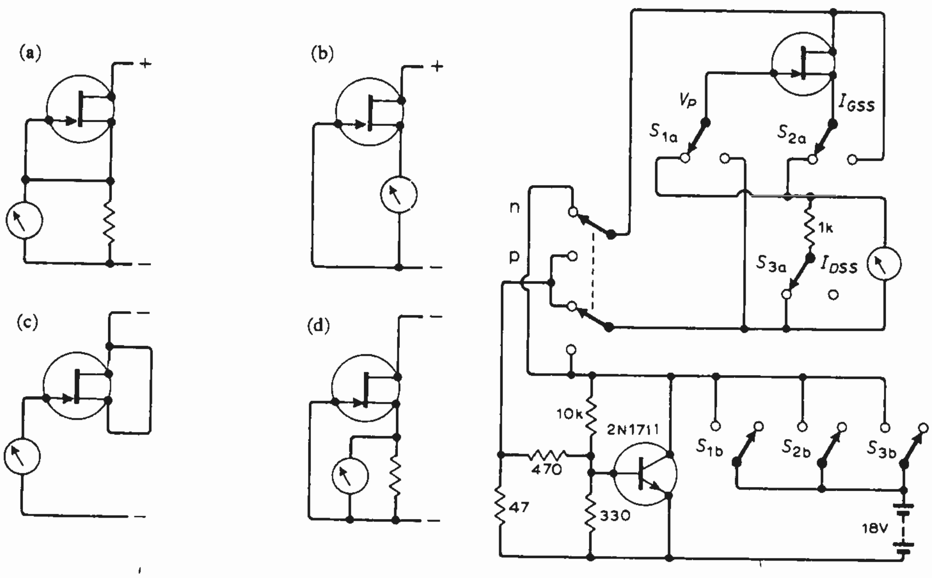
Simple tester for f.e.ts

Zero-bias drain current, gate cut-off voltage and gate leakage current in junction f.e.ts can be measured with a high-impedance voltmeter (20,000Ω/V or better) using the circuits shown in figures (a), (b) and (c) respectively. Using an external voltmeter, a simple tester can be built around three push-button switches as shown (right). With the source resistor value of 1kΩ, I_{DDS} is read in mA on the voltage scales of the meter. Two small 9-V batteries give sufficient voltage to test most f.e.ts. The f.e.t. is protected by the protection circuit (2N1711) which works on the loadline principle, and limits the power to the test circuit to about 200mW.

The circuit has an additional refinement. By depressing the V_P button and the I_{DSS} button simultaneously, the circuit shown in (d) is made. The meter reading now gives both drain current in mA and gate-source voltage in volts, thus giving a third point on the $I_D V_{GS}$ characteristic.

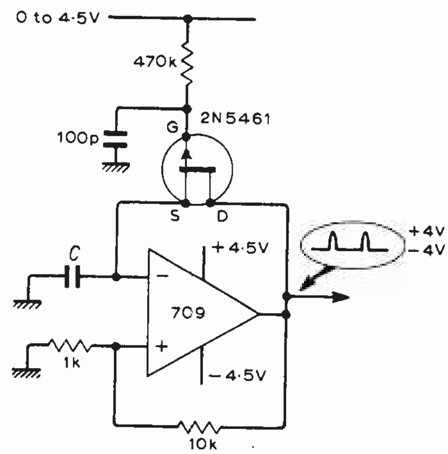
The complete tester can be built in a box measuring 10 × 7 × 5cm.

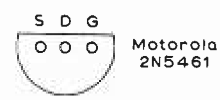
Jostein Skjelstad,
Ålesund,
Norway.



Voltage-controlled filter and oscillator

In the item under the title "Voltage-controlled high-pass filter" in July Circuit Ideas (page 349) the first paragraph, referring to a voltage-controlled oscillator, was, we regret, included without the accompanying diagram. The simple v.c.o. circuit, shown below, uses a 709 op-amp and a junction f.e.t. to give a logarithmic frequency-voltage relationship over five octaves, and a pulsed output. Capacitor C is chosen to give the required frequency range.

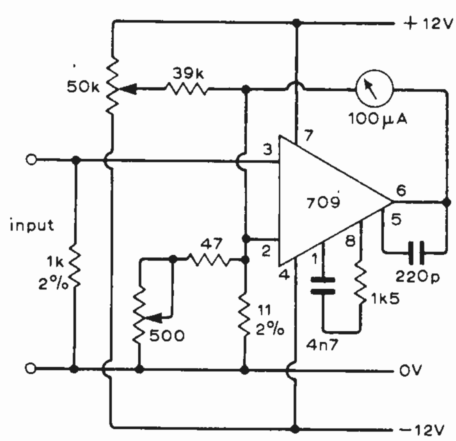




The voltage-controlled high-pass filter circuit (July issue) used two f.e.ts and these should be matched. This can be done by ensuring the voltage across a 1-kΩ source resistor — with the gate earthed and the drain at −9V — is within 10%, preferably 2%. A 100-kΩ resistor was omitted from the circuit, between the gate end of the 1MΩ resistor and potentiometer wiper. Circuits contributed by David G. Malham, University College, Cardiff.

Op-amp meter amplifier

This circuit uses a 709 operational amplifier to increase the sensitivity of a moving-coil meter. With the values given, full-scale deflection of the meter is obtained for a 1μA input. Different ranges can be obtained by altering the values of the resistive network connected between pin 2 of the 709 and the 0-V rail. Accuracy and linearity of the circuit is within 1% over the ambient range of 0 to 50°C. There is no significant warm-up drift. By adjusting the 50-kΩ potentiometer the circuit can also be used to give a 500-0-500pA range. The 500Ω potentiometer is used to adjust the amplifier gain.



In contributing to Circuit Ideas readers should say how their circuit is an improvement on previous ones, preferably in the first sentence. We will pay £5 for each idea published.

About People

Eric Eastwood, C.B.E., F.R.S., director of research of the General Electric Company and chief scientist of the Marconi Company, takes office as president of the Institution of Electrical Engineers on 1 October. Dr. Eastwood, who was born in 1910, read physics at Manchester University and, after graduation, began research in spectroscopy. He continued this work at Cambridge University and received his Ph.D. in 1935. During the war he served in the R.A.F. on the technical problems associated with the use of radar by the Fighter Defences. After the war, Dr. Eastwood joined the Nelson Research Laboratory of the English Electric Company. When English Electric acquired the Marconi Company, Dr. Eastwood was transferred to the Marconi Research Laboratory, Great Baddow. He was appointed director of research to the Marconi Company in 1954 and became research director of the English Electric Group in 1962. Dr. Eastwood has been greatly interested in the application of radar as a research tool in atmospheric studies. With the aid of the Marconi Experimental Radar Station at Bushy Hill, Essex, he has been able to apply radar methods, not only to the study of tropospheric propagation and the aurora, but also to carrying out comprehensive ornithological investigations on bird migration.

Air-Vice Marshal Sir Victor Tait, K.B.E., C.B.E., B.Sc., deputy chairman of Ultra Electronic Holdings, has retired from the board and also from that of its subsidiary Ultra Electronics Ltd. Sir Victor has been associated with Ultra Electronics either as chairman or deputy chairman for 17 years. Between 1917 and 1945 Sir Victor served in the Royal Flying Corps and the R.A.F. and was director of radar and director-general of signals at the Air Ministry between 1942 and 1945. He was operations director of B.O.A.C. from 1945 to 1956 and chairman of International Aeradio from 1946 to 1963.

F. J. Brooks, newly appointed commercial director of Racal-Amplivox Communications Ltd, of Wembley, was formerly manager of the marketing department at Racal Communication Ltd, in Bracknell. Mr. Brooks joined Racal in 1960 as a sales engineer and has successively held the posts of assistant sales manager (U.K.) and marketing manager (Europe). Racal acquired Amplivox, manufacturers of headsets and microphones, a few months ago.

J. A. D. Timms has been appointed as assistant managing director of British Communications Corporation Ltd, a company in the Racal Electronics Group. He retains his responsibilities as director of supplies and services. Mr. Timms has been with the company since its inception in 1946.

J. Treeby Dickinson has retired from the post of chief engineer in the Technical Centre of the European Broadcasting Union, but his services are being retained by the E.B.U. on a part-time basis for the time being. Mr. Dickinson has held the position for twenty-two years, a period covering the setting up of the E.B.U. Technical Centre in Brussels, the establishment of its Receiving and Measuring Station at Jurbise-Masnuy and the introduction and expansion of Eurovision. He joined the B.B.C. in 1937 and has been on secondment to the E.B.U. since 1950. Prior to joining the B.B.C. he was in industry engaged on receiver development.

ITT Components Group Europe, has appointed **Philip Allin** as director components for the United Kingdom. Mr. Allin, 41, was previously general manager of the Valve Division of ITT Components Group at Paignton, Devon. Educated at Cambridge where he took the mechanical sciences tripos, he entered industry as a research physicist, joining ITT in 1965. He became manager of the Valve Division in 1968.

David Farrar, Dip. Tech., M. Inst. P., who joined English Electric Valve Company in 1959 and obtained a diploma of technology in applied physics after completing his apprenticeship, has been appointed a sales engineer for TV camera tubes. He has worked in the camera tube research department at Chelmsford, investigating, among other things, the factors controlling the vacuum in a variety of photo-tubes.

Peter Buck, M.I.E.R.E., who has been managing director of Westrex, London, has been appointed vice-president of the Westrex division of Litton Industries and will direct all Westrex activities outside the United States. Since joining Westrex in 1946 as a recording engineer, Mr. Buck has served as general manager and director for several regional offices in the Far East, Australia and Europe.

Consumer Microcircuits, now in Witham, Essex, have appointed **N. A. B. King** and **G. J. Bates** as design engineers. Mr. King was with Marconi, where he was directly concerned with the first m.o.s. designs, and latterly with G.E.C. Mr. Bates was previously with G.E.C. where he spent one year as a photolithographic engineer, followed by a period as m.o.s. integrated circuit engineer.

