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# wireless world 

## Electronics, Television, Radio, Audio

## JANUARY 1976 Vol 82 No 1.481

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[^3]

This month's front cover shows a group of Plumbicon* lead oxide photocoriductive television camera tubes available from Mullard Ltd.
*registered trade mark for television camera tubes.

## IN OUR NEXT ISSUE

Radio controlled clock. Design for self-setting digital clock operated by time-of-day code now transmitted by MSF Rugby.

Novel waveguide. New type of structure for dipolemode propagation of microwaves offers practical advantages of smaller size and lower attenuation.

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Principles and characteristics of the thermistor and the somewhat neglected thermocouple, outlining basic measurement methods.

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| P8001 | Monochrome <br> Luminance $+$ <br> Chrominance channels | 30 mm diameter Separate mesh construction | P8131** | Monochrome Luminance $+$ Chrominance channels | Similar to P8130 series but with variable integral fight bias |
| P8003 | Red channe\| <br> Luminance/green channels Monochrome channel | 30 mm diameter <br> Separate mesh with extendedred response. with or without IR faceplate filter | P8132 | Red Channel <br> Luminance/green channels <br> Monochrome channel | Similar to P8130 series but with extended red response. with or without IR faceplate filter |
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## Models of behaviour

The turn of the year sees the world in the trough of an economic recession, the deepest for Britain since the second world war. We are told there may be an upturn in 1976. Will there be another boom? We recognize, at any rate, that we are in the grip of another business cycle - and people have been experiencing these now for over a hundred years. Thoughtful engineers observe that if these events really are cyclical there must be a system at work somewhere, or rather a complex of interacting system. And systems can be measured, analysed and controlled to make them do what we want. But are the politicians, economists, business leaders, trade union officials and others who run our affairs really aware of this? Are they capable of even conceiving that there are mechanisms at work, let alone trying to make use of this knowledge?
A small elite, of course, is aware. In the writings of the econometrists, for example, you will find plenty of talk of input, output, gain, positive and negative feedback loops, damping, stability and the like. The fact that many of these terms are consciously or unconsciously borrowed from electronics is no accident, for electronic engineering has provided a whole universe of concepts and analogies which are highly flexible and readily adaptable to the complexities of economic systems.

In this issue we introduce a new series of articles which may make a small contribution to the understanding of systems in human affairs (economic, biological, sociological and so on) in terms of the models provided by electronic physical processes, circuits, systems and engineering concepts. It is based on Professor Chaplin's pioneering work on A-level electronics courses for schools - and this is a significant point. In the sixth form, young people's ideas about the world are still reasonably fluid. They are not yet bound to the conceptual framework of the historian, the economist, the physicist, the geographer, the biologist or the engineer. At this age it's not a bad idea for our future politicians, etc., to study a set of physical models embodying universal principles. In this respect the subject is rather like the classics. It certainly provides a better mental frame of reference for understanding the systems in human activities than those offered by the conventional disciplines.

Electronics, of course, has also given us instruments by which our models can be quantified - analogue and digital computers. We have, for example, Professor Stone's input-output model of the British economy at Cambridge, handling about 300 identities and 700 relationships on a digital computer, by which it is hoped to forecast the state of the economy in 1980. Other computer models have been concerned with the activities of man on a global scale, predicting quantitatively the interaction of population growth, industrialization, malnutrition, depletion of natural resources and deterioration of the environment.

We must be very careful, though, not to lead people astray by simple deterministic models which don't take account of all the variables. Processes in the real world are often non-linear, probabilistic rather than deterministic, have time delays, and include behaviour caused by such things as human panic and euphoria. The impractical ideal is for our electronic models to provide ways of representing all these characteristics. The practical realization is never to accept that the ideal has been achieved.

# Audibility of phase effects in loudspeakers 

# Are "linear phase" loudspeakers a gimmick? 

by H. D. Harwood, B.Sc.

BBC Research Department


#### Abstract

The fact that most loudspeakers have a non-uniform phase-frequency response has long been known and the audibility of the effects of the phase changes in loudspeakers has been debated in the literature many times. This article reviews the position in the light of present-day knowledge and especially of the fact that some manufacturers are now selling loudspeakers in which it is claimed that particular attention has been paid to the phase-frequency response. The article is concerned with the effects of phase from two aspects - that of sound quality and that of stereo image formation.


Taking the question of sound quality first, it is a commonplace observation that the polarity of the connection to the loudspeaker has almost no effect on sound quality. A few observers have claimed to be able to detect such a change on particular types of programme but even if this is accepted as correct the effect is so small that no programme producer, disc manufacturer, pickup manufacturer or loudspeaker manufacturer has ever thought it worthwhile to attempt to specify which polarity is correct. Furthermore, no producer attempts to bother about polarities when mixing the outputs of single multi-microphone pickups for light music programmes. This result is at first sight somewhat surprising, as a programme waveform is well known to be appreciably asymmetrical and it is not at all obvious that polarity should have so little effect on such items as the beating of a drum.

This lack of perception of phase over the whole band can be confirmed by changing the relative phase of the loudspeaker output by means of two all-pass type networks. This can be done on an A-B basis using one loudspeaker and if the circuits are carefully constructed for both amplitude, phase and distortion, the effect is quite inaudible. This has recently been checked at the BBC Research Department using a modern high quality loudspeaker and various types of programme. These included a violin piano sonata, informal speech with mixed voices, light orchestra, choir and symphonic orchestra. On this stringent type of test it is essential to hold the amplitude response to very close limits and for the non-linearity distortion to be of a very low order if the comparison is not to be invalidated by such factors.

The effect on sound quality of relative phase changes over the whole audio band of $90^{\circ}, 60^{\circ}$ and $30^{\circ}$ was also tested and in no case was any audible effect produced.

In all these experiments of relative phase shift, the absolute phase shift introduced is very much greater than the relative shift and varies smoothly with frequency as shown in Fig. 1. The change in phase between 30 Hz and 10 kHz is of the order of $500^{\circ}$ yet the effect of this on sound quality is still inaudible. Note that this change does not even correspond to a simple time delay, the delay at the bass is roughly 6 ms while that in the treble is 0.15 ms .
The inaudibility of such phase changes has been confirmed by Bauer ${ }^{1}$, yet ${ }^{-1}$ the distortion produced by such ${ }^{-}$a network on the waveform of a rectangular wave was clearly shown in his article. The need for a loudspeaker to be able to reproduce accurately a square wave ${ }^{2}$ is therefore more than doubtful.

From these observations it is concluded that sound quality is not affected either by constant or by smoothly varying phase changes over the range normally met in loudspeakers. It is quite likely that such changes produced by


Fig. 1. Absolute phase-frequency response curve of all-pass networks used for phase shift experiments.
the use of very long horns would produce audible effects, but such devices are rare in practice.
In contrast, the effects of rapid local phase changes on sound quality must now be considered. Various workers, such as de Boer and Craig \& Jeffress, have shown that audible effects can be produced when using artificial signals, by what are, in essence, rapid local phase changes. The conditions which are necessary to achieve this audibility are not clear, except that authors such as Goldstein" have suggested that, for example, the difference between an amplitude modulated signal and a frequency modulated signal (which differ only in phase) is only audible within the critical bandwidth, at the corresponding carrier frequency.

On the other hand examination" of the audibility of resonant peaks of various Qs and frequencies for minimum phase shift networks has shown that under certain conditions the audibility of a peak of given height actually decreases as the Q increases, that is as the rate of change of phase increases. This surprising result, since confirmed by Moulana ${ }^{7}$, was accompanied by evidence that under these particular circumstances the subjects were listening to the steady-state condition. On reversing the phase of the resonant channel relative to that of the straight through channel it was possible, by adjusting the level of the signal in the resonant circuit, to cancel out thie coloration so that the effect of the resonance was inaudible. This cancellation can only occur for the steady state condition; the time response cannot so be cancelled. Under these conditions the effects of phase wère obviously inaudible.

Attention has also been drawn" to the
anomaly that under other conditions, which again cannot yet be specified, subjects arrive at precisely the opposite conciusion, namely that the audibility of a resonant peak of given height increases with increasing $Q$. Whether in this last-mentioned case it is the phase change or the amplitude-time response which is audible might be thought to be immaterial, as for these networks the two are obviously irrevocably tied together. On the other hand if a simple resonant element such as a tuning fork is struck, the audibility clearly falls with the amplitude and does not follow the constant phase changes.

It is also a matter of common observation that narrow interference crevasses are inaudible. However, these by their very nature must contain very sharp phase changes as well as the more obvious amplitude changes, yet they appear to have no effect on sound quality. One rather interesting example of this was tested at BBC Research Department by P. C. J. Hill in connection with another investigation. The amplitude-frequency response curve of the circuit used is shown in Fig. 2. For reasons connected with the original experiment the bandwidth was limited to 6 kHz , and the effect of this circuit was compared with that of an otherwise unadulterated programme restricted to the same bandwidth. Recordings of speech, piano music and dance band were used in the assessment; in the associated subjective grading of 1 to 5 , one unit represented only "slightly worse than the stan" dard." The ratings obtained were 0.6 for speech, 0.8 for piano and an improvement of 0.3 for the dance music! Obviously the overall effect of the irregularities due to the combined effects of phase and amplitude was very small indeed, even though the phase response must have been irregular.
Again, no one objects to listening in a live room, yet the effects on phase of the first reflections and of the eigentone structure must be devastating and vary rapidly with the position of the source or listener. The effects of even appreciable changes in head position are however inaudible.
If now we revert to the condition of lumped constants with high Qs, the question arises as to the limit on $Q$, or rate of phase change, which is inaudible. For example I have often found it necessary to damp cross-over networks of second and third order to remove audible coloration. However it adequate damping is used the crossover frequencies are completely inaudible even though the users in the BBC have the advantage of direct comparison with the live programme. This condition appears to be all that is necessary subjectively even though the phase changes are considerable.

In this connection the BBC monitoring loudspeaker type LS5/5 (ref. 8) is recognised to possess a sound quality of the very highest order yet the phase


Fig. 2. Amplitude-frequency response curve of circuit used in Dr Hill's experiment.