Orbit Controls Ltd., the Cheltenham industrial electronics engineers, have announced the appointment of **C. P. Crompton,** M.I.E.R.E., as a director Mr. Crompton joined the company as a senior project engineer on its formation in July 1968 and was appointed chief engineer in 1970. He will continue to be responsible for design and development. Prior to joining Orbit, Mr. Crompton, who is 44, held senior engineering posts with Advance Electronics Ltd., Racal Communications Ltd. and the Sperry Gyroscope Co.

Peter Turner, who is 33, and joined Hewlett-Packard as an instrument and component sales engineer in 1969, has been appointed manager, components group. His appointment coincides with the announcement that Celdis Ltd., had been appointed sole distributors for Hewlett-Packard's range of opto-electronic devices.

Celdis Ltd have appointed **John Eeles** as product manager for the H-P range of opto-electronic products. Mr. Eeles, who is 25, served an apprenticeship with Racal Communications, and subsequently worked on mobile transmitter maintenance with the B.B.C. before joining Celdis as a sales engineer two years ago.

Alan Fletcher, works manager of Devon Transformers Ltd, of Penzance, Cornwall, for the past three years, has been appointed sales manager of the company and of the associated company, Ferro-Mag (Electrical) Ltd.

J. G. O'Keeffe, M.Sc., assistant director of engineering with Radio Telefis Eireann since 1963, has retired but is being retained as a consultant and he will continue to serve as chairman of the European Broadcasting Union's technical committee on v.h.f. sound broadcasting. Mr. O'Keeffe joined Eire's Department of Posts & Telegraphs as an assistant engineer in 1932 and later became engineer-in-charge, Athlone transmitter. He transferred to Radio Eireann headquarters and in 1954 became its chief engineer. He is succeeded as assistant director of engineering by **E. J. Slowey,** B.Sc.(Eng.), F.I.E.E., who, since 1966 has been manager, central engineering services, television. Mr. Slowey graduated in electrical engineering from Queen's University, Dublin, in 1935. He joined the radio and broadcasting section of the Department of Posts and Telegraphs in 1936 and by 1940 was engineer-in-charge of the Radio Eireann studios.

John F. Willsher has been appointed managing director of a newly-formed EMI subsidiary, EMI Sound & Vision Equipment Ltd., Hayes, Middlesex, established to concentrate under a single management the diverse technical activities of EMI in sound and vision systems and related fields. The new company co-ordinates the activities of four divisions — EMI Pathe Equipment Division of which **G.C. McLean** is director; EMI Television Equipment Division, **J. D. Tucker,** director; EMI Industrial Components; and EMI Service, the central technical equipment maintenance and service organization, with **A. Hunt** as director. Mr. Willsher joined EMI in January. He was previously managing director of Berkey Technical (UK) Ltd., a specialist television, film and theatre equipment company, which he joined in 1969. For over 20 years previously, he held various appointments within the George Kent group of companies, latterly as managing director of international operations.

Norman Weisbloom has become technical manager of Mullard's Government Electronics Division. Mr. Weisbloom joined Mullard in 1953 and his career since then has been mainly in semiconductor management. Latterly, he has been commercial product manager for integrated circuits.

David J. I. Gray, F.I.Mech.E., has joined the board of Goldring Manufacturing Company. Mr. Gray, who is 47, joins Goldring from Pye where he was latterly managing director of Pye TVT and previously director of Pye Business Communications. Prior to joining Pye in 1968 he was managing director of George W. King, of Stevenage.

Radio Reception Developments

New techniques for communications systems described at I.E.R.E. conference

Technical development in radio receivers has been extremely slow, compared with that in, say, electronic computers and control systems. The superheterodyne was invented in 1919 and is still the basis of the majority of receivers; likewise, such techniques as synchronous detection, single-sideband operation multiplexing and diversity reception, which are periodically brought out of the cupboard, dusted and re-applied, have been known for many decades. The only technical changes which have broken new ground are not in basic systems at all but those which have been forced or made desirable by the demand for more channels, i.e. the gradual move into higher and higher frequency bands, and the availability of new active devices – the replacement of the valve by the transistor and the transistor by the integrated circuit. One reason for this slowness of development is that, if one includes sound and television broadcasting, for every one transmitter there are thousands of receivers – an enormous, widely spread investment which places an economic restraint on changes of transmission systems. If the system must remain in a particular form for many years there is obviously not much scope for the receiver designer.

In radio telephones and radio communication systems, however, considered as a separate class, there is much less of a disparity between the numbers of receivers and transmitters, and consequently any change of transmission system – for example, between types of modulation in mobile radio – is not such a drastic matter. It was perhaps for that reason that a recent conference on Radio Receivers and Associated Systems, organized by the I.E.R.E. in association with the I.E.E. and I.E.E.E. at the University College of Swansea (4th-6th July), was largely devoted to developments in radio communication — as distinct from, say, broadcasting or radio navigation. Thirty-three papers were presented, in five sessions, and the following notes are a selection from these.

Diversity reception for mobile radio. Birmingham University's Department of Electronic and Electrical Engineering has been working on spatial diversity reception systems for aircraft and land vehicles for some years. At the conference five of its members presented a group of papers on automatic methods of compensating for the rapid fading (hundreds of fades per second, upwards) due to multipath reception which occurs in v.h.f. mobile radio links. Propagation at v.h.f. is principally by way of scattering, and the multiple reflection and diffraction from buildings and other obstacles cause standing-wave patterns to be set up, through which the mobile receiver moves. Amplitude variations of typically 20dB are experienced at the receiver (see Fig. 2). Since the distance between adjacent maxima and minima in the standing-wave pattern lies in the region of 0.5m to 1m it becomes feasible to mount an array of spaced aerials on the vehicle and combine additively the separate signals in such a way that the fading in the summed signal will be diminished.

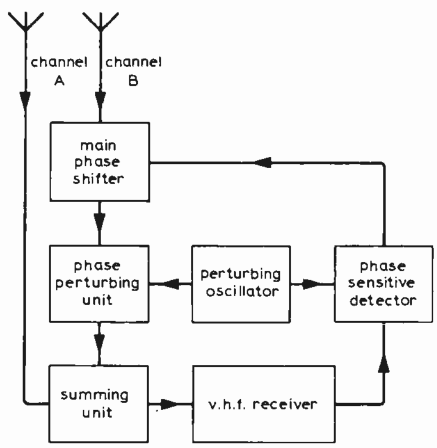


Fig. 1. Basic principle of self-phasing aerial system for space diversity reception in mobile radio. (All three channels are not shown.)

The Birmingham schemes are based on automatic adjustment of the phases of the signals from the separate aerials; and one paper, for example, by J. D. Parsons and P. A. Ratcliff, described a three-channel "adaptive" system, using this principle, which was built around a conventional v.h.f./a.m. receiver. Fig.1 shows the basic components of the method, not including, for simplicity, all three channels. One aerial signal (channel A) is considered as a reference signal and is connected directly to a summing unit. The other (channel B) is passed through a phase-shifter capable of changing the signal phase through 2π

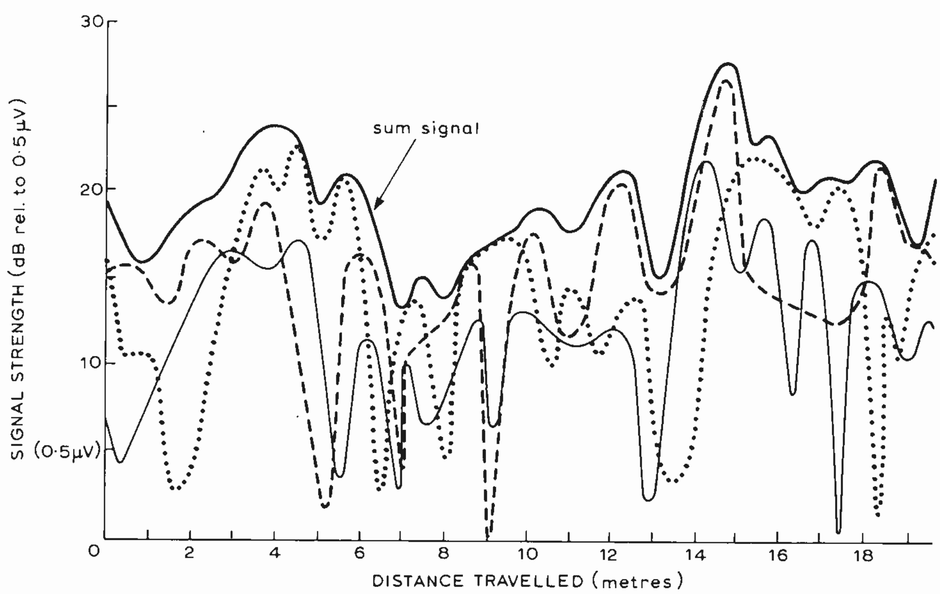


Fig. 2. Recordings of signal amplitudes in three aerials mounted at 0.4 wavelength spacing on a van, travelling through central Birmingham. The sum signal is given by the three-channel self-phasing aerial system shown schematically in Fig. 1.

radians. The phase-shifter output is passed to a phase-perturbing unit, a small-deviation phase modulator, and provides phase-difference information which is used to set the main phase shifter. The phase-perturbed signal then enters the summing unit, the output of which feeds the receiver.

Amplitude modulation at the perturbing frequency is produced in the sum signal. This modulation is in phase with the perturbing frequency when signal B lags signal A and the sum signal, and in anti-phase when signal B leads signal A and the sum signal. When signal B is in phase with signal A only a small amplitude second harmonic of the perturbing frequency is produced. An a.m. error signal is extracted from the detector of the receiver and phase detected with respect to the phase-perturbing oscillator. The polarity of the filtered d.c. output thus obtained indicates the direction in which the main-phase shifter must be changed to co-phase the signals A and B. This basic method is extended to three channels, a separate control loop being provided for each aerial and a separate perturbing frequency for each channel.

Results obtained in a severe "multi-path" area in the centre of Birmingham are shown in Fig.2. The individual aerial signal strengths are plotted in broken and thin lines, with variations as high as 27dB, but the maximum variation of the sum signal is only 15dB and its fading rate is greatly reduced.

Another diversity system includes a new technique for phase measurement. Described by J. D. Parsons and M. Henze, the system has not yet been tried out but a prototype is being built for use in the 450-470MHz mobile radio band. The phase measuring technique, based on single-sideband modulation with carrier, is used to detect the phase-difference between each of the received signals and their sum; and control signals are generated which adjust quantized phase-shifters in such a way that the signals are put approximately in phase. The system includes a single, standard radio receiver capable of detecting either amplitude-modulated or frequency-modulated signals. There is no need for modification to the receiver circuitry.

The third diversity system from Birmingham University, described by M. J. Withers, is based on a principle similar to that of a technique used already in radar. It utilizes a phase shifter which is swept continuously and therefore causes two independently received signals to come into phase periodically. By making the periodicity short compared with the modulation it is possible to satisfy the sampling theorem of information theory and extract the original modulation. A phase modulator, in the signal path from one aerial, repeatedly changes the signal phase in steps of $\pi/4$ from 0 to 2π. The combined signals are fed into a conventional v.h.f./a.m. receiver and the original modulation is retrieved by an envelope detector. Mobile tests have not yet been made but Mr. Withers said that an "on-the-air" trial at a fixed location showed no degradation of speech quality. If the system is successful in mobile tests its main advantage will be that a basic area coverage receiver can be used with a minimum of modification and the addition of an extra aerial, a combining "hybrid" and a phase modulator.

Selective calling for mobile radio. Moving into higher and higher frequency bands is not the only way of meeting the demand for extra communications channels: there is the reduction of channel spacing (e.g. in mobile radio from 50kHz to 25kHz and now to 12.5kHz) and the sharing of particular channels. The last expedient is now commonplace in mobile radio, and though simple to achieve, it has the serious practical disadvantage of causing any one "mobile" to listen to an almost continuous stream of speech, only a small proportion of which is directed at him. This problem has been investigated over the past few years by the introduction of various systems of selective calling. In these, by the use of signalling codes, a particular "mobile" can be called while the remainder are kept muted. There are single-tone and multi-tone systems, and the multi-tone systems can be subdivided into simultaneous and sequential systems. J. E. Philips and E. J. Slevin, of G.E.C. Mobile Radio (part of Marconi Communications Systems Ltd), described a five-tone sequential system which they claimed "would appear to provide the optimum solution for present requirements" — optimum, that is, in its balance of price and calling capacity.

The tones available — chosen from a C.C.I.R. marine specification so that the system can be used for ships – are ten in number, between 1124 and 2100Hz. To call a "mobile" from the base station any five of these tones are transmitted in sequence, so that in all 10^5 tone combinations, or codes, are available. The duration of each tone, determined by the hardware consideration of the Q of filter coils in a decoder at the receiving end, is 40 milliseconds.

In the mobile station, the audio signal from the receiver is fed via the decoder to the loudspeaker. The h.t. for the decoder is taken from the transmitter/receiver supplies and so no internal connections to the set need be made. Since the set is disconnected from the speaker except when the "mobile" is called no speech and/or noise is heard. Thus the required setting of the volume control is not known. To overcome this an h.t. switch has been included which connects the decoder to 9V and 5V supply rails when the incoming noise and/or signal level is above a predetermined threshold, which may be reached by adjustment of the volume control. The 5V rail is then used to light a lamp on the front panel, indicating to the operator that the volume control of the receiver is adjusted correctly.

The first tone received merely operates the h.t. switch. The second tone, forming the first of the actual calling code, is fed to a limiter which enables the decoder to work over a large dynamic range (this may be as much as 25dB since at low s/n levels the receiver's a.g.c. system is inoperative). From the limiter the tone signal is fed to a tuned circuit (filter), the frequency of which is predetermined by the particular code of the "mobile". The code to which each decoder responds is determined by a wire matrix which selects various tappings on a coil to give the various tuning frequencies. These tappings are selected in the required order by sequencing logic. The first code tone, on entering the filter, will cause the tuned circuit of the filter to resonate because the circuit is programmed to tune this frequency. Once the filter response has built up above a certain level an integrator begins to operate and its output, upon reaching the trigger level of a Schmitt trigger, will actuate it. This Schmitt trigger then operates a sequencing network which selects the next tap of the coil – determined by the wire matrix code – which causes the tuned circuit to tune to the second code tone, and so on, until all five tones have been decoded. After the last tone has been decoded a relay is actuated; and this causes a lamp to be lit on the front of the decoder unit and connects the audio signal from the receiver to the loudspeaker.

The authors demonstrated a selective calling equipment working on this principle and stated that the decoder to be carried by the "mobile" would eventually be manufactured as a 2in × 1in thick-film circuit.

Improving s/n in f.m. demodulation. In a communications f.m. receiver, when the noise accompanying the carrier at the receiver input is large enough for the so-called threshold region* to be entered, the envelope of the carrier-plus-noise combination shows a significant number of dips. These are experienced as sharp noise peaks, heard as clicks, in the receiver output. J. H. Roberts, of Plessey Avionics and Communications, described several

"Threshold" is defined as the point, on the graph of input carrier/noise ratio vs. output signal/noise ratio, at which the output signal/noise ratio departs by 1dB from its linear relationship with the input carrier/noise ratio.

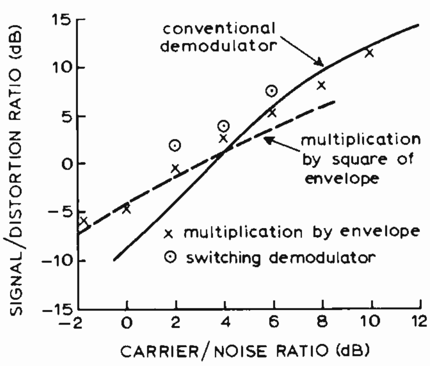


Fig. 3. Reduction of click noise in f.m. reception by multiplying the base-band output by a function of the i.f. output. The graphs of signal/distortion ratio vs. carrier/noise ratio are for different multiplying functions.

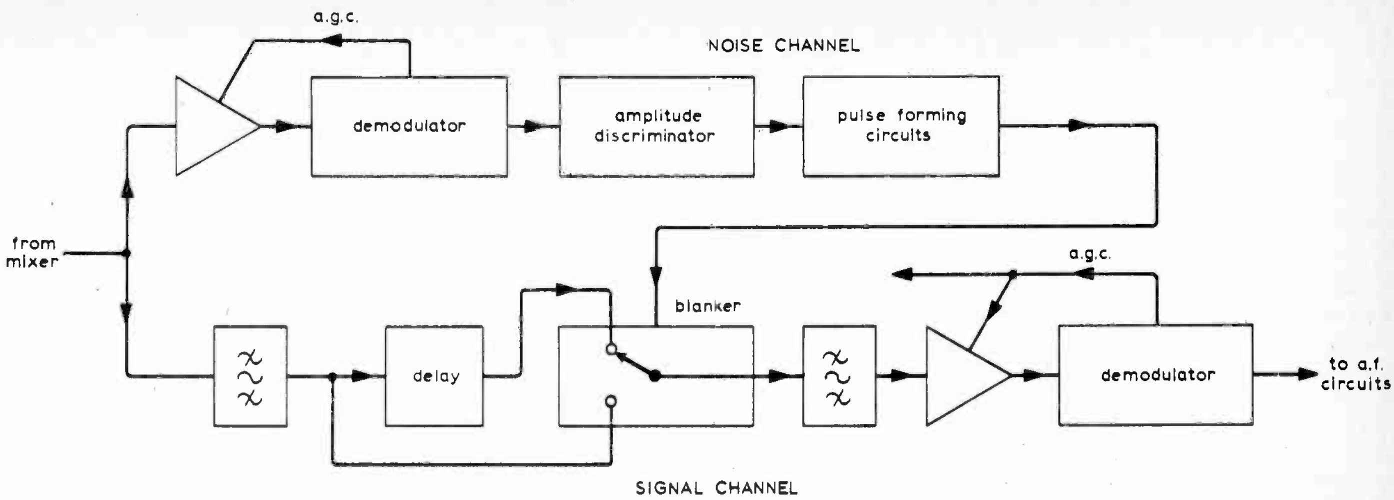


Fig. 4. Impulse noise blanker fed from mixer of a.m. receiver. In one channel each noise impulse initiates a blanking pulse which is applied to the "signal" channel to reduce its gain at the appropriate time.

methods by which this effect could be mitigated, all based on the idea of multiplying the base-band output of the f.m. receiver by a function of the envelope of the i.f. input. This has the effect of reducing the intensity of the click noise. In fact the click noise is exchanged for other types of base-band interference, but an overall improvement in signal/distortion ratio is said to be obtained at carrier/noise ratios well below the threshold point in the receiver. The practical situation in which Mr. Roberts was interested was with low carrier/noise ratios of −2dB to +6dB.

The envelope multiplication was tried out by means of computer simulation and some of the results are presented in Fig 3, which is for the top end (1MHz) of a 240-channel f.m. radio system with a base-band range of 60-1052kHz. Generally, there is improvement in signal/distortion ratio at low values of carrier/noise ratio but degradation at high values. Multiplication by the envelope function seems to be superior to multiplication by the square of the envelope. The best performance (circled dots) is obtained by what is described as a switching demodulator. In this there is a continual switching between conventional demodulation (solid line) and multiplication by the square of the envelope (broken line) – each switch to multiplication being made when the instantaneous level of the envelope falls below a specified threshold and a noise click is likely to occur.