Fig. 3 (a) Phase-frequency response curve of monitoring loudspeaker type LS5/5 (b) Phase response plus $1 / 4 \mathrm{~ms}$ time delay.
response, measured at 1 m , is as shown in Fig. 3(a). The loudspeaker is a three-unit design with cross-over frequencies at 400 Hz and 3 kHz . The phase irregularity at the first frequency is small but that at the last is large, the high frequency unit being acoustically in front of the middle frequency unit, although mechanically the front surfaces of all three units are in the same plane. The major phase irregularity caused by this is totally inaudible and no complaints of the sound quality in this region are made.
The phase irregularity shown in curve 3 (a) can be largely removed by the addition of a $1 / 4 \mathrm{~ms}$ time delay applied to the whole loudspeaker. The results are shown in Fig. 3(b). The curve is now a rough approximation to a straight line, and this in turn was shown earlier to be inaudible. It is also well known that changing the listening distance by about 3 in on the axis of a loudspeaker produces no audible change. This confirms the innocuous nature of curve (a). This experiment also makes nonsense of absolute phase angle curves, for the addition of a small time delay has completely changed the shape of the curve. If on the other hand time
delay itself were plotted such an addition would merely result in a small vertical shift; the axial time delay/ frequency response curve is shown in Fig. 4. It is clear that this has an innocuous shape.

It seems therefore that the situation as far as the effect of phase on sound quality is concerned, is rather like that of the old kady passenger flying for the first time. Some time after take-off she looked out of the window and noticed that the wings were flapping up and down. Calling the steward she complained that the plane was unsafe and demanded that it should be stopped. The steward assured her that the plane was quite in order and proceeded to pull down the cabin blind, asking the lady if she could now feel any effects due to the wing movement. She could not and was quite happy. As regards the effects of phase on sound quality therefore until someone can demonstrate that they are audible on programme items as opposed to artificial signals, I shall proceed to pull down the blind and pretend that such a factor does not exist.

The second aspect of phase in relation to loudspeakers is connected with the formation of stereo images. Here our ignorance is even more marked than for sound quality as we do not know just what factors are necessary to determine image sharpness.

It would appear at first sight that if the two loudspeakers of a pair were identical in amplitude and phase response the image sharpness would not depend on how uniform with frequency either of the characteristics were. In practice the situation does not seem to be as simple as this. Pairs of differing types of loudspeaker even when matched for amplitude-frequency response do not all give the same sharpness of image even for a symmetrically placed listener. Thus in two types designed by the author the same bass units and high frequency units were used for the two unit system. One design employed the units in a vertical array but for the other design the tweeters were mounted concentrically in the centre of the bass unit on a perforated panel. This last-mentioned design is reckoned by the users to have appreciably better image forming properties than the in-line design even though the same units and similar cross-over networks are used. That the improvement is not due to the concentric arrangement is shown by the fact that a later design ${ }^{8}$ which uses three units in a vertical array gives a sharper image still. It is this last loudspeaker whose phase-frequency response is given in Fig. 3(a). Furthermore, another commercial concentric design gives the worst image sharpness yet known.

A futher experiment has been carried out inserting one of the all-pass phase delay networks whose response is shown in Fig. 1 in series with each of the type LS5/5 loadspeakers and determining the effect both on the sharpness of a
central image, and, in view of the non-constant time delay, on possible chromatic aberration effects when using male speech as the input signal. No effect was discernible, even on an A-B test.

From this result it is not clear how the use of a circuit to make the phase-frequency response linear would help to improve the image width.

In this connection it is worthwhile recalling data published some years ago ${ }^{9}$. which show that for a central image the image width from a pair of loudspeakers is actually narrower under live conditions than under free-field conditions. The only obvious explanation of this is that the reflection from the floor in the live room assists in the detection of the direction of the apparent source. This has been confirmed recently during, recent tests in a free-field room.

Under these last-mentioned conditions subjects complained that the stereophonic image was more diffuse than that to which they were accustomed using the same loudspeakers in a normal live listening room. At my suggestion, a floor of 6 mm hardboard was put down in the free-field room, and all the subjects were then content. The acoustic axis of the loudspeakers was about 1 metre from the floor and if the distance of the observer from the loudspeaker is taken as 2 metres then under the simplifying assumption of no phase change on reflection from the floor there will be an interference crevasse at the listener's ear at 115 Hz and at intervals of 230 Hz thereafter. The phase irregularities corresponding to this kind of interefernce must be very large, but far from destroying the stereo image it is seen to actually assist in improving the sharpness.

This does however raise the interesting question as to whether the loudspeaker amplitude response at, say, $-45^{\circ}$ in the vertical plane (which is usually ignored) is important in this regard.

Even if it were eventually found that phase had a second-order effect on image formation, it is difficult to see how it can be effective above about 2 kHz . It is known from studies of the stereo effect that the relative phase. between the stereo signals is important up to about this frequency, but above this the ear appears to use the envelope of the signal rather than the absolute phase. It is therefore once again extremely difficult to imagine that producing a phase-linear loudspeaker will assist in any way in improving the production of a stereo image above 2 kHz .

A further piece of evidence is coming to light in experiments I have started to measure the image width produced by various designs of loudspeaker. This work has only just commenced but preliminary results indicate that when using male speech as the source material the image width for wide range material is narrower than that obtained for any


Fig. 4. Axial time delay/frequency response corresponding to Fig. 3(a).
one octave band of the same material, indicating that the brain is using some form of summation of the information received.

To conclude, any effect either of wide band or narrow band variations of phase on sound quality is at most of a second order, and is certainly minute compared with first-order effects such as coloration, and compared with these can be ignored by loudspeaker designers and users even though not by psycho-acousticians. Until, therefore, someone can demonstrate that phase effects of this nature affect the sound quality of loudspeakers on programme, and not just on highly artificial signals, the appropriate treatment appears to be to pull down the blind and pretend that phase, in this sense, does not exist.

From the aspect of stereo image formation we must admit our ignorance as to all the factors involved, but in view of the data presented earlier, I will continue to listen to the siren voices of the manufacturers, but again until some audible effect can be demonstrated I will remain absolutely agog - with indifference!

Thanks to the Director of Engineering, BBC for permission to publish.

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Books Received

Foundations of Wireless and Electronics, 9th edition by M. G. Scroggie. The first edition was published in 1936 and since then, nearly 250,000 copies have been sold, which in itself is a recommendation for the current ninth edition. Like previous editions, the book covers basic theory, starting with elementary "first principles" where no previous technical knowledge is assumed. Subsequent chapters discuss a.c. theory, valve and transistor theory, equivalent circuits and oscillators. The book then covers radio senders, transmission lines, radiation and aerials, and detection.
Final chapters deal with low and high frequency amplification, selectivity and tuning, the superheterodyne receiver, cath-ode-ray tubes, electronic waveform generators and switches, computers, and power supplies. As with all of the author's work, the text is clear, concise and very readable with many circuit and pictorial diagrams where necessary. Price $£ 3.75$. Pp.521. Newnes-Butterworth, Borough Green, Sevenoaks, Kent TN 158 PH.

The Autumn 75 series of D.A.T.A. books is now available which covers relays, microwave tubes, memories, linear and digital integrated circuits, current/discontinued diodes and transistors, power semiconductors, thyristors and, a new addition to the range, opto electronics. Most of the books have two editions a year and contain a section with suppliers' names and addresses. The books are also provided with a cross index which speeds up the location of a device, and devices with similar characteristics are listed together. London Information (Rowse Muir) Ltd, Index House, Ascot, Berkshire SL5 7EU.

We have recently been able to examine the Bugbook system of teaching the use of digital integrated circuits. Bugbooks I and II are intended to lead a student from a state of complete innocence, via an extremely methodical procedure, to an understanding of fairly complex t.t.l. functions (registers, counters, memories, etc.). Each section consists of an introduction to the subject area, a statement of objectives, definitions and symbols when necessary, discussion of the subject and experiments. The section finishes with a review of the work. Teachers' course books are available.

Experiments are carried out with the aid of a universal socket board and a selection of small printed-board assemblies (clocks, indicators, power units, etc) which plug into it, a rather neat idea being the application of pin-identifying stickers to the integrated-circuit packages (bugs).

Books and hardware are expensive. The two books cost $£ 12.82$ and a typical accessory board, consisting of a p.c.b. about $32 \times 20 \mathrm{~mm}$, two resistors and an I.e.d., two lengths of wire and two croc. clips, costs $£ 7.36$. The course material is exhaustive (but not exhausting, if a case quoted in the foreword is typical) and, with reservations over the price, can be recommended for examination. The books, produced by E \& L Instruments, Derby, Connecticut, can be obtained from Hepworth Electronics, Kidderminster, Worcs.

## Electronic systems - 1

# An introduction to the nature of the Universe (or an atom) 

by W. E. Anderton
Assistant Editor, Wireless World

This article heralds the beginning of a series of tutorial articles on the subject of electronic systems, their theory, applications and consequences. The content will be based on a proposed Advanced Level course for schools which has been undergoing a trial at nine schools for a period of three years. This course is the culmination of six years' preparation by Professor G. B. B. Chaplin of the University of Essex, who, with the assistance of his colleagues, has detailed most of the proposed syllabus. Wireless World will use this course as a basis for the forthcoming series. We shall, however, diversify and expand its content where we feel necessary, with the hope that a contribution will be made to what we believe is an excellent cause.

The subject matter of "Electronic systems" is split into (a) communication systems, (b) computer systems and (c) feedback systems. The theory is applicable to almost any subject and helps. the student to appreciate that apparently unrelated phenomena are often embodiments of fundamental principles with universal application. The articles will also aim towards a more rigorous and academic knowledge of these fundamental principles through study of their application to electronics.

Since the human body itself is made up of many systems and man throughout his life is surrounded by systems, an understanding of them (whether they are biological, sociological, environmental, mechanical or electronic) is necessary so that science can benefit our society and enable it to progress. All systems have certain fundamental principles in common, and it so happens that they can be described particularly easily and effectively in terms of electronic systems. Electronic experiments to illustrate these principles can be set up quickly and cheaply and their flexibility encourages initiative on the part of the pupil.
Each section of the series is to be introduced by relating man and his needs, including his relevant physical and mental characteristics, to the electronic system under discussion. The content will not emphasize technology,

but will rather seek to establish in the reader a knowledge and understanding of fundamental principles such as feedback, stability, storage and retrieval of information, signals and noise, information content and decision making. Some current technologies will also be discussed as applications of these basic principles.

The word "system" seems vague at first as it can apparently be applied to

Sixth form students working on one of the A-level electronic systems course experiments.
any subject under the sun without explaining anything useful about it. But this is not at all true as a definite theory of systems, how they work and how they can be usefully manipulated has been evolved. This theory can be applied to politics, chemistry, biology, sociology, evolution and so on ad infinitum, the system being defined to encompass just what parameters may be involved with a particular problem. The largest system imaginable is the Universe and one of the smallest is the atom, but no matter how large or small, the same theory can be applied. Our concern will! naturally be with electronic systems theory and with the consequences and benefits that can be achieved.

First, the starting point must be a definition. According to the Concise Oxford Dictionary, a system is "a complex whole, a set of connected things or parts, an organized body of material or immaterial things."
The first point to note is that a system is a complete entity and can thus be represented pictorially by a box (Fig. 1) within which everything is subject to the laws of how a system operates. To determine what these laws are it is necessary to divide the system into its basic component parts, and these can be



Fig. I. A system can be represented as a single box in its most simplified form.
obtained simply by examining the dictionary definition. A system is complex and therefore needs to be capable of processing or working out its complexities so that they are ordered and not disordered. It is a set of connected things and therefore needs communication to make the connections. Finally, it is organized and therefore needs control to achieve organization.

One more factor must be added to the list of processing, communication and control and this is the application of feedback, which is involved with all three. Here information is fed back from output to input so that a close watch can be kept on the way things are going within the system. If they are going astray then the information fed back to tell this to whoever or whatever is in control can be used to adjust the system's operation and straighten out any problems that have arisen. More rigorously, feedback is the process whereby a measured portion of information is taken from the output of a system or its component parts and is fed back to the input where it is added in such a way as to increase the input (positive feedback) or decrease the input (negative feedback).

As an example of how a system can be split into its relevant component parts for analysis, consider Fig. 2. This is a block diagram of the system representing a town's community, with the problem of controlling the numbers of its population to a comfortable level. The elements of feedback, control, communication and processing can be identified as follows:
At the "input" to the town there are births and immigrants who swell the population and at its output there are deaths and emigrants who diminish it. The problem is to maintain a balance

## Suggested reading

Two books which will develop a greater understanding of the interdisciplinary nature of systems are:

Wiener, N. "The Human use of Human Beings," Sphere Books, London.
Wiener, N. "Cybernetics: or Control and Communication in the Animal and the Machine," The M.I.T. Press, Cambridge, Massachusetts.

The following topics will be covered in subsequent issues:
Introduction to communication systems
Modulation
Transmission
Reception and demodulation
Audio perception and bandwidth
Audio systems and transducers
between the actual population and the facilities that are provided in the town such as housing, hospitals, shops and parks.

Information is fed from output to input.so that all relevant facts can be supplied to the town hall where they are processed. The results are communicated to the town's services such as police, hospitals, etc., who are then able to control to maximum efficiency of services which they offer. Communication between the town hall, the social services and the town's population is effected by means of newspapers and radio broadcasts. This briefly shows how a social system is built of the prime factors. It will be the aim of the following series of articles to examine each in turn in terms of electronic systems.
Fundamental to the analysis of a system is the recognition of whether it is deterministic (entirely predictable according to formal laws) or stochastic (not entirely deterministic, but subject to probability). This means recognizing the fact that we may be able to say from one analysis that "a system definitely does this" and from another "a system definitely does this, but . . ." For practical purposes, electronics is exact


Fig. 2. Simple analysis of a social system. There would be many more factors to take into account which have not been considered here.

Audio systems and amplifiers
Measurement in audio systems
Visual perception
Visual systems - scanning and detection
Television systems
These describe the "communications" part of the syllabus and will be followed by related experiments:

- Frequency response of audio systems
- Amplitude demodulation
- Power in electronic circuits

Television waveforms
Topics continuing from this will be under the main headings of feedback and control systems, computer systems and basic electronics.
enough to forget about the "buts" and this is why the application of systems theory to electronics is such a powerful and important tool.

## Sixty Years Ago

Wireless World has always been conscious of the need to make instructional material as interesting as possible - there is never anything wrong with a little humour in even the weightiest piece of reasoning. But we have never approached the level of popularization advocated by Mr Stephen Leacock, as reported in our issue of January 1916. One wonders what he would have made of avalanche diodes and buffer stages.
"A matter under discussion at the moment is the actual value to the student of a mathematical training, and whether it is necessary to go deeply into the subject in order to become a competent electrical engineer. Many have complained that the ordinary methods of imparting mathematical knowledge are far from interesting, and the famous Canadian humourist, Mr Stephen Leacock, in his new book Moonbeams from a Larger Lunacy, particularly emphasises this point of view: He writes as follows:
"Here, for example, you have Euclid writing in a perfectly prosaic way all in small type such an item as the following:
"'A perpendicular is let fall on a line BC so as to bisect it at the point $C$, etc., etc.'
". . . just as if it were the most ordinary occurrence in the world. Every newspaper man will see at once that it ought to be:
> "AWFUL CATASTROPHE. PERPENDICULAR FALLS HEADLONG ON A GIVEN POINT.

The Line at C said to be Completely Bisected. President of the Line makes Statement."


## Atomic heart power

A nuclear-powered pacemaker that could bring a new lease of life to thousands of people suffering from certain types of heart trouble is to be made at the Harwell Atomic Energy research Establishment. Conventional pacemakers with chemical batteries have to be replaced surgically about every three years. Trials indicate that the implanted lifetime of the Harwell units could reach 10 or even 20 years.

The nuclear battery that powers the pacemaker consists of a heat-producing nuclear source, containing less than a fifth of a gramme of plutonium oxide, and a miniature thermocouple, all in a strong metal housing. The assembly is encapsulated in an epoxy resin known to have lasting compatibility with living tissue. Internationally agreed safety standards require stringent tests on the batteries. These ensure that there is no radiation hazard and that they will withstand the most severe accident including cremation. The isotope plutonium 238 has been chosen for the heat source because its power output falls by only one per cent per year, and its radiation, mainly alpha particles, requires a minimum of shielding for implantation. This isotope is specially prepared for the pacemaker batteries and is markedly different from the very long-lived plutonium isotopes associated with nuclear fuel.

## Braille information service

A pilot scheme to produce a monthly braille information service for blind computer programmers and analysts has been started by INSPEC, the information services division of the Institution of Electrical Engineers, and the Warwick Research Unit for the -Blind. Each month, INSPEC supplies the Warwick Research Unit with a computer-readable magnetic tape containing abstracts of recently published
articles from computer science journals. The Research Unit uses an RXDS Sigma 5 computer system with a braille embosser on-line to produce selective listings from the tapes, which are converted into contracted braille by an automatic translation programme. The pilot service has now been running for several months with a limited group of users. Both the INSPEC source tapes and the service operated by the Warwick Research Unit are being made available free to the blind user or his employers. Computing has been able to offer particularly good opportunities for the employment of blind staff on equal terms with their sighted colleagues, and there is now a significant group of blind computer programmers and analysts in the UK. It's in the hope of finding a solution to their information needs that the INSPEC/Warwick project has been started.

## London honours Blum lein

The Greater London Council is to erect a commemorative plaque on the wall of the house where Alan Dower Blumlein was living at the time he met his death in a Halifax bomber crash in 1942. In bestowing this honour, Blumlein's native city of London is placing him among the immortals in recognition of his many achievements in science and technology. Blumlein's inventions
included modern gramophone stereo recording techniques, the closely-coupled inductive ratio-arm bridge, the cathode follower circuit and the socalled "Miller" integrator. The BBC adopted the television broadcasting specification on which Blumlein worked in the 1930's, in preference to Baird's, when they opted for the 405 -line, 50 frames/s interlaced system still in use today. During his short working life, Blumlein was awarded an average of one patent each six working weeks, a record not yet equalled. The official unveiling of the plaque, which is to be placed on a house in Ealing, West London, will be in 1977 on June 7, the anniversary date of Blumlein's death near Ross-on-Wye, Herefordshire.

## Microprocessor comes clean

A new range of dedicated microprocessors just introduced is suitable for all types of washing machines and offers the seven main wash programmes recommended by the International Standards Organization's system of fabric care labelling, together with additional pre-wash, rinse and spin, biological and fast wash programmes. Another feature of the circuit is that it allows for the automatic release of a fabric conditioner during the final rinse. It replaces the conventional electromechanical timer and various safety

Monitoring speech. Strain gauges mounted on flexible metal sensors are used to obtain information on lip and jaw movements made while speaking. Honeywell tape and oscillograph instrumentation records the facial movements for analysis in Seattle, USA, at the University of Washington's Speech Physiology Laboratory.

features are incorporated, including door interlocks, no machine action or heating without water present, and a "child-proof" programme lock-out which, after a programme has been selected, inhibits change of programmes in mid-operation without first resetting.
The heart of the processor, the ITT 7150 , is contained within a 28 -pin d.i.l. package and consists of two r.o.ms which perform the control functions. Up to ten machine functions can be controlled by means of a maximum of 20 different programme steps which provide the various combinations of fill, wash, heat, pump and spin. The package has been designed for m.o.s. applications and is capable of withstanding adverse conditions, e.g. $85^{\circ} \mathrm{C}$ at $85 \%$. relative humidity under reverse bias.

## Head-up for Harrier

The electronics content of military aircraft continues to grow. A system incorporating an integrated head-up display and weapon aiming computer, with cursive and raster outputs and designed specifically for carrier aircraft, is to be supplied for the Sea Harrier. The display system will consist of three units; a computer which generates a very wide range of air-to-air and air-to-surface weapon aiming and display generation functions; a pilot's display which presents the pilot with all the appropriate flight data and weapon aiming symbology superimposed on his view of the outside world; a pilot's control panel for mode selection and display control. The computer unit will contain: a high-speed digital processor having an associated store of 20,000 words of which 16,000 words will be fully reprogrammable core store; hybrid analogue and digital input and output peripherals, to interface with other aircraft systems and sensors; and a symbol generator to provide deflection wave forms to the pilot's display unit. There will also be a video synthesizer section which, under control of the central processor, will give symbology outputs in raster form for use with raster type displays. The inclusion of a core store will give considerable flexibility during development and will greatly simplify changes to the role of the aircraft during service. Smiths Industries is to supply the display system.

## Single-tube colour TV camera

Work is continuing on the single-tube television camera developed by Siemens and demonstrated at the Montreux television symposium. The "Interplex" camera, as it is called, uses a new type of dichroic strip filter. In contrast to a normal three-tube camera,


Market research has told us that the average age of our reladership is getting lower. Is Richard Tucker of Lutterworth in Leicestershire the youngest reader at $21 / 4$ ?
the colour analyser used to break down the image arriving from the lens into red, green and blue components is integrated within the complex pick-up tube itself. This has made it possible to reduce the size of the camera by dispensing with the accessories for colour registration.

Signal information supplied by the television camera tube in the 4.43 MHz subcarrier band is converted into standard PAL signals in a decoder using comb filter systems and other circuitry. Black-and-white and colour information is separated by the comb filters, the spectral lines of the video signals being broken down into chrominance and luminance information. The silicon tube is said to achieve a resolution of approximately 6 MHz in red-green-blue operation. The signal produced by the camera is compatible with black-andwhite sets, and can be recorded without decoding on conventional video tape recorders with colour capability, preserving a claimed 6 MHz resolution.