Reducing impulse noise. Much of the noise which affects a.m. radio communication is impulsive in character, notably that generated by the ignition systems of motor vehicles. One method of countering this is by the use of an amplitude limiter, or "clipper", but an alternative approach at present being investigated is the use of a "blanker" – a circuit which reduces the receiver gain to a very low value for the duration of the noise pulse. As an example, Professor W. Gosling described the work that is being done at the University College of Swansea on a.m. and s.s.b. receivers, particularly that of analysing the process of blanking and deciding on the best blanking function (pulse shape) to use. The basic technique is to take the signal from the receiver's mixer and split it into two channels (Fig. 4), a "noise" channel, in which the noise impulses are detected and used to initiate blanking action, and a "signal" channel. The signal channel is conventional except that it includes a blanking circuit and also that a delay line is inserted to ensure that the noise pulse is given sufficient time to bring the blanking circuit into operation *before* the noise pulse arrives at it.

A trapezoidal blanking function has been chosen, in preference to a rectangular function, for a variety of reasons; for example, the Fourier components, which may fall within the passband of the receiver, decline more rapidly, and there is less drastic "ringing" of the i.f. filter which follows the blanking circuit. Trapezoidal blanking almost eliminates high-frequency ringing (about 3kHz), but a lower frequency ring remains, and will typically worsen the s/n ratio by 3-4dB.

Fundamentally this difficulty arises because the narrow band i.f. filter through which the received signal must pass cannot be designed both to give the required adjacent channel selectivity and also to have a good response to transients, whether noise pulses or blanked breaks in the received signal. The blanker reduces the amplitude of the transient to equal the instantaneous amplitude of the received signal, and this is its advantage over a receiver not equipped with noise reduction circuits, but a serious transient still remains. A possible approach to this problem is to reduce the magnitude of the blanking transient, for example by holding the signal amplitude constant during the blanking period rather than reducing it to zero. Since blanking operations must be performed at i.f., and at a signal level of only tens of microvolts, circuits which will perform this function adequately present serious difficulties.

One solution under investigation at Swansea is shown in Fig. 4. Here the i.f. is 5.2MHz, the pre-blanker i.f. filter has a bandwidth of 100kHz while the post-blanker filter is a conventional s.s.b. crystal i.f. filter with a passband of 2.5kHz. The necessary delay is provided by a 64μs ultrasonic glass delay line. The received signal normally passes through the broad-band delay line, but when a noise pulse is detected the delay line is switched out of circuit and the signal passes directly to the subsequent receiver circuits. Provided the delay is short compared with the shortest modulation period, there will be little change in the signal amplitude; hence the transient superimposed on the incoming signal will be much reduced.

Automated frequency changes for h.f. Readers who follow the regular feature H.F. Predictions in this journal will be well aware that the operating frequencies of h.f. radio communications systems have to be changed throughout the day in order to take advantage of the changing ionospheric propagation conditions along different routes. This normally entails a good deal of repetitive and time consuming manual operation, but the Post Office radio telegraph station at Somerton, Somerset, has introduced an automated system for changing the frequencies of receivers which relieves staff of this work, reduces human errors and allows the operator to concentrate on fault localization and clearance. The engineer in charge of the station, Mr. L. Wilks, described the system.

The basic method adopted is to use the distant-end transmitter to control the receiver frequency changes within a basic schedule provided by a 60-position drum-type timer making one revolution in 24 hours. This timing schedule is set by cams inserted in the drum and these operate switches which are arranged to give a "time slot" of 48 minutes during which a frequency change to a receiver can be made. To allow for transmitter and/or receiver frequency variations, the receiver performs a frequency searching operation (±500Hz at a speed of 40Hz per second) after making a frequency change. The receiver reverts to its normal a.f.c. locked condition when a signal of correct keying speed and acceptable telegraph distortion is recognized.

World of Amateur Radio

Moonbounce and Moon reception

Moonbounce (earth-moon-earth) contacts by amateurs have in recent years been reported on the 144, 220 (U.S.A.), 420, 1215 and 2300-MHz bands. These contacts include Sweden (SM7BAE) to New Zealand (ZL1AZR) on 144 MHz; California (WA6HXW) to Australia (VK2AMW) on 420MHz; California (WB6IOM) to U.K. (Peter Blair, G3LTF) on 1215MHz; and an all-American contact (W3GKP to W4HHK) on 2300MHz.

But a novel use of 2300MHz moonbounce equipment has been reported by Paul Wilson, W4HHK, and Richard Knadle, K2RIW. This has been the direct pick-up of Apollo mission S-band communications, culminating in the successful reception of voice transmissions from the Apollo 16 command service module and the 1-watt (2276-2279MHz) transmitters left on the moon's surface. Under F.C.C. regulations there is nothing to prevent American amateurs from attempting to receive such transmissions provided they do not disclose the contents.

The equipment used includes a commerically built 18ft dish aerial at W4HHK and a 12ft home-made dish at K2RIW; at both stations low-noise parametric amplifiers are used.

Most of us tend to be satisfied with rather less direct contact with the Apollo missions — but recently I received from the Goddard Amateur Radio Club a QSL card for a contact with WG3SFC which operated during the Apollo 16 flight from the NASA Goddard Space Flight Center. The operation of such "special event" stations has been proving very popular in the United States recently, with special prefixes allotted — so much so that the F.C.C. has now appealed to American amateurs to cut back such requests unless the event is of real public interest.

A French high-altitude balloon (N5) to be launched on September 17th, and expected to reach a height of some 28km, will carry an amateur radio transposer accepting signals on 432.1-432.2MHz and retransmitting them between 145.6-145.9MHz.

Latest estimate for the launch of Oscar 6 is November. The U.K. Ministry of Posts & Telecommunications has confirmed that amateurs with Class B licences may use the translator although this has an output in the 28MHz band.

Final close-down of three old-timers

The deaths of three well-known amateurs have been reported recently. John Curnow (G6CW), of Nottingham, was one of the first British amateurs to use the 1215MHz band. George Courtenay Price (GW2OP), of Pembroke, in pre-war years collated and presented information on aerials in the former R.S.G.B. "Contact Bureau". S. C. MacNamee (EI2A), of County Meath, was for many years v.h.f. manager of the Irish Radio Transmitters Society.

Courses for would-be amateurs

The numbers of British amateur licences continues to grow at a rate approaching 1000 a year. By the end of May 1972 the total of 17,760 (Class A 14,218; Class B 3315; TV 227) was just over 2000 more than in March 1970. September sees the enrolment of students and the start of courses for would-be amateurs at a number of adult education centres throughout the country. These courses cover the syllabus of the Radio Amateurs' Examination of the City and Guilds of London Institute, and most students will aim at taking the examination next May.

Centres known to be offering courses of this type are listed below (based on information from the organizers or supplied by the R.S.G.B.).

London and Home Counties: Beckenham and Penge Adult Education Centre, Beckenham; College of Further Education, Boreham Wood; Robert Haining Centre, Mytchett, Camberley; Cove Further Education Centre, Cove, nr Farnborough; Harlow Technical College, Harlow; County High School for Girls, Ilford; Western and Purley Further Education Centre, Croydon; Holloway Institute, London N19; Slough College of Technology, Slough (also offering "advanced class" mainly for licensed amateurs); Mid-Herts College of Further Education, Welwyn Garden City; and Wembley Evening Institute, Copland School, High Rd, Wembley.

Provinces: Holte Adult Education Centre, Birmingham 19; Corby Technical College, Corby, Northants; St Hugh's Secondary School, Grantham, Lincs; Gosforth Evening Institute, Gosforth, nr Newcastle-upon-Tyne; Newport College of Further Education, Nash, Newport, Mon.; College of Further Education, Oxford; North End Further Education Centre, Portsmouth; Cannock Chase Technical College, Cannock, Staffs.

Without comment

From the *Wirral Amateur Society Newsletter:* "At the sale of surplus equipment last month one might have thought that home construction was a dying craft. Items which would have been snapped up a couple of years ago were withdrawn without a bid. Yet the same members were seen patronizing the trade stands at the Belle Vue Convention as though money was about to go out of use. There can be nothing more satisfying in amateur radio than to work some one, even locally, on a receiver and transmitter you have built yourself."

In brief

An international amateur TV contest, organized jointly by the British Amateur TV Club, A.T.A. in Belgium and A.G.A.F. in West Germany, is to be held on September 23, 24, 30 and October 1, including sound and vision modes on 70cm and 23cm, and sound-only on 144MHz. B.A.T.C. is also holding a national contest concurrently . . . . . The GB3LDN 23-cm beacon station at Shooters Hill in south-east London (possibly the only beacon on this band anywhere in the world) is active on approximately 1297.89 MHz . . . . . The Gibraltar 70-MHz beacon, ZB2VHF, was heard by several British amateurs during July . . . . . The 28.2-MHz beacon, 3B8MS, in Mauritius is now operating . . . . . The new Astronomer Royal, Sir Martin Ryle, holds the amateur call G3CY, though this does not appear to be used very often in recent years — but one of my 1939 QSL cards confirm a 7-MHz c.w. contact with him at a time when his equipment included a l-v-l receiver and a Zepp aerial! . . . . . The 1972 Scouts and Guides Jamboree-on-the-Air is planned for October 21-22. U.K. national organizer is L. R. Mitchell, G3BHK, 28 Darwall Drive, Ascot, Berks. . . . . The Star Short Wave Club is holding two special junk sales in aid of the Radio Amateurs Invalid and Bedfast Club on September 6 and September 27 at New Inn Hotel, Bramley Town Street, Leeds 13 at 19.30 and would be grateful for suitable "junk" (T. Leeman, G8BUU, 115 Asket Drive, Seacroft, Leeds LS14 1HX).

Pat Hawker, G3VA

New Products

Flying-spot c.r.t.

Fast decay time and very low gain noise are two claimed features of a new addition to the range of flying-spot scanning tubes produced by the Electron Tube Division of EMI Electronics. Known as the type MX71, this new 7in diameter, magnetically focused and deflected, cathode-ray tube is primarily intended for use in both 16mm and 35mm colour flying-spot telecine machines.

The high brightness tube incorporates a new EMI fine grain phosphor, the type GGO, which has a broad spectral emission band peaking at 520nm making the tube equally suitable for scanning both positive and negative film without having to modify the colour masking matrices. The decay time is approximately 150ns to 10% and, coupled with the very low grain-noise, contributes to high signal-to-noise figures in all channels.

The high burn resistance of the phosphor is said to give substantially improved tube life under normal operating conditions. The MX71 has a neutral tinted face plate, reducing flare and halation, and is a direct plug-in replacement for EMI's MX69 tube.

Typical characteristics:

resolution	0.1mm
heater voltage	4.0V
heater current	1.0A
e.h.t. voltage	25 to 30kV
cut-off voltage	−90 to −150V

EMI Electronics and Industrial Operations, Blyth Road, Hayes, Middx.

WW325 for further details

Buffered controller

Perex Ltd, of Reading, Berks, manufacturers of off-line data processing systems, have announced their Type 501 controller unit. The 501 provides the facility to enter data at burst transfer rates up to 500,000 words per second, or from low speed inputs such as a teleprinter. Data is recorded on up to four cartridges, by the Tri-Data CartriFile, with extremely low data error rates and high reliability. The combination is described as the Perex 1004 magnetic tape system.

A range of buffer "lengths" is provided by the Type 501 controller in units of 512 words of sixteen bits. The combination of the Perex product and the Tri-Data CartriFile gives a capability to store up to 170,000 sixteen bit words on each of four magnetic tape cartridges. Applications of the system include data logging, and off-line printer stations, where data can be stored for fast transfer to a computer. Price of the 501 controller is £720 ex works. Sintrom Electronics Ltd, 2 Arkwright Road, Reading RG2 OLS.

WW324 for further details

P-N-P Darlingtons

G.E. Electronics (London) Ltd, agents for the Unitrode Corporation, have announced the addition of p-n-p Darlingtons in 5A and 10A versions, to complement their existing range of n-p-n Darlington devices. These new U2T151 and U2T351 series offer voltage ratings up to 120V and minimum current gains of 2000 at the rated d.c. collector current and offer the advantages of a monolithic two-transistor circuit for less than the price of comparable discrete devices. Available in TO-33 lead mounted and TO-66 chassis mounted hermetically sealed packages. G.E. Electronics (London) Ltd, Eardley House, 182/184 Campden Hill Road, London W8 7AS.

WW 302 for further details

Wiring saddles

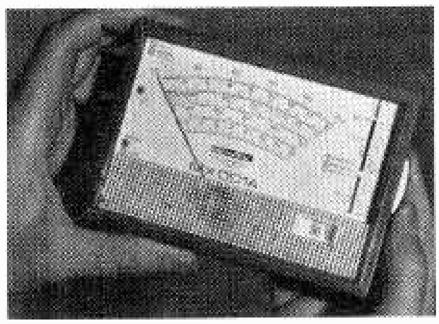
A range of wiring saddles has been announced by component distributors, Peter Bowthorpe Raynor and Associates Ltd. Designed and manufactured by the Richlok Corporation – for whom PBRA are sole U.K. representatives – the saddles are designed to minimize assembly time by

use of a patent self-locking fixing device. Each saddle is fixed, simply by snapping it into a 0.187in diameter hole in the chassis or mounting surface. Made in high de-electric strength nylon, and available in four sizes, the saddles provide a positive, spring locking action on cable bundles up to 1 × ½in cross section. They can be mounted in any plane, or inverted, and require no tools. Peter Bowthorpe Raynor & Associates Ltd, Holmethorpe Avenue, Redhill, Surrey.

WW326 for further details

Pocket multimeter

The MX001A pocket multimeter has been introduced by ITT Metrix and is available in Britain from ITT Components Group Europe. Priced at £12.75 the MX001A can measure voltages up to 160V d.c. using switched ranges, and 500V and 1600V on separate sockets with a sensitivity of 20,000Ω/V; up to 500V a.c. switched ranging, with 1600V on a separate socket; and currents up to 500mA d.c. (5A on separate socket) and 1.6A a.c. Resistances up to 5MΩ can also be measured.



Weighing 400g, the MX001A is fully protected against overload by both fuse and diodes. Range selection is by a thumbwheel switch, with the range selected being shown at a window below the scale.

A full range of accessories is available, including a filter probe for TV line voltage measurements, 15kV a.c./d.c. probe, 30kV d.c. probe, range multiplier resistor box (3000 to 6000V a.c./d.c.) and an adaptor for resistance measurements from 20kΩ to 50MΩ. In addition a carrying case and rubber shock protector are available. ITT Components Group Europe, Instrument Sales, Edinburgh Way, Harlow, Essex.

WW321 for further details

Monolithic regulator

Eurosem International Ltd have announced the introduction by Raytheon Semiconductor of their monolithic 5V regulator, types RC and RM 109. The design features include output currents in excess of 1A, internal thermal overload protection, no external components required and t.t.l. and d.t.l. compatible. These integrated circuits are designed for local regulation on digital logic cards and they may also be used to regulate voltages

above 5V where required. Both military and industrial temperature range types are available in the solid-kovar TO-5 header package and the TO-3 power package; they are equivalents to the LM 109 and LM 309 types. Eurosem International Ltd, Haywood House, Pinner. Middx HA5 5QA.
WW 303 for further details

Frequency synthesizers

A pair of synthesizers have been developed to digitally control the frequency of the tuning heads in the Watkins Johnson model RS-160 PAN/MAN receiving system. The FS-101 synthesizer operates from 2 to 300MHz, and the FS-102 from 2 to 1000MHz. Control of the synthesizer is either through a parallel or serial digital frequency command word input. The command word can be supplied by a computer, a command word generator or a data link. The synthesizer input is t.t.l. compatible and in operation converts the desired frequency data to a $\pm$ 10V signal which tunes the receiver's v.c.o. and preselector to the commanded frequency. Up to 10% of the band may be tuned in less than 10ms, and a tuning head can be tuned from band edge to band edge in less than 100ms. Frequency control is maintained through a phase-locked loop. The receiver's local oscillator is phase-compared to the synthesizer reference source and a correction voltage generated to correct the receiver's v.c.o. and to tune the associated preselector. Long-term internal reference stability is 1 p.p.m. per year over a 0 to 50°C temperature range. Provision for an external 1MHz reference has been incorporated. Applications Engineering, Watkins-Johnson International, Shirley Avenue, Windsor, Berks.
WW 320 for further details

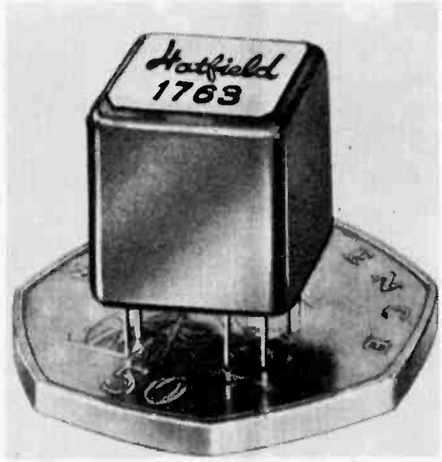
Component identification

A new series of marker cards that provide a quick and permanent means of identifying wires, cables, switches and other components has been introduced by Thomas & Betts. The cards are part of the 'E-Z Code' system and are designed so that the legend, which can be written by hand, is permanently protected by a transparent laminated cover. Made of vinyl plastic, they will withstand water, oils or solvents, and will resist galvanic corrosion, mildew, fungus, most chemicals, and the effects of abrasion.

Four different sizes are stocked in books of 30 to 400 markers, the largest being $1 \times 4\frac{7}{8}$in. The markers are self-adhesive, the adhesive backing being protected by a pull-off tab. Printed markers with various legends are also available. Thomas & Betts Ltd, 90-93 Cowcross Street, London EC1M 6JR.
WW 317 for further details

Miniature mixer

The latest unit in the Hatfield miniature r.f. mixer range is the type 1763. This unit, sealed in a metal can with a glass-to-metal header, is unaffected by humidity and its r.f. screening enables use in strong electromagnetic fields. The 1763 has frequency

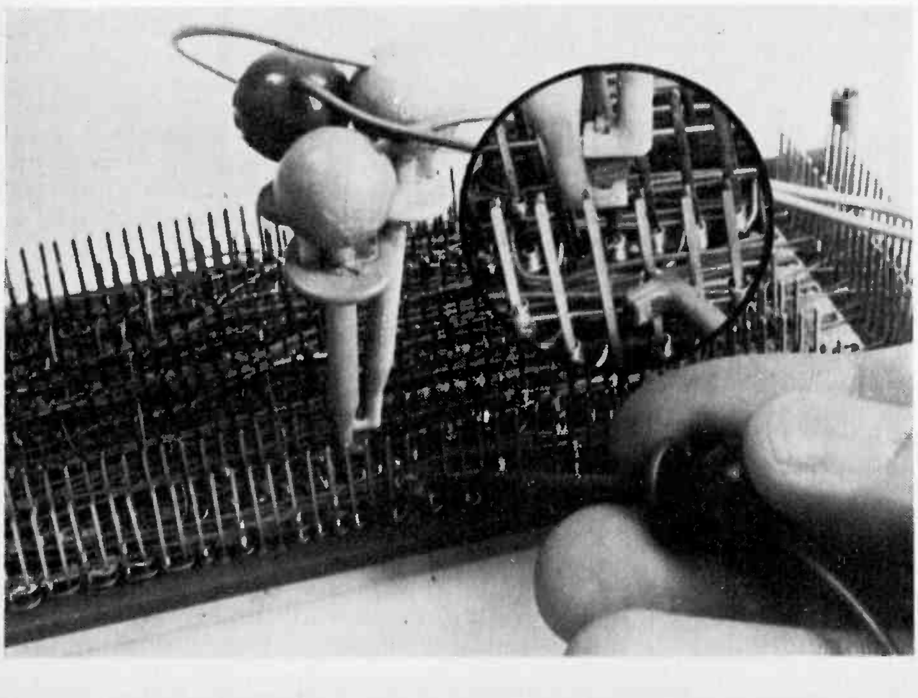


ranges in the A and C ports of 0.1–500MHz and in the B port of d.c. – 500MHz. The temperature range for storage is $-55°$ to 100°C while operating ranges are from $-40°$ to 100°C. Like other mixers in the miniature range the size is extremely compact at $14 \times 14 \times 14$mm, excluding soldering pins. Hatfield Instruments Ltd, Burrington Way, Plymouth, Devon PL5 3LZ.
WW322 for further details

Miniature hook test probe

The full range of U.S. manufactured E-Z-Hook test connectors with a "hypodermic needle" action which clamps them on to fine wires and pins in circuits is now available in this country from British Central Electrical Company Ltd, distributor for the U.K. and Ireland. The smallest in the range, the versatile E-Z-Mini-Hook X-100, can be connected by lead to any testing apparatus, such as a frequency meter or an oscilloscope. Having a fine tip, it can be safely and conveniently used in confined spaces and complex circuits. The E-Z-Mini-Hook costs 30p and comes in 10 colours for coding. Ready-wired to a plug, it costs from 58p, including 32in. of lead.