## Traffic information service

The BBC is still working on its m.f. system for broadcasting traffic information bulletins to road users (see May 1973 issue p. 239). They claim it is more economical than other schemes and will not involve depriving car radio listeners of their normal programmes.

A network of about 80 low-power medium wave transmitters would be set up in the pattern of a grid covering the whole country. They would all operate on one frequency just within the tuning range of a medium-wave receiver. Each station would only transmit for perhaps
half a minute in every eight, and neighbouring stations would always transmit at different times from one another. At regular intervals, therefore, every motorist could be provided with a bulletin containing information about the area through which he was travelling. The motoring receiver could either be separate from the normal car radio or cassette-player, or, for drivers who wished, special receivers could be fitted to either machine so that it could be played in the normal way and would be interrupted by the strongest local traffic signal.

This system is rather similar in principle to a radio paging scheme for national coverage, using one frequency and a cellular transmitter layout, developed by the Martin Marietta Corporation in the USA.

The BBC say they would be willing to develop and operate the service in collaboration with a motoring organization. Cost of establishing the transmitters would be about $£ 21 / 2$ million. The BBC thinks that a project of this nature should not be a charge on the licence fee - the finance would have to come from some other source.

There is a well established system of traffic information broadcasting operating in West Germany, called Auto-fahrer-Rundfunk-Information (see May 1973 and April 1974 issues).

## Rogers packs up

It's sad to record the demise of yet another small British company making high quality products. Rogers Developments (Electronics) Ltd, manufacturers of audio equipment including amplifiers, tuners and monitor loudspeakers, ceased operating on October 31 1975. The company was formed by J. D. Rogers in 1947 and came to considerable prominence during the hi-fi boom. Following changes in the firm's management on June 17, 1975, when the founder was made chairman and technical director, Mr Rogers "was compelled to resign" (as he writes to us) on August 27. At the time of writing we understand the company will be put into liquidation. Farewell to a much respected and friendly firm.

## Licence for Teletext?

In a recent memorandum to the Annan Committee on the future of broadcasting the $B B C$ says it would expect its CEEFAX part of the Teletext service to attract an incremental licence fee in much the same way as colour television does.
Viewdata, the Post Office's television screen information service using the domestic telephone line, will of course be charged for on a subscriber basis (November issue, p.532).

# Wireless World Teletext decoder 

## 3 - Line and clock dividers

by J. F. Daniels

Last month we looked briefly at the overall operation of the decoder by considering a simplified block schematic diagram. I also mentioned that the heart of the circuit lay in the design and operation of the clock and line divider circuits. This month we shall look in greater detail at the operation of these two circuit blocks, as these together provide most of the waveforms used elsewhere in the decoder circuitry. We shall begin by looking at the operation of the line dividers.
The main function of this part of the circuit is to provide row-address information for the store during the operation of writing data into the store and when reading it; i.e., when the data is
displayed as characters on the screen. Other outputs are also obtained from this part of the circuit, namely separated field syncs, line address information for the character generater read-only memory, the vertical component of the display blanking waveform and a field blanking waveform which goes to " 1 " during the period from lines 11 to 21 and 324 to 334 . This latter waveform is used extensively elsewhere in the decoder to

Fig. 1. The line divider circuitry. $\mathrm{IC}_{3}$ is a 74121 monostable, $I C_{5}$ and $I C_{14}$ are 7490 decimal counters, $I C_{11}$ is a 7474 dual D-type edge-triggered flip-flop and $I C_{20}$ is a 74177,/ 74197 binary counter.
discriminate between data acquisition and data display: i.e., data is only "looked for" during the periods just mentioned. If all 625 lines' were examined for data, false framing codes would be detected at random during the video information, and cause lines of "rubbish" to be read into the store. It would of course be possible to look for data only on lines 17-18 and 330-331, but it is possible that broadcasters may change the positions, or increase the number, of data lines transmitted, and so the above method was adopted to avoid later modification in this respect.

It may appear, from a quick glance at the circuits, that the i.cs are numbered somewhat erratically, but this is


because they are designated according to their position on the p.c. boards -$\mathrm{IC}_{1-40}$ being on the uppermost board and $\mathrm{IC}_{41-80}$ on the lower circuit. During circuit descriptions, multigate i.cs will be referred to by their IC number, and output pin number. For example $(7,12$. indicates the inverter gate for field sync pulses in Fig. 1.

## Line dividers

Negative-going mixed syncs from the analogue board trigger monostable 3, which is adjustable by means of a preset potentiometer. Elswhere in the circuit the variable width of the output pulse is used to shift the display horizontally, but for this part of the circuit the output pulse width is not critical, so long as the pulse is wider than the input sync pulse. By gating together the input and output waveforms of this i.c. in a NOR gate ( 4,13 ), the broad pulses during the field blanking interval are separated from the mixed sync waveform. Capacitor $\mathrm{C}_{2}$

Fig. 2. Waveforms from Fig. 1, showing the detection of broad pulses.
removes spikees which are caused by the delay through $\mathrm{IC}_{3}$. The method by which the broad-pulse separation is achieved is more clearly shown by Fig. 2. waveforms (a), (b) and (c).

The broad pulses are used to reset the first part of the line-divider chain, which consists of two i.cs forming a divide-by- 80 circuit. The first i.c., a divide-by-10 counter, provides lineaddress information for the character generater ro.m. and sets the height of each "character box" at 10 lines per field.

In the "upper case only" version of the decoder, only seven of these lines

Fig. 3. Waveforms in the circuit of Fig. 1.
are used for the display of alphanumeric characters, and gates $(6,10)$ and $(4,10)$ serve to inhibit character generation during three of the ten lines. These gates are not required when the upper and lower case version is built, as the blanking of unused character lines is achieved in a later part of the decoder.
The operation of the rest of the circuit is best understood by referring to Fig.3, which shows some of the more important waveforms through the circuit. The divide-by-8 counter $\mathrm{IC}_{14}$ is fed with one pulse every 10 lines from $\mathrm{IC}_{5}$ and its outputs $\mathrm{Q}_{\mathrm{B}}$ and $\mathrm{Q}_{\mathrm{C}}$ are gated in $(13,8)$ to provide a start pulse for the display period at line 51 (also 364). $\mathrm{C}_{4}$ removes any spikes caused by timing differences between the $Q_{B}$ and $Q_{C}$ outputs of the counter, and $C_{5}, R_{1}$ and $R_{2}$ shorten the pulse into a narrow negative spike to set the vertical display-blanking flip-flop, formed from gates $(15,3)$ and $(15,6)$. The positive transition at the output of gate $(15,3)$ is coupled via $C_{6}$ and $R_{3}$ into gates $(12,8)$ and $(6,4)$, which has the


effect of setting the outputs of the 5 -bit counter, formed from ICs 11 and 20, to zero. From this point onwards, the 5 -bit counter counts through the characterdisplay rows until the counter reaches 24 (the end of the page display). This point is detected by gate $(13,3)$ and its output is used to reset the vertical display-blanking flip-flop.
During lines containing data the 5 -bit counter is preset to the correct row number, which is available in binary form at the output of the Hamming corrector during the control and row address groups, at the start of each data line.
The time at which this data is entered into the presettable counter is controlled by gates $(4,4)$ and $(13,6)$. The least significant bit of the row address arrives first, and a strobe pulse from the clock divider circuits (to be described later in this article) set $\mathrm{IC}_{11}$ at the appropriate time. The next data group to arrive contains the remaining four bits of row address information and this is written into $\mathrm{IC}_{20}$, again at a time determined by a strobe pulse from the clock divider circuits. The other inputs to gates $(4,4)$ and (13,6) are waveforms which only allow the strobe pulses to set the counters during lines which have been confirmed as containing valid information. These waveforms are used in a number of places in the decoder and we shall call them data allow (DA), for a waveform going to " 1 " when a framing code is detected and returning to " 0 " at the end of that line and DA for its inverse.

A 3 -input NOR gate ( 12,6 ), provides the field blanking waveforms which goes to "l" during lines $11-21$ (324334), by gating together the $Q_{A}$ and $Q_{C}$ outputs of $\mathrm{IC}_{14}$ together with the vertical display-blanking waveform. It can be seen from the waveform diagram in Fig.3. that this waveform also goes to

Fig. 4. Layouts of typical rows of data bits. A display row (row 5 of magazine 1) is shown at (a) and at (b) is a page-header row.
" 1 " during the equalizing and broad pulses. This is of little consequence, however, as no video is present on these lines so that no false framing codes could be detected here.
Before continuing with a description of the clock divider circuits, it would be as well to consider in greater detail how each of the data rows is made up.

## Data row format

Display row. The data bits are transmitted in groups of eight, and each line of data contains 45 of these groups, or bytes. The first five groups have the same function on all rows and a typical row is shown in Fig. 4(a). The following 40 groups consist of 1 S0-7 coded characters, the eighth bit in each group being used as an odd parity bit. The magazine number and row address information are coded in a different manner from the display data, being more heavily protected in a form of Hamming coded. Only four useful bits of data can be extracted from each of these eight-bit groups, but a single error in the group can be corrected and an even number of errors i.e., 2, 4 or 6 bits in error, can be detected.

Page header row. The page-header row (row zero) contains eight more of these Hamming protected groups following the row address number, and only 32 groups of 1S0-7 coded data for display, as shown in Fig. 4(b). The two groups immediately following the row address number contain the page number units and tens respectively and these together with the three bits allocated to
the magazine number, enable any combination of page numbers between 100 and 899 to be selected.
The next four groups contain information indicating the time of day. The first group contains four bits of "minutes units" information and the second, three bits for the minutes tens. The spare bit in this second group is used as a "clear page" control, which goes to " 1 " whenever the following page contains new information and the old page must be erased from the store. The next group contains a 4 -bit "hours units" message and the fourth group contains a 2-bit "hours tens" indication. There are two unused bits in this group, the first being used to indicate a newsflash page, and the second a subtitle page.

## Clock dividers

There are two more groups of Ham-ming-protected data before the page header display starts, and these will eventually contain more control information.
The clock-divider circuit block provides a large number of pulses and waveforms for use elsewhere in the decoder, as follows: a display lineblanking waveform which goes to " 1 " during the 40 -character-wide display period; the "data allow" waveform which goes " 1 " on detection of a framing code and returns to zero 42 bytes later; 6 -bit column-address for addressing the store, which is used when writing data into the store, and when reading from display; strobe pulses which discriminate between the Hamming-coded words in normal rows and in the page header row and finally, groups of pulses used elsewhere for the formation of the vertical elements of the alphanumeric characters.
Figure 5 shows the clock-divider circuit diagram, and Fig. 6 the associated waveform diagrams.


Fig. 6. Waveforms seen in the circuit of Fig. 5 during data lines.


Write mode．Detected framing code pulses，which are one clock pulse wide （ 144 ns ）and negative－going，are invert－ ed in gate $(22,10)$ and fed into gate $(9,12)$ together with the differentiated output of monostable i7．This is the＂framing－ code allow＂monostable and its differ－ entiated output is about $2 \mu$ s wide and timed to occur at the time of the frame code，ie，about $15 \mu$ s after the start of line syncs．In this way，only valid framing codes are allowed through gate $(9,12)$ ． This method was used to avoid allowing through random framing codes detect－ ed in the colour burst，which can occur with certain types of data－slicing cir－ cuit．The monostable triggering is inhibited by means of the $B$ input on all except lines 11－2l．The valid framing codes are then used to trigger the＂data allow＂flip－flop made up from gates $(9,8)$ and（9，6）．During lines 11－21，the first two clock－divider i．cs are held in the reset condition until the DA－waveform， goes to＂ 1 ＂（ 18,8 and 18,11 ）i．e．，until the frame code occurs．It should be noted that i．cs of the 74 H series are specified for gates 22 and 9 ．Although all the prototype decoders worked satisfactor－ ily with ordinary gates here，it would be possible under worst－case conditions for the time delay between detecting the framing code and removing the reset from the dividers to exceed the maxi－ mum permissible time of 144 ns ．By using 74 H series gates，the time delay will always be less than 144 ns ，even under worst－case conditions of supply voltage and gate delay．


A second binary－to－decimal decoder $\mathrm{IC}_{47}$ is connected to the outputs of the second clock－divider i．c．after they have passed through the four AND gates of $\mathrm{IC}_{2}$ ．These gates serve to allow the outputs of $\mathrm{IC}_{47}$ only during the＂allow＂ waveform at the output of gate $(6,12)$ ． This waveform consists of narrow， positive－going pulses derived from decoder 42，gated with a waveform which only allows the pulses for the first 13 bytes of data．The output pulses of decoder 47 then become narrow，nega－ tive－going pulses，timed to occur one during each byte of data from byte 3 to byte 12：i．e．，decoder 47 output 1 occurs during byte 4 －output 9 during byte 12 ． These pulses are used to time each group of Hamming coded bits in the row and page recognition circuitry，and for presetting the line divider i．cs as mentioned earlier

Read mode．Resetting of the clock dividers in during the＂read＂mode is

The final prototype decoder，which measures about $40 \times 22.5 \times 6 \mathrm{~cm}$ ．The ROLL switch enables pages to be viewed in rapid succession．We understand that a complete kit of parts for the decoder will be available from Catronics Ltd， 39 Pound Street， Carshalton，Surrey．

The newest character code，which is now in use Reference 1 indicates that these undefined control characters will be allocated in order of decreasing serial number．Codes referenced 2 are to assist compatibility with standard data codes．All character rows start in the＂steady，alphanumeric white＂ condition，unboxed and unconcealed unless control characters indicate the contrary．

| Bits |  |  |  | $b_{7}$ $b_{6}$ $b_{5}$ | $\bigcirc 0$ | ${ }^{C} 01$ | ${ }^{0} 10$ | $\begin{array}{ll}0 & 1 \\ & 1\end{array}$ | 100 | ${ }^{1} 01$ | $\begin{array}{lll}1 & 1 & \\ & & \end{array}$ |  |  |
| :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: | :---: |
| b4 |  | b2 |  |  | 0 | 1 | 2 | 3 | 4 | 5 | 6 |  |  |
| 0 | 0 | 0 | 0 | 0 | NUL ${ }^{\text {2 }}$ | DLE ${ }^{(2)}$ | 目 | $\bigcirc$－ | ＠${ }_{\text {＠}}$ | $P$ P | ，$\square^{\text {a }}$ | p | 日 |
| 0 | $\bigcirc$ | 0 | 1 | 1 | ${ }^{\text {Alphan }}$ Red | ${ }_{\text {Graphics }}^{\text {Red }}$ | ＋$\square^{6}$ | 1 1 日 | A A | $Q \quad$ Q | E | व | $\square$ |
| 0 | 0 | 1 | 0 | 2 | Alphan $^{\text {Green }}$ | Graphics Green | 1 四 | 2 E | $B$ B | $R$ R | －1 日 | $r$ | E |
| 0 | 0 | 1 | 1 | 3 | Alphan $^{\text {yellow }}$ | Graphics $\qquad$ | £ 日 | $31 \square$ | C ： C | $s$｜S | I日 | s | － |
| 0 | 1 | 0 | 0 | 4 | ${ }^{\text {Alpho }}{ }^{\text {n }}$ Blue | Grophics ${ }_{\text {Blue }}$ | ¢ | 4 1目 | D ${ }^{\text {D }}$ | T，T | ¢ 18 | 1 | B |
| 0 | 1 | 0 | 1 | 5 | $\mathrm{Al}^{\text {a }}{ }^{\text {anan }}$ Magenta | $\begin{array}{\|c\|} \hline \text { Graphics } \\ \text { Mogenta } \end{array}$ | \％｜回 | 5 目 | E E | $\cup{ }^{\cup}$ | 8 | $u$ | E |
| 0 | 1 | 1 | 0 | 6 | Alphon <br> Cyan | Grophics Cyan | \＆ | $61 P$ | $F$ F | $\checkmark$ V | 들 | $v$ | － |
| 0 | 1 | 1 | 1 | 7 | ${ }^{\text {Alphon }}$ White | $\text { Grophics }{ }_{\text {White }}$ | $\square$ | 7 7 | G ${ }^{\text {G }}$ | $w$ w w | g 1 B | w | － |
| 1 | 0 | 0 | 0 | 8 | Flosh | Concealed Display | 回 | 8 ，E | H H | $\times$ x | ¢ 1 E | $\times$ | $\square$ |
| 1 | 0 | 0 | 1 | 9 | Steady | （1） | E | 9 回 | 111 | $Y$ | E | $y$ | $\square$ |
| 1 | 0 | 1 | 0 | 10 | End Box | （1） | ＊1 8 |  | $\checkmark$ | $z \mid z$ | j E | $z$ | E |
| 1 | 0 | 1 | 1 | 11 | Start box | ESC ${ }^{(2)}$ | $+\square$ | $\square$ | K K | $\leftarrow 1 \leqslant$ | k E | ${ }^{1} 4$ |  |
| 1 | 1 | 0 | 0 | 12 | i | （1） | 1 $\square^{\text {D }}$ | $<$ 回 | L \｜L | 12 1 <br> 12  | $1 \square$ | II |  |
| 1 | 1 | 0 | 1 | 13 | v | （1） | B | $B$ | $M$ ，M | $\rightarrow{ }^{\rightarrow}$ | m $1 \square$ | $3_{4}$ |  |
| 1 | 1 | 1 | 0 | 14 | $\underline{\text { SO }}{ }^{(3)}$ | D | $E$ |  | $\mathrm{N} / \mathrm{N}$ | 414 | n | $\div$ |  |
| 4 | 1 | 1 | 1 | 15 | SI ？ | ${ }^{\prime}$ | $\underline{\square}$ | $\square$ | 0 O 0 | \＃\＃ | $\bigcirc \quad 1$ |  |  |

done by the Q output of monostable 3. This reset is combined with the "write". mode reset by gate $(18,11)$.
The first of the clock-divider i.cs feeds a binary-to-decimal decoder ( $\mathrm{IC}_{42}$ ) of which only eight of the ten outputs are used. Each output is a negativegoing pulse, which occurs at every eighth clock pulse and is used to time the vertical components of the characters. Each character box is eight clock pulses wide, and for the formation of alphanumeric characters, five of these eight-clock-pulse intervals are used, the other three forming the space between characters.

Experiments were carried out with prototype decoders to reduce the character boxes to six pulses wide, with only one pulse space between characters. However, the extra circuit complexity was not thought to be worthwhile, as the display seemed as good with a character-to-space ratio of $5: 3$ as with 5:1. Graphic characters, however, fill the complete character box, in both the vertical and horizontal directions.

Address generator. This is common to "read" and "write" modes, i.cs 11 and 19, together with the inverted $Q_{D}$ output of ${ }_{,} \mathrm{IC}_{1}$, providing the 6 -bit address for the store, obtaining reset by the output of the line blanking flip-flop formed from $(18,3)$ and $(18,6)$. This flip-flop is set by output 3 of decoder 47 and reset at the end of the line, 40 -character boxes later in the display mode, or 40 bytes of data later during the write mode. This same reset pulse is also used to return the data allow waveform to " 0 ". The data allow waveform may also be reset to zero by another input to gate $(9,6)$, taken from the Hamming corrector circuit, which gives a negative-going pulse if an even error is detected during the row address group and inhibits the writing of any data which may occur on such a row. During the read mode of operation the clock dividers and column-address dividers are both reset by the $\overline{\mathrm{Q}}$ output of monostable 3, and by adjusting the width of this pulse, the display can be shifted horizontally to the centre of the TV screen. Also in the read mode, the clock input to the dividers is switched to the output of an adjustable frequency oscillator, to enable the overall width of the display to be adjusted.
This concludes the description of the divider circuits. In the next article we shall look at the operation of the serial-to-parallel converter, framingcode detector, data latches, and data store.

The Teletext decoder will also be usable with Viewdata (November issue, p.532) when this Post Office service becames available. Later in the year we shall be publishing details of the Viewdata system and how it is compatible with Teletext.

# New medium- and long-wave broadcasting plan 

Better long-wave coverage of the UK and better reception of Radio London are two of the few benefits for Britain to be expected from the 1975 Geneva Plan for l.f./m.f. broadcasting just completed under the auspices of the International Telecommunication Union. Due to take effect in three years' time on November 23, 1978, the Plan covers frequency assignments in the 120 m.f. channels available for broadcasting in ITU Regions 1 and 3 (which means all the world except the Americas - see map in August 1974 issue, p.266) and in the 15 l.f. channels available at present for Region 1 only. It supersedes the Copenhagen Plan of 1948 - which was for the European broadcasting area only - and is to last for eleven years. The Plan is in fact a thick book listing stations, frequencies, transmitter powers etc., and is an "Annex" to the "Final Acts of the Regional Administrative l.f./m.f. Broadcasting Conferences (Region 1 and 3)", part of which took place at Geneva in October-November 1974 (Dec. 1974 issue, p. 481 ; Jan. 1975 issue, p.10) and part in OctoberNovember 1975. J. G. Spencer's article "The future of medium- and long-wave broadcasting" in the August 1974 issue explained the technical problems involved in the re-planning.