The E-Z-Mini-Hook is made of durable nylon with a gold-plated beryllium copper conductor and hook. Flash-tested up to 4kV, it will operate on running test up to 2kV, and has a 1.5A capacity with a minimal voltage drop. British Central Electrical Company Ltd, Briticent House, New Street, Ringwood, Hants.
WW323 for further details



Two new Plumbicon tubes

The latest additions to the Mullard range of Plumbicon television camera tubes are designated types S42XQ and S43XQ and have a minimum resolution of 1400 lines. 'Anti-comet tail' electron guns enable the tubes to present much more detail in the brightly lit areas of the picture. The tubes also have an internal light biasing system that greatly reduces lag under poor lighting conditions. The tubes have diameters of 50mm. The S43XQ has a plain glass faceplate, but the S42XQ has a faceplate made of dark-cladded fibre optics so that it can be coupled easily to an image intensifier tube. Suitable for use in closed-circuit television systems in medical schools and medical diagnosis with X-ray intensifiers, they can also be used with advantage in flight simulator systems, and long-range military reconnaissance and surveillance as well as in

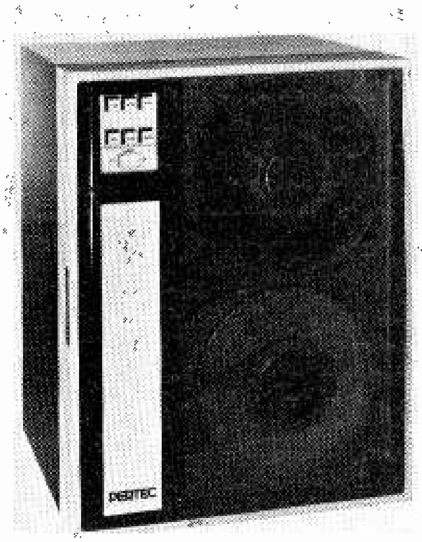
(continued on page 451)

high-quality monochrome and colour television broadcasting systems. Mullard Ltd, Mullard House, Torrington Place, London WC1E 7HD.

WW 314 for further details

Digital tape transport

Pertec's most recent 10½in reel synchronous digital magnetic tape transport features a new quick-lock hub which automatically seats the reel, retractable tension arms, contoured head cover for easier tape threading, a new blade-type tape cleaner and a fast rewind speed of 200 i.p.s.



Designed for use with data entry, remote terminal and mini-computer systems, the Pertec T8000 series transports are IMB and ANSI compatible and available in read-after-write, read/write and read only models, with tape speeds up to 45 i.p.s. and data transfer rates up to 72,000 characters per second. Standard configurations available include 7 or 9 track, NRZI* or 1600 characters per inch phase encoded, or the new NRZI/PE electronically switchable tape formats. The transports also feature single adjustment electronic write de-skewing, dynamic braking, photosensing arm positioner, low power consumption, temperature stable head guide assembly and remote edit capability. The T8000 transports provide forward and reverse read with programmed data recovery capability for reading older or marginal-amplitude tapes. Pertec, 21 London Street, Reading, Berks.

WW 312 for further details

*NRZI – standard abbreviation for non-return zero indexing

"Hall-effect" switches

Quarndon Electronics now have available the Sprague ULN-3000 series of magnetically activated electronic switch utilizing the Hall effect. The ULN-3000M is a monolithic integrated circuit comprising a Hall cell, a trigger circuit and an amplifier encapsulated in an 8-pin plastic dual-in-line package. It requires a 5V supply, will operate with small permanent magnetic fields, has high reliability and small size. Having a grounded-emitter and open collector 20mA output the ULN-3000M can be interfaced directly with d.t.l., t.t.l., m.o.s., r.t.l., s.c.r. triacs, transistors and relays. Price 100 up £0.86. Quarndon Electronics (Semiconductors) Ltd, Slack Lane, Derby.

WW 313 for further details

Electronic message display

Development work carried out by Stabletron on electronic visual displays has resulted in the launching of a new display system, called ANCHOR. It is based on a technique originally developed by the B.B.C. where the messages can be immediately "typed" on to a television screen and relayed to remote displays up to large cinema screen size, or the information can be stored on computer magnetic tape or punched paper tape. The display system generates television-type signals, so it is suitable for use as a caption generator to overwrite on to moving video pictures. The system can operate in black and white or provide multi-colour messages.

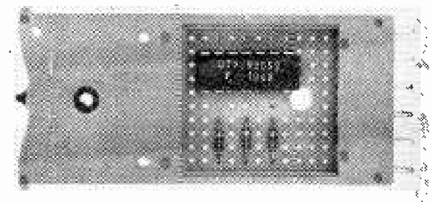
The design permits multicolour data displays to be generated at negligible extra cost, the current model allowing six distinct colours plus white to be shown on the screen using standard television c.r.t. equipment. The keyboard unit comprises a standard alpha-numeric keyboard plus a number of function keys. This enables the operator to type in messages and use built-in text editing hardware. A cursor may be moved in any direction around the screen, and messages can be inserted, deleted or moved in "zip" or "crawl" mode. Transmission or storage of information can be controlled. ANCHOR can be manually or computer driven, via keyboard, paper-tape, magnetic tape, disc, core store, or over Post Office lines. Stabletron Ltd, 266 Hunts Pond Road, Titchfield Common, Fareham, Hants.

WW 318 for further details

Thumbwheel Multiswitch

Contraves Industrial Products Ltd have announced a new series of miniature thumbwheel switches giving optimum character definition with respect to minimum panel space. One out of the range, designated Type C, provides a high degree of mechanical protection for logic elements such as diodes, which can be mounted on the printed circuit board within an enclosed cavity in the switch housing. A wide selection of binary coded outputs is available.

Dimensionally compatible with the frontal dimensions of the Type H Multiswitch, namely, 33mm high and 10mm wide, the Type C switch has terminations avail-

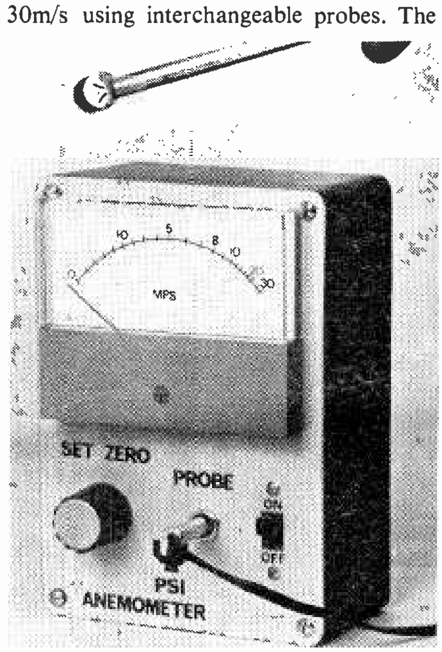


able for solder tag, or for wire-wrapping and other high-speed connection methods. Contraves Industrial Products Ltd, Times House, Station Approach, Ruislip, Middx. HA4 8LH.

WW328 for further details

Air velocity meter

Prosser Scientific Instruments' new portable air velocity meter, designated AVM 500, displays wind speed on a meter, or can be connected to a recorder. Priced at £42.00, the meter covers the range 0 to 30m/s using interchangeable probes. The



moving-coil readout meter is graduated in both metres per second and m.p.h. A recorder output is fitted. Accuracy is 10% of reading under normal conditions. Prosser Scientific Instruments Ltd, Lady Lane Industrial Estate, Hadleigh, Ipswich, Suffolk.

WW 306 for further details

Miniature plugboard

Sealectro are now able to offer the Miniboard programming system with bussed upper and individual lower contacts. Previously available in bussed/bussed form only, the Miniboard matrix programming and switching system is custom built in hole configurations to suit each customer application. Low cost coupled with extremely compact construction are the main features of this product.