Britain's improved l.f. coverage will come partly from an additional assignment for $200 \mathrm{kHz}, 50 \mathrm{~kW}$ transmitter at Burghead, Scotland, to be synchronized with the existing 200 kHz station at Droitwich. This will give improved long-wave reception in the north of Scotland. For the central, heavily populated area of Scotland around Glasgow and Edinhurgh there is a further l.f. assignment, for a 227 kHz transmitter at Westerglen, also with a power of 50 kW . (Both Burghead and Westerglen are m.f. stations at present.) Thus England and the Scottish mainland will have complete coverage from two frequencies in the l.f. band. Unfortunately this gain has to be "paid for" in the Plan by the assignment of a new 200 kHz frequency to Warsaw, at 200 kW , for day-time operation. This will not worry us within the UK but will interfere with reception of Droitwich on the Continent. At the moment the BBC has not decided whether the new UK coverage on l.f. will be used for Radio 2 or Radio 4.

Radio London, the m.f. 1457 kHz local radio station at Brookmans Park, will under the new Plan give better reception because a co-channel transmitter at present sited on the coast of Albania at Durres is planned to be moved 60 km inland to Lushnje, where it will lose the benefit of "sea gain", and the night-time interference with Radio London will be reduced by $4-5 \mathrm{~dB}$. Unfortunately some
of the other UK local radio stations are likely to suffer worse co-channel interference at night, particularly Radio Solent at Fareham which may be 10 dB worse off.

Night-time co-channel interference could also be worse for listeners to the Radio $3647 \mathrm{kHz}, 150 \mathrm{~kW}$ transmitter at Daventry, for under the Plan there are three new assignments on this frequency, for stations in Albania, Greece and Libya, and all for 300 kW transmissions. To obtain protection against this interference it will be necessary to increase the power of Daventry, and the UK delegation stated at Geneva that it reserves the right to do this when the new transmitters come on the air. (Theoretically 2 megawatts would be needed but this is unlikely to be achieved because of cost.) These three new assignments in Radio 3's "province" well illustrate the principle adopted by the conference that existing stations should not have priority over new claims for the channels just because they are well established. However, the prospect of next-door neighbours Albania and Greece both transmitting 300 kW on the same frequency seems somewhat unlikely.

Two further problems for the UK from the Plan: the power of Radio Ulster $(1340 \mathrm{kHz})$ will have to be increased from 100 kW to 250 kW to cope with increased co-channel interference power from Budapest; and the power of Radio 4 Wales ( 881 kHz ) will have to be increased from 150 kW to 300 kW to deal with increased co-channel power from East German and Yugoslav stations.
As mentioned in the January 1975 issue, the channel spacing will remain at 9 kHz throughout the l.f. and m.f. bands. The actual frequencies assigned in the Plan at m.f., however, are slightly adjusted to make them integral multiples of the channel spacing; for example Radio 3, Daventry, at present 647 kHz , will become 648 kHz ; Radio Ulster's 1340 kHz will become 1341 kHz ; Radio London's 1457 kHz will become 1458 kHz .
Some commentators have described the 1975 Geneva Plan as a bad plan. "It started as a shopping list and it ended as a shopping list but everyone signed it and called it a plan" were the remarks of one delegate. Criticisms centre around the "power race" (of competitively increasing transmitter powers) and the fact that the Plan carries with it a long and untidy list of 72 reservations and 30 counter-reservations by different countries, claiming rights to change frequencies, powers etc. if they wish to after the Plan has come into force. Perhaps it's as well the Geneva Plan is to last for a maximum of only eleven . years.

# John Logie Baird and the Falkirk Transmitter 

# The activities of the pioneer of television between 1925 and 1928 

by P. Waddell, W. V. Smith and J. Sanderson

Dr. Waddell and Mr Smith are at the University of Strathrlyde. Mr Sanderson is Curator of the Falkirk Museum

On the evening of Tuesday, the twentysixth of January 1926, John Logie Baird demonstrated television to an audience of Members of the Royal Institution. No detailed description, with photographic evidence, of the equipment used for the demonstration seems to exist, although many authors have made guesses ${ }^{1}$. It has been suggested that the instrument now at the South Kensington Science Museum was used, or that a very large disc unit with a single spiral of lenses was seen, but Baird had suddenly become secretive and information is hard to come by. The next two and a half years are poorly recorded and this article is an attempt to discover what Baird was working with during this period.

April 1925 is the beginning of the relevant period in Baird's activities. A. S. Goodchild and R. Best, both employed by Will Day Ltd of Leicester Square, had constructed a television transmitter to Baird's design ${ }^{2}$. The instrument was to be used in the first public demonstration of television, given by Baird at Selfridge's in Oxford Street, and was delivered to him at 22 Frith Street, Soho, in late 1924. This unit was first described in Wireless World ${ }^{3}$ by Baird and some indication of his canniness can be obtained from the fact that he attempted to deceive his readers into thinking he was less advanced than he was. He gave a diagram of a large, basically unbalanced, single-spiral lens disc, which he knew would give a low-grade picture due to a necessarily low speed of rotation and vibration, which would blur the picture. Photographs, and a further article ${ }^{4}$, show that he was really using a double spiral of lenses, which made possible a higher speed and smoother running, and a radially-slotted disc subdivided each of the eight lines into 50 elements to produce a 400 -element picture. (As a comparison, a modern television display contains about 450,000 elements.)

Later in 1925, in June, Television Ltd was formed as a means of financing Baird's research, Day and relatives of Baird supplying much of the capital. Day had also found the Frith Street laboratory and his name appears on


1A. J. L. Baird at 22 Frith Street, Soho, London, Dec. 1924, with his "original" television transmitter and receiver.

1B. J. L. Baird, Sept. 1926, his "original" transmitter now presented to the

Science Museum, Kensington, London. Note the white painted double spiralled receiver disc has been removed, and that a rotary slot has been placed over the photocell, necessitating a change in the chopper disc mounting arrangement.

some patents which he had helped to finance from July 1923, while Baird was at Hastings.
In this way, the work was able to continue at Frith Street and further experiments and demonstrations took place. The face of William Taynton, an office-boy from another floor of the building, was televised by equipment of unknown design and in January 1926 Baird and Capt. O. G. Hutchinson (who had bought Day's share of Television Ltd) demonstrated televised human faces to the Press ${ }^{5,6}$. Then, after the R.I. demonstration ${ }^{7}$, Baird clamped down and details of the equipment used were lost.
Newspaper reports and magazine articles are of some assistance, though words and pictures in some of these are at variance and serve only to cloud the image, many of them not giving details of equipment used. Tracing the development of the equipment, the first reference is found in the Daily Telegraph ${ }^{8}$, in which mention is made of a lensed-disc transmitter, with no further information.
A step forward is indicated by a report in The Engineer ${ }^{9}$ of June 18th 1926 which describes and illustrates a grid being used to subdivide the image, the "grid" being a bundle of optical fibres with which Baird claimed to have eliminated the need for a lensed disc. A full description of this transmitter was not seen until June, $1927^{10}$ and the patent on the use of optical fibres was not applied for until 1927, being obtained the following year - a further instance of Baird's reticence. Discrepancies between the description of the optical fibre system and the photographs of Capt. Hutchinson's televised face exist; the photograph which illustrated a number of articles of this period, showed the usual picture made up of ' vertical strips - a fibre system would have given a cellular appearance. Deception again, perhaps?

In September 1926, Baird gave a double-spiral lens-disc transmitter (two sets of eight lenses) to John Hart of Hart Radio Supplies in Falkirk and, six days later, donated the transmitter/receiver now on display at the South Kensington Science Museum. The authors have found no references to the Falkirk transmitter, except in the Falkirk Museum records. One difference. between this unit and the Science Museum transmitting section is that the Falkirk unit has two, interlaced spirals of eight holes, providing a 16 -line picture, while the Science Museum transmitter has two non-interlaced. spirals, giving an 8 -line picture. Further small differences include the use of three lens-retaining screws for each lens on the Falkirk unit - two for most lenses on the S.M. transmitter. Both units are provided with a rotary spiral slot over the photocell, which increases the number of picture lines, with no increase in the number of lenses. This slot was patented in January 1925,

2. Suggested 32 lensed disc as used at Royal Institution demonstration by $E$. Larner in 1929: book carries a foreword by J. L. Baird.

3. First ever televised photograph, Captain O. G. Hutchinson, business manager and director of Television Ltd and friend of Baird's since 1911 in Glasgow, when they were apprentices together at the Argylle Motor Works, Alexandria, Dunbartonshire. Photograph published in various magazines from June 1926. Photograph has peculiar aspects, one half of the image is apparently a profile, the other half being a frontal view.
although it does not appear in magazine articles or newspaper photographs of the first half of 1925 .
In the first full description of the rotary slot - an article in Electrical Review ${ }^{11}$ - Baird still discusses an 8 -line transmitter and the diagrams in the article match the S.M. transmitter. A second article in December ${ }^{12}$ also shows these features. Then, in 1927, Baird began to talk of 16 -line transmitters, using choppers and rotary slots, as in
the Falkirk and S.M. units. In a paper ${ }^{13}$ read by Baird in March 1928 he describes, in words and in diagrams, a 16 -line transmitter and clearly states "using equipment similar to that shown

I gave a demonstration of true television to the Royal Institution on January 27th, 1926". (Baird always referred to January 27th, not 26th). He also says that a London to Glasgow television demonstration, using G.P.O. lines, was performed on May 24th and 26th 1927 using the same equipment, i.e., the Royal Institution transmitter.

It is interesting to note that in the Falkirk Museum register of entries it states that the Falkirk transmitter was used by Baird to transmit human faces from Glasgow to London on 27th January 1927. Newspaper articles from April 22nd $1927^{14}$ talk of Baird sending his first batches of television signals to New York from London; other articles talk of signals from London, Hull, Leeds, Birmingham and Glasgow to New York. It should be remembered that Baird's publicized transatlantic demonstration was on February 8th 1928, so once again he was ahead of his well publicized public events, making sure nothing would go wrong on the big day.
Another article from March 1928 clearly talks of images from Glasgow to London ${ }^{15}$. A photograph of Baird (Fig. 6) standing beside the Falkirk transmitter, somewhere in the West of Scotland, has been traced. Four people date it to 1928 and almost certainly to Mr Hart's radio show exhibition in Falkirk. The photograph, reproduced in this article, clearly states "radio engineers" and the word "Glasgow" also appears on the board behind Baird's head. The dummy is unlike any seen in published work by Baird. Interestingly enough, one author, P. Waddell, has seen a photograph of Baird wearing the same overcoat on a visit to Scotland, dated January 1927. It should be noted that Baird did visit Scotland in early 1927, arriving in Glasgow on February lst 1927 for his lecture in the St. Andrews Halls, Glasgow on February 3rd 1927.

Other mysterious events concerning Baird at this period and the equipment he was using are seen in the following articles. in Nature, January 15th $1927^{16}$ it was stated that Baird had lectured on January 6th 1927 to the 17th Annual Exhibition of the Physical and Optical Societies at Imperial College, London. A "model" of the transmitting portion of the lecturers "original" apparatus was displayed. Which one? The one from the Science Museum reborrowed, or the one given to J. Hart (still not sent to Falkirk), or a new and third unit? Backing up the third model concept it is clearly stated in a book by S. A. Mosely and H. T. Barton Chapple from $19311^{17}$ that Mr Baird had given a model of his original transmitter to Glasgow University, sometime in early 1927. This is also mentioned in Nature July 1927, and some parts of this transmitter are still in existence.
4. J. L. Baird's "original" 16-lensed disc transmitter and receiver. Note spiral slot over the photocell in the transmitter to increase number of TV lines, without increasing number of disc lenses. In the receiver light from a lamp at $K$ is allowed through rotary slot on disc $H$ then through the lensed disc, to be focused on screen $F$.

It would appear from the above evidence, and from other information we are not at liberty to disclose, that Baird was working in secrecy in Scotland on television well away from his main competitors. The reason for his secrecy from mid 1925 until early 1928 may never be fully known, but why did he never reveal in later years exactly what had been used in January 1926? The authors, having further information, can make educated guesses and assumptions as to Mr Baird's movements at this period.

Baird stated in Radio News, June 1927, that he had abandoned 16-line lensed disc transmitters since his last article in Radio News, which was September 1926. He also had abandoned fibre optics by the time of writing reference 12, i.e. June 1927, and was now investigating multiple-lensed-disc "optical levers". The optical lever enables a rapid scan of the image but only required low speed discs - in general up to five discs were used, four of which had a circle of lenses and not a spiral. The fifth disc could either be a single spiral or an oscillating or notching mirror in order to move the images over the photocell. A full description of such a system is given in a reference ${ }^{18}$. However, in Modern Wireless May $1928^{19}$ it is clearly stated that "they would have to wait and see if it worked.'
The transatlantic TV demonstration by Baird on February 8th 1928 meanwhile had taken place using a lensed-disc transmitter. A photograph in Illustrated London News at the time clearly shows one lens with three screws at $120^{\circ}$ around the lens, but what style of disc, two spirals or one, is not clear. In July $1928^{20}$ Television revealed a photograph of an optical lever using mainly slotted discs and not lensed discs; no large single-spiralled lensed disc can be seen. Shortly afterwards Baird abandoned lensed discs and went over to small aluminium discs having 30 small holes in a single spiral.

Une further mystery remams. Mr Baird donated the so called Royal Institution receiver to the Science Museum in 1930. The receiver consists of a disc having 30 holes in a single spiral and one's immediate reaction is that he must have used a 30 -lensed disc, lenses in a single spiral. Why he left the receiver presentation until 1930 is not known.

Highlighting the utter confusion as to what Baird used in late 1925 into 1926

5. Falkirk Museum transmitter, given by J. L. Baird to "Hart Radio Supplies" of Falkirk on 5th September, 1926. Mr
J. Hart of Falkirk has apparently no

are shown in two references. In a. reference from January $1926^{6}$, Capt. Hutchinson describes the transmitter as having a nine-foot diameter disc with 50 lenses in spiral formation (single or double?) and rotating at a high speed. A
reference from late $1926^{21}$ describes Baird as using a large single-spiralled lensed disc transmitter in February 1926, although they could not be sure since "unpatented devices appear to have been used". A third reference from

September $1926^{22}$ shows a diagrammatic visualization of Baird's equipment at 2TV studios; what appears to be either a large-diameter single spiral or a circle of lenses is shown, followed by the rotary spiral slot over the photocell.

Baird only provisionally patented the double spiral system in late 1926, for an obvious reason. The double spiral with its well-balanced characteristics and small diameter enabled high rotational speeds to be achieved and was a commercial proposition. In the Spring of 1928 Baird started selling the dou-ble-spiralled small aluminium disc "Televisors" at Selfridge's, the spirals being of small holes and not lenses. One

6. Baird with "Falkirk" transmitter (3 screws around each lens) in West of Scotland. Photograph dates to either early 1927 (Jan.) or by witnesses to 1928.

7. Royal Institution demonstration television receiver? Donated in 1930 by J. L. Baird to the Science Museum, Kensington. 30 holes in a single spiral.
spiral acted as a flying-spot transmitter and the other spiral, on the same disc, acted as the receiver. No synchronization problems existed and the units provided a lot of interest for the amateur TV enthusiasts. He was, in fact, creating a television market! So once again Baird in late 1924 with his double-spiralled lensed disc arrangement was $31 / 2$ years ahead of such a system being fully patented.

The authors would be delighted to hear from anyone with photographic or first-hand evidence on the actual equipment used by Mr Baird.
John Logie Baird, with 178 patents in his own name, is under serious study by several people at the moment and will undoubtedly emerge as one of Britain's finest researchers in tackling not only the technical but the business aspects of creating a television audience.

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## Meetings

## MANCHESTER

20th. IEE - "Minicomputers - wherever next" by M. Judd at 17.45 at UMIST, Sackville Street.

28th. IEE - "Advanced developments and miniaturization in the telecommunications field" by J. G. L. Rhodes at 17.45 at Renold Building, UMIST.

## MIDDLESBROUGH

29th. [EE/IMC - "Application of microcomputers to industrial control systems" by J. Gallacher at 18.30 at Cleveland Scientific Institute, Corporation Road.

## NEWCASTLE-ON-TYNE

5th. IEE - "The manufacture of thick-film, precision active filters" by A. W. Greenwell at 18.15 at M4 21 Merz Court, University of Newcastle-onTyne.
12th. IEE - "My Dear Watson" by G. Phillips at 18.30 at The Refectory, The University. Newcastle-on-Tyne, Kings Road.

NEWPORT, I.o.W.
31st. IEE/IERE - "Microwave landing systems' by R. Barrett and J. M. H. Chambers at 19.30 at the Isle of Wight College of Arts and Technology.

## OXFORD

141h. IEE - "More and more on less and less (integrated circuit design)" by P. C. Newman at 19.00 at Oxford Polytechnic, Headington.

## POOLE

14th. IEE - "Hi-Fi" at 19.30 at Arndale Library.

## PRESTON

14th. IEE - "Inertial navigation" by Cmdr. C. J G. Young at 19.30 at Torella Restaurant, 76 Friargate.

20th. IEE - "Reduction in hi-fi noise" by K. J. Gundry at 19.30 at Preston Polytechnic.

RUGBY
20th. IEE - "Cinema engineering" by H. Rigby at 18.15 at Lanchester Polytechnic.

## SHEFFIELD

21st. IEETE - "Electronics in rolling mills" by D Turner at 19.30 at Granville College of Further Education, Granville Road.

## SOUTHAMPTON

21st. IEE/IERE - "Electronics in schools" by G F. Bevis at 19.30 at Southampton College of Technology, East Park Terrace.

## SWINDON

6th. IEE/IERRE - "Automobile electronics" by C. S. Rayner at 18.15 at The College. Regent Circus.

## WEYMOUTH

22nd. IEE - "Data processing and recording" at 18.30 at South Dorset Technical College. Newstead Road.

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# What's in a name? 

## More views on technical terms

By Cathode Ray

As long ago as 1934 (Nov. 23 issue) I took it upon myself to attack, by ridicule, the many absurd and inappropriate "technical terms" that had come into use in the world of wireless. The attack was renewed in 1945 (April issue), "non-linear" (distortion) being one of the targets. It was encouraging to note that by 1954 the British Standards Institution had admitted "non-linearity" to BS. 2065 as an alternative, and by 1955 "non-linear" had been banished from BS.661. Although I have continued to fire at "targets of opportunity" whenever they arose, the fact that an article devoted wholly to misconceived terms has not seemed needful for the last three decades is evidence of an improvement in this respect, though how much (if any) is due to my campaign and how much to a growing sense of logic among electronic engineers I can't say - nor can anyone else.

But seeing (in WW, May 1975, and elsewhere) the word "slew" used in a sense apparently quite unconnected with its actual meaning, has stirred the old instinct once more. I do not, of course, refer to the past tense of "slay", but to a type of movement. (It can alternatively be spelt "slue".) The origin of the word is obscure, but there is no doubt that an essential feature of the movement it refers to is turning or twisting. A sailor might use it to refer to the swinging round of a boom, or even a turning movement of the whole vessel. So it could hardly change in an unchanging direction, as applied, for example, to deflect a cathode ray (you see why I am affronted by this usage!) across an oscilloscope screen. I hope that whoever started misusing this word will kindly explain why he did so. He would make an interesting case history in psychology, no doubt under the heading "Humpty Dumpty complex". (Mr H. Dumpty was, of course, the character who claimed the right to make words mean just what he chose.)

Since I first protested against the use of "ground" (instead of "earth") by non-Americans, and in particular such expressions as "grounded base", or even "earthed base" (instead of "common base") by anybody at all, there has been a marked reduction in the latter offending usage. I won't repeat the whole story; the nub of the matter is that in this context the description "grounded"
or "earthed" is not the essential feature of the type of connection so described. A transistor (or triode valve) used for amplification has a pair of input and a pair of output points, and its performance depends very largely on which of its three electrodes is common to both pairs. The fact that quite often this electrode is also earthed, directly or indirectly, is beside the point; it need not be, in order to qualify as any one of the three possible configurations. It is quite possible for a transistor to have one electrode common and a different one earthed. So the usage complained of is clearly wrong.
Incidentally, it is meaningless to apply a common-electrode distinction to an oscillator, but the Editor considers this too obvious to need explaining. I hope he is right.
My impression is that there has been a noticeable improvement in rejection of such absurdities as "d.c. current", "d.c. voltage", and "i.f. frequency". I wish I could say the same of "at d.c." (meaning "at z.f.") and "d.c. to 10 MHz " (meaning " $0-10 \mathrm{MHz}$ "). Why should frequency be stated in terms of current?
I'm afraid "mixer" (applied to frequency changers) is a lost cause. If this wrong term were abolished, students would be more likely to grasp that just adding together signals of different frequencies (as in the correctly-named audio mixer) does not create other frequencies, any more than just mixing nitrogen and hydrogen produces ammonia.
My only other complaint just now is of a misusage that is far more widely distributed than in technical literature, being spoken by broadcasters and written in Times leaders, to give but two examples. But whereas in these examples it can often be dismissed as mere poetic fancy, in technical contexts ambiguity must be avoided at all costs. It happens that in English there is one word, and one word only, to mean "one or the other but not both". That word is "either". When one wants to include two alternatives there is the correct word "each". Or, if you prefer, "both". As if the choice of these two were not enough, however, certain lewd fellows of the baser sort, as St Luke might well have described them (Acts, ch.17, v.5), perversely reject both of these and see fit to use (or rather misuse) instead the only word that means not both. (Again
the Bible provides a fitting comparison, with the rich man who wantonly took the poor man's only lamb rather than kill one from his own flock. David declared indignantly that this man actually himself - should surely die.) People who say or write "on either side" when what they mean is "on each side" or "on both sides" are guilty of murdering the English language. As a result of this, when one wants to make quite sure that "either" (correctly) is meant there is no way left of doing it. To this day I'm not sure what the writer of an article in Electronics \& Power meant when, describing a new electric train, he wrote that there was a motor unit "at either end". Was he using or misusing the word? Only he could tell, and if he really meant "either" he was not to blame for the failure to communicate his meaning unambiguously; the blame would lie on those who won't use either of the words provided but have to steal the one word available for conveying a certain different meaning.