In one form the Miniboard provides the equivalent of six single-pole, single-throw, normally open switches with three

inputs each, achieved in an overall size of 0.5 × 1.2in. In use this type of contact arrangement permits the mounting of diodes on p.c. cards using shorting pins to programme a diode matrix instead of using the more expensive diode plugs. Connections to the new beryllium copper single sockets are solder type. Standard finish is nickel silver, but gold plating can be provided. Programming Division of Sealectro Ltd, Walton Road, Farlington, Portsmouth PO6 1TB.
WW 319 for further details

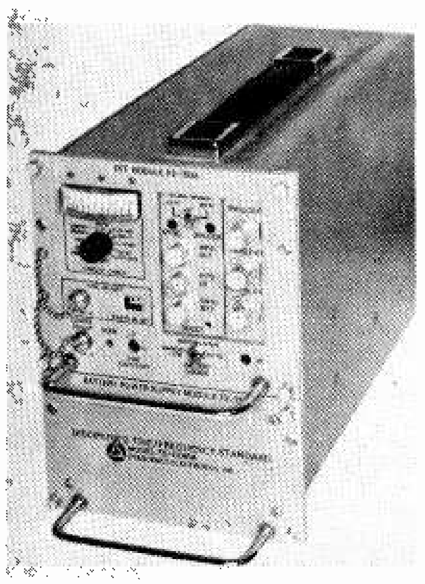
D.c. to d.c. converter

Electroplan introduces a d.c. to d.c. converter with 5V d.c. input and dual programmable output voltages of ±6V to ±15V d.c. This unit operates from a standard 5V logic supply requiring less than 1A at full output load and is easily programmed by a single resistor. A small moulded case encapsulates the circuitry with the input and output connections through printed circuit pins. The positive and negative outputs are isolated from the input and are short-circuit protected. Any output may be connected to any input terminal.

Output characteristic per output channel is 40mA at 12V with output ripple (noise) less than 25mV peak to peak. Dimensions are 31 × 63 × 15mm with the pins on a grid of 0.1in. Price £10 each. Electroplan Ltd, P.O. Box 19, Orchard Road, Royston, Herts SG8 5HH.
WW 309 for further details

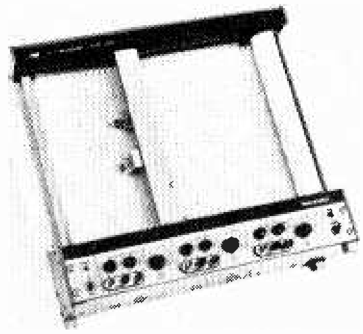
Portable time/frequency standard

The model FE-1050A made by Frequency Electronics Inc. of New York is a portable time/frequency standard with a stability rising to 1.5 parts in 10^{12}. Using a voltage-controlled crystal oscillator, whose frequency is controlled by a voltage stored in a memory storage device rather than a fixed frequency crystal oscillator, the instrument has been designed so that the memory voltage is constantly updated when a phase lock input is present. In the event of interrupted operation, the frequency is maintained at its last locked value with a resolution of 2.5×10^{-12}. High spectral purity outputs of 5MHz and 1MHz sine waves are available from front panel connectors, also a 100kHz logic compatible square wave and a 1 pulse/second output. The model FE-1050A will operate from normal mains power supplies, a 20 to 30V d.c. source or for up to seven hours from the built-in battery pack of rechargeable nickel cadmium cells. Euro Electronic Instruments Ltd, Shirley House, 27 Camden Road, London N.W.1.
WW 315 for further details



X-Y recorders

Yokogawa have announced the Series 3070 X-Y recorders with a range of standard and optional features. Sensitivity of 100μV/cm, f.e.t. chopper with constant 1MΩ input and 0.3% accuracy are characteristics of the YEW 3070 units. The



3077 single-pen and 3078 dual-pen units offer c.m.r. of 180dB at d.c., full width zero shift and remote or manual pen lift. The 3070 series of recorders are available in single or multiple range options with paper take-up accessories for the versatile roll chart. Martron Associates Ltd, 81 Station Road, Marlow, Bucks.
WW 316 for further details

Improved d.i.l. connector

A newly designed d.i.l. connector for p.c. boards, available from Molex, has a grid contact arrangement for improved support and assembly flexibility and tolerance. The new 1938-8 Soldercon terminals are available in chain form or pre-cut strips to fit specific i.c. requirements.

They hold typical on-centre spacing of 2.54mm between terminals of d.i.l. packages. Terminal contact pins alternate front and back on successive terminals, however, and increase to 5.08mm the in-line hole spacing on the p.c. board, with a hole pattern on a staggered 2.54mm grid. Terminals are supplied with a break-off carrier strip to make positioning and soldering easier. The carrier is removed after soldering with a simple hand tool. Terminals can be specified in tin-plated brass or gold-plated (over nickel) brass, and on special order, phosphor bronze. Molex International, 14 Yeading Lane, Hayes, Middlesex.
WW 307 for further details

Two-watt i.c. audio amplifier

A new Mullard integrated circuit has been developed for use in audio amplifiers. Type TCA160 has an input impedance which can be modified with a series input resistor making it suitable for use with ceramic pickups as well as in small television and radio receivers. It will give an output of more than 2W into 8Ω and can be used with supply voltages between 5 and 16V although an unloaded supply voltage of 18V can be tolerated safely. Crossover distortion between 7.5 and 16V is low, hence the TCA160 can be used in battery-operated equipment. It needs few resistors and capacitors to form a complete audio amplifier. The TCA160 is available with a dual-in-line, or "Quil"* encapsulation; both can be supplied with a fitted heatsink that will enable an output of more than 2.5W to be attained. Mullard Ltd, Mullard House, Torrington Place, London WC1E 7HD.
WW 310 for further details

*staggered pin d.i.l.

S-band travelling-wave tube

The range of S-band travelling-wave tubes produced by The M-O Valve Co. Ltd, has been further extended with the introduction of the TWS36. It is a conduction cooled power amplifier giving a small signal gain of better than 30dB and a minimum saturated output power of 18W in the frequency range 1.7 to 2.3GHz, designed for use in telecommunication systems.

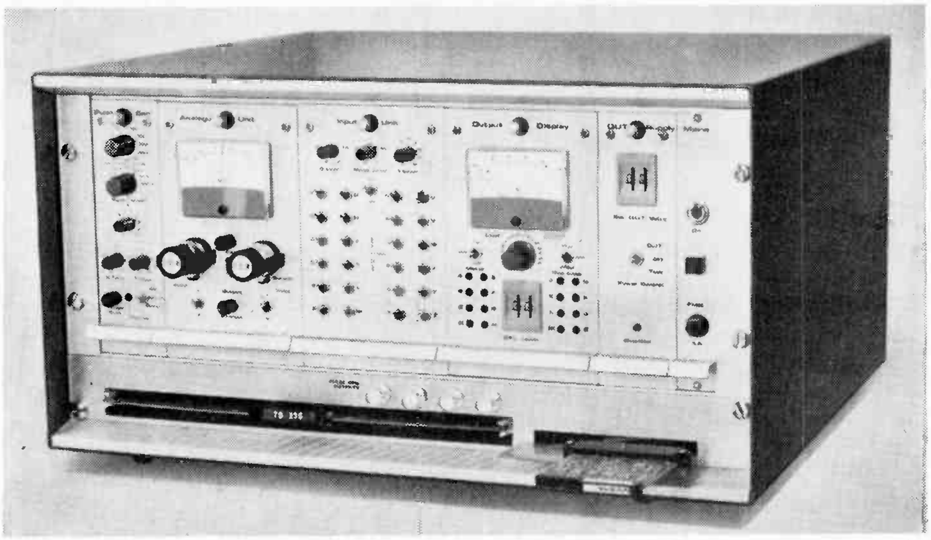
The amplifier is housed in a periodic permanent-magnet focusing assembly and is designed for linear operation at output levels up to 10W. Coupled helix transitions to type N 50Ω connectors provide a good match, and permit extended bandwidth operation at slightly reduced gain over the frequency range 1.6 to 4.0GHz.

Ancillary fittings, including a heat sink, can be supplied which allow the TWS36 to be substituted as a direct replacement in equipments employing the type 7642/TWS10 and thereby eliminating the need for forced air cooling. The M-O Valve Co. Ltd, Brook Green Works, London W.6.
WW 301 for further details

Multipurpose test set

The 2000 series test equipment has been introduced by Quantum Developments Ltd. It is a modular system with p.c.b. interconnection and is suitable for field testing, small batch production testing, and design evaluation of p.c. cards containing linear and logic i.cs. In standard form it consists of a crystal-controlled double-pulse generator, variable level 24-bit input unit with noise injection, 16-bit output display with l.e.ds, metered and loaded, analogue source and monitor with Weston cell calibration, and stabilized variable supplies. Special purpose modules are available or made to order. Quantum Developments (Norfolk) Ltd, 13 Kings Arms Street, North Walsham, Norfolk.

WW 329 for further details



Time delay generator

The Eldorado Model 650 digital delay generator has variable delays in one nanosecond steps from one to 999,999,999ns and a delay accuracy of ± 1ns over this range. Jitter is specified at less than ± 100ps and delay repeatability is better than 1ns over the full range. It can be used in non-synchronous systems and is adaptable to multi-channel delay systems, which can generate a series of unequal time delays that are all referenced to a common trigger.

WW327 for further details

Low-pass 330-1000MHz filters

The Spinner Company of Munich, West Germany, have developed a new range of coaxial low-pass filters combining good electrical characteristics with very small dimensions. These filters have been designed by applying Tschebyscheff coaxial (9th degree) techniques and a substantial reduction in overall length has been achieved by using helical inner conductors. Features include 35dB attenuation at 1.3 times the cut-off frequency, maximum reflection coefficient of 0.1 between cut-off and 0.3 of cut-off frequency. Hayden Laboratories Ltd, Hayden House, 17 Chesham Road, Amersham, Bucks.

WW 308 for further details

Auto-ranging frequency meter

The Model CR400 frequency meter marketed by Dynamco gives readings at frequencies down to 0.01Hz. Ranging is automatic – or can be externally programmed – and the unit has an l.e.d. display. Frequency range is 0.01Hz to 20MHz and the crystal timebase has a stability of 3×10^{-8} per day. Frequency is indicated by measuring the period of the input signal (from 1 to 10^7 periods) and automatically calculating the reciprocal value. As well as frequency measurement, the CR400 can be used as a period meter (50ns – 100), a tachometer (0.0006 to 6000 r.p.m.) and a timer (1μs to 100s). Output facilities include b.c.d. measurement outputs and an oscilloscope monitor point to check the waveshape and the internal timebase. Dynamco Ltd, 91 Beddington Lane, Croydon, Surrey.