There are of course quite a number of English words which through ignorance or carelessness are so often wrongly used that one is forced to abandon them altogether for fear that they would probably not be rightly understood. Such words as "infer" (in place of "imply") and "protagonist" (supposing it to be the opposite of "antagonist"). But they are not very relevant to Wireless Warld, so I refrain from further comment.
English is a wonderful language for precise communication - if it is used with due care.

Editor's footnote: Our respect for the English language has to be tempered sometimes by the overriding need for effective communication in a complicated subject. This means that we have to follow common usage in technical language, just as everyone has to in ordinary language. Common usage may not produce the mest elegant or logical expressions but it's what everyone understands. And, of course, it does bring changes in terminology. The term "antenna", for example, is gradually replacing "aerial" in the UK, and although some people may deplore this as an Americanism it is in fact a perfectly good, and descriptive, English word to be found in the Oxford Dictionary. We apologise to readers if our terminology is not always consistent, but we live in a rapidly changing world.


CITIZENS' BAND IN EUROPE

November's World of Amateur Radio mentioned the increased sales of Citizens' Band transceivers in the USA. You may be unaware that Citizens' Band is not limited to the USA. Here in West Germany we have a flourishing CB system and in Scandinavia and several other European countries CB is well established. The German authorities have allocated frequencies from 26.965 to 27.275 MHz at a spacing of 10 kHz , which gives about twenty-six channels, with certain exclusions. Power is limited to two watts input, and two types of transceivers are available, one type being licence and payment free and the other costing DM5 per month (about $£ 1$, but in cost-of-living terms more like 50 p ).
There is a considerable number of $C B$ clubs in Germany, several of them American with USA forces constituting the members. The club to which I belong has roughly 100 members. Basic operating procedure is to allocate one frequency as a "monitoring" frequency and when you contact your party you shift to one of two "conversation" frequencies. Callsigns are generated by the CB-ers themselves. Traffic on CB is totally dissimilar to amateur radio traffic; although there are the keen DX-ers, most people use mobile-tomobile or mobile-to-fixed transceivers for convenience and assistance.

Despite the dire fears of the British Post Office, little interference with television is caused by the licence-free or the registered users, and the Bundespost has gained rather than lost revenue from the CB-ers, who mostly use $C B$ in addition to the telephone rather than instead of it. It does indeed seem a pity that the B.P.O. are so down on CB. Driving through England, the deadly silence of my $C B$ set was a miserable contrast to the many interesting chats I have had with fellow German and American CB-ers in Germany, not to mention the useful information one can get about traffic jams,
fog, snow, and "Smokey the Bear" sitting at the side of the road with his
"Camera Machine"!
Robin Adrian Flood,
Darmstadt,
West Germany.

## Pat Hawker comments:

Although Citizens' Band Radio has become an anathema to very many radio amateurs (we have learned to fear that recruits and frequencies may be spirited away from our side of the fence) it constantly surprises me that there is not a more persistent and vocal demand for CB facilities in the UK. Many countries now have such facilities, and it would surely not be beyond the wit of the Home Office to devise a system that would avoid most of the well-recognised abuses yet still give the British citizen the advantages of two-way radio at low cost.
But in his reference to "Smokey Bear and his camera machine" Robin Flood rather gives the game away - and the Home Office is not likely to be anxious to see a British CB system used for the dubious purpose of giving other motorists advance warning of police speed traps - for that is the meaning of such Smokey Bear messages, and one of the reasons for the current CB boom in the United States. Officially in Germany the CB channels are intended for specific purposes, including police, fire, Red Cross, small businesses, sports clubs, etc. The problem that many countries have found is that once you grant CB facilities there is no easy way to check how they are being used - hence Smokey Bear.

## EXPLOITATION OF VIEWDATA?

I was much interested by the review in your November issue ("Wireless World Teletext decoder" by P. Darrington) of methods which have been used to transmit additional information over a standard television channel.
If they do not already know it, your readers might like to have the discussion of this topic by Prof. Stafford Beer brought to their notice. In his book "Platform for Change" (Wiley 1975, p. 236) he discusses the social and political implications of systems which could, in effect, allow electronic monitoring of what pages of the newspaper an individual chooses to read. There are obvious possibilities for exploitation by advertisers and propagandists.

Stafford Beer does not advocate the abolition of Viewdata and other developments (such as data banks of personal information) which threaten to infringe personal freedom. The potential benefits are enormous if the developments are properly handled. What he argues is. that attention should be given urgently to deciding what constitutes proper
handling and to bringing it about. Systems should be designed, right from the start, so that decisions about access to information can be made according to proper democratic procedures, and enforced.
Your presentation of this series of articles and a design for a decoder is an admirable way of bringing the possibilities to the notice of a responsible section of the community.
A. M. Andrew,

Department of Engineering
and Cybernetics,
University of Reading.
The Government is about to issue a White Paper on the question of personal privacy and the use of data banks and other electronic systems. - Ed.

## AUDIO AMPLIFIER LOAD SPECIFICATION

Now that voltage-dependent current limiters are in widespread use, it is important that users and reviewers should be aware that such circuits, when made too potent, can severely restrict' the output-voltage-producing capability of an amplifier when feeding partially-reactive loads of the type presented by normal loudspeakers. I find myself in agreement with every word of Peter Walker's letter in the December issue, and I think the only controversial point is whether his practical criterion, that a good amplifier should be capable of feeding a load of $\mathrm{R} \pm \mathrm{jX}$, without increase in distortion, is the most sensible choice of criterion. In the absence of detailed information on the complex impedances of numerous commercial loudspeakers, it does, indeed, seem to me to be a reasonable and sensible compromise.
It is really commercial pressures and trends that have given rise to the widespread use of voltage-dependent current limiting, for it enables a manufacturer, with little increase in cost, to increase substantially the (resistive load) power rating of an amplifier, without sacrificing reliability. Some-' times it is deemed to permit the use of faster, though less robust, output transistors. Though this may lead to impressive-looking figures and squarewave photographs in reviews, it may well represent a step in the wrong direction nevertheless, for there will inevitably have been some sacrifice in reliability, and excessively wide bandwidths are in any case quite pointless in an audio amplifier.
It is an unfortunate, but quite inherent, feature of voltage-dependent current limiters, that they tend to make the audible effects of overloading, with some types of load impedance, much more pleasant. This is because the limiter, while operating, gives the amplifier a negative incremental output resistance - an increase in the collec-
tor-to-emitter voltage of the "on" transistor reduces the current it can turn on. If the load has an impedance at some frequency of the same nature as that of a tuned circuit, oscillation will ocicur while the limiter is acting, giving unpleasant squittery sounds.

There is much to be said for making nice simple, robust amplifiers, without voltage-dependent current limiters, even though, for a given price, these may offer less spectacular power-rating and bandwidth figures; but commercial pressures tend to be against this philosophy.

It has sometimes been said that, even in the absence of voltage-dependent current limiters, reactive, or partiallyreactive, loads are "difficult" for amplifiers to cope with, and are liable to give feedback distortion. In a well-designed and conservatively-rated feedback amplifier, without limiter, even a fully reactive load gives rise to no particular distortion difficulties and is, indeed, likely to give less distortion than does a purely resistive load of the same value. I gave the results of some measurements verifying this theoreti-cally-predictable result in a letter to the editor in January 1955. If the distortion does increase on reactive loads, it is usually because the feedback stability margin is insufficient, giving oscillation, or near-oscillation, during part of the cycle. The transient testing technique described by $R$. C. Bowes in the December 1962 issue, applied with loads of appropriate phase angles, is helpful here.
Peter J. Baxandall,
Malvern,
Worcs.

## "THE CONSULTANTS"

It is clear from John Dwyer's very comprehensive article on consultants that he invested a considerable amount of time in research and discussions with individuals in the industry. Unfortunately, the very range of his discussions seems to have led to some quoted comments being abbreviated to the point where a degree of distortion has resulted. Since one of the things which emerges strongly from his article is the controversy surrounding the role and relevance of consultants, it is important to make sure that the complex issues involved are not muddied by incorrect attributions (November issue).
Mr Dwyer devoted considerable space to the very important question of whether learning by the consultant is carried out at the client's expense or his own. How you tackle this question depends partly upon what is included as "learning". In routine situations a consultant should be fully briefed already. If he is not, he should most certainly acquire the necessary skill and background in his own time and not at the client's expense. This point hardly
needs stating. However, my comment, which he paraphrased as "if you're working for the government they don't mind your learning at their expense" was made to emphasise a particular point. This was that a different type of learning is involved when a client, which may often be a government agency, retains a consultant to carry out a specific type of work for him, involving the breaking of new ground. Examples include the development of a new technology or provision of an independent assessment of an existing technology unbiased by any preconceptions. In such cases familiarisation with the technology and so development of the consultant - learning are inevitable consequences of the work which the client is sponsoring. He himself is well aware of this and happy that it should be so.
John Dwyer also implies that the receivership (not bankruptcy) of the group to which Cambridge Consultants belonged may still be having its effect: this anxiety could be quickly dispelled by considering our financial and technical success since that time. However, a consequence of that experience has been to make us at CCL crucially aware of the essential need for really solid product development and production engineering if a product is to be successful in the market. This is a very important lesson which we have not forgotten and I regret that John Dwyer seems to have done so by mislaying the essential point of my comment on that topic!
My comment he quoted on the section reproduced from my article in Communications International (referring to the accessibility of consultants) was that specialist skills which a client may need may often exist within the large corporations, such as Marconi. However, these skills often cannot be tapped from outside the organisation. They may not even be available to other members of the corporation, because of competing demands for the time of the specialists involved. In