WW305 for further details

TO-3 beryllium-oxide washer

The latest addition to Jermyn's range of heat-transfer washers for semiconductors is the type A26-3030, which incorporates a disc of very high purity beryllium oxide. The constructional technique employed, economizes in the use of expensive beryllium oxide. This disc is moulded into a plastic surround, which has a TO-3 outline and provides a thermal performance of 0.1°C per watt and a dielectric strength of better than 14kV per mm. Jermyn Manufacturing, Vestry Estate, Sevenoaks, Kent.

WW311 for further details

Two-frequency tone squelch

Alpha Electronic Services have produced a two-frequency sub-audible tone encoder/decoder for the two-way mobile radio user. The Alpha SS-80J "dual tone", measuring $2\frac{3}{4} \times 1 \times \frac{5}{8}$in, can provide two distinct audio frequencies for the purpose of controlling or selecting repeaters, base stations or mobile units. The Alpha SS-80J employs thick film modules that fit into an edge connector on a small printed circuit board, giving ease of maintenance, modification or change of frequencies. Tone frequencies are also determined by a plug-in thick film module which contains a twin-T, *RC* network. Tone frequencies are available from 20 to 250Hz. Application engineering notes are available. Alpha Electronic Services Inc., 8431 Monroe Avenue, Stanton, California 90680 U.S.A.

WW 304 for further details

Literature Received

For further information on any item include the WW number on the reader reply card

ACTIVE DEVICES

The range of "C-Line" industrial semiconductors (p.u.ts, power Darlingtons, transistors and i.cs, plastic s.c.rs, zener diodes, rectifiers and bridges) is described in a catalogue received from G. E. Electronics (London) Ltd, Eardley House, 182/184 Campden Hill Road, Kensington, London W8 7AS WW401

The complete product range of microcircuits, germanium and silicon transistors, packaged circuits and ancillary devices, offered by Newmarket Transistors Ltd, Exning Road, Newmarket, Suffolk, is described in a "product portfolio" WW402

Specifications, application ideas and interface circuitry of the Motorola MC1400 series of c.m.o.s. logic devices are included in a booklet "McMos from Motorola". GDS (Sales) Ltd, Michaelmas House, Salt Hill, Bath Road, Slough, Bucks WW403

Series SSD-200 data books give information on all RCA solid state products. RCA Solid State-Europe, Customer Technical Data Services, Sunbury-on-Thames, Middx, TW16 7HW. The series includes:

- 201 — "Linear Integrated Circuits and M.O.S. Devices — Data" WW404
- 202 — "Linear Integrated Circuits and M.O.S. Devices — Application Notes" WW405
- 203 — "C.O.S./M.O.S. Digital Integrated Circuits" WW406
- 204 — "Power Transistors and Power Hybrid Circuits" WW407
- 205 — "R.F. Power Devices" WW408
- 206 — "Thyristors, Rectifiers and other Diodes" WW409

We have received a brochure giving electrical characteristics, test circuits and typical performance curves of series 54/74 low-power, t.t.l. circuits. National Semiconductor (U.K.) Ltd, The Precinct, Broxbourne, Herts WW410

A wall chart listing the range of Raytheon linear i.cs provides circuit diagrams and principal specifications. Jermyn Industries, Vestry Estate, Sevenoaks, Kent WW411

Logic diagrams and condensed specifications of m.o.s. circuits are included in a catalogue (May 1972). General Instrument Microelectronics Ltd. 57–61 Mortimer Street, London W1N 7TD WW412

PASSIVE DEVICES

A catalogue of electronic accessories, covering araldite and aluminium knobs to veroboard and wander plugs is issued by Tronic Sales Ltd, 1 Buckwell Place, Wellingborough, Northants, NN8 4LR WW413

A fuseholder incorporating a neon lamp, which lights when the fuse blows, is described in a data sheet from A. F. Bulgin & Co. Ltd, Bye-Pass Road, Barking, Essex WW414

We have received a set of leaflets describing "Nip-E-Boards" — a range of ready-made p.c. boards with facilities for edge connection. Nip Electronics, P.O. Box 11, Beaconsfield Road, St. Albans, Herts WW415

Mullard Ltd, Mullard House, Torrington Place, London WC1E 7HD, have published a wall chart giving details of the company's capacitor and resistor ranges WW416

Quick reference to the miniature relays, solenoids, lever keys, timers and counter timers manufactured by Keyswitch Relays Ltd, Bendon Valley, Garratt Lane, Wandsworth, London S.W.18, is contained in a leaflet we have received WW417

More than 1600 components, from batteries and capacitors to instruments and tools, are listed in the 1972/73 stock catalogue from ITT Electronic Services, Edinburgh Way, Harlow, Essex. Price £1.50

The Erie U.K. '72 component catalogue combines the manufacturer's listings with a distributor's stock catalogue. Components range from resistors and capacitors to semiconductors and thick film devices. Erie Electronics Ltd, South Denes, Great Yarmouth, Norfolk WW418

EQUIPMENT

Tektronix U.K. Ltd, Beaverton House, Harpenden, Herts, have sent us supplement No. 5 to the March 1971 product catalogue — a leaflet including a description of the TM500 instrumentation system WW419

"Your Guide to Digital Instrumentation" is a brief survey of the range (d.v.ms, counter timers, service and standards) available from SE Laboratories, (Engineering) Ltd, North Feltham Trading Estate, Feltham, Middx WW420

Highgate Acoustics, 184 Gt. Portland Street, London W.1, have sent us leaflets describing the Harman Kardon model 630 and model 930 a.m./stereo f.m. receivers WW421

Two data sheets are available on new equipment from Venner — a division of AMF International Ltd, Kingston By-Pass, New Malden, Surrey:

- model 7734 frequency counter (for measurements up to 5MHz. Price £98) WW422
- model 754 signal generator (85kHz to 100MHz with digital indication. Price £250) ... WW423

Brief descriptions of equipment ranging from power supplies and signal sources to digital and analogue measuring instruments are described in a brochure sent to us by Farnell Instruments Ltd, Sandbeck Way, Wetherby, Yorks LS22 4DH WW424

We have received a leaflet describing Olektron d.c.-1000MHz broadband components and sub-systems for receivers, transmitters, aerials and TV distribution. Tony Chapman Electronics Ltd, 3 Cecil Court, London Road, Enfield, Middx. WW425

The Rovac digital multimeter, described in a leaflet from Roband Electronics Ltd, Charlwood, Horley, Surrey, measures up to 2000V and 15A, a.c. and d.c., and up to 200MΩ WW426

Operating specifications of power supply ranges "OA" and "IC", for op-amp and i.c. operation respectively, are given in a leaflet received from Coutant Electronics Ltd, 3 Trafford Road, Reading, RG1 8JR WW427

A leaflet we have received describes a battery-operated on/off electronic cycling timer, series AL. Cycling times of up to 60min are available, with 2-10s on/off switching periods. Tempatron Ltd, 5 Loverock Road, Reading WW428

The Honeywell series 3 illuminated pushbuttons and indicators is the subject of a brochure. A choice of matrix, individual or strip mount pushbuttons is available. Honeywell Ltd, Micro Switch & Keyboard Group, Charles Square, Bracknell, Berks. .. WW429

"Powercard" is a range of power supplies for linear and digital i.cs described in a specification sheet. ITT Components Group Europe, Standard Telephones and Cables Ltd, Rectifier Division, Edinburgh Way, Harlow, Essex WW430

We have received a leaflet describing the "Masteranger" solid-state multitest set (1.5mV to 1500V, 0.15μA to 150A a.c./d.c., resistance ranges to 10,000MΩ). Conway Electronic Enterprises Ltd, Weston, Ontario, Canada WW431

A data sheet is available describing the model 514 portable electronic clinical thermometer. Response time of the thermometer is 5s to steady reading. Polkinghorne Industries Ltd, Lillyhall Industrial Estate, Workington, Cumberland WW432

APPLICATIONS

The contents of a manual on reed switches include terms and definitions, operating magnetic reed switches, physical and electrical characteristics and applications. Hamlin Electronics Ltd, 14 New Road, Southampton, SO2 0AA WW433

The first edition of "Reichert Information" (bi-monthly from June 1972) traces the history of the C. Reichert of Vienna company, and some of the products (optical instruments) which are handled by the Instrument Division of British American Optical, 266 Bath Road, Slough, Bucks. ... WW434

"Coilwinding Techniques" by G. W. Strobel is a paper presented at Coilwinding '72. Several methods of coilwinding are described including layer, bobbin, sky, scramble, orthocyclic and saddle winding. The Plessey Company Ltd, Plessey Windings, Ilford, Essex WW435

"Timing Fundamentals in Color and Black-and-White Television Systems, PAL-B/PAL-M/SECAM" explains the basic principles necessary for correctly timed and phased programming in a moderate size television station. R.C.A. Great Britain Ltd, Lincoln Way, Windmill Road, Sunbury-on-Thames, Middlesex WW436

GENERAL INFORMATION

The "Compendium of Degree Courses 1972" provides information about 400 sandwich, full-time and part-time degree courses and includes entry qualifications. The courses lead to degrees of B.A., B.Ed., or B.Sc. Council for National Academic Awards, 3 Devonshire Street, London W1N 2BA.

A leaflet we have received describes courses available, entry qualifications and diplomas awarded in the department of engineering technology of Twickenham College of Technology, Egerton Road, Twickenham, Middx. TW2 7SJ.

Two publications have been sent to us by the British Standards Institution, 2 Park Street, London W1A 2BS.

- BS4503:1972 "Specification for 9-track magnetic tape for data interchange. Part 2. Tape recorded at 1600 rows per inch, phase encoded" .... Price £1*
- BS4798:1972 "Specification for the electro-acoustic response in cinema auditoria and motion picture control rooms" .... Price 70p*

*(plus 20p for orders under £2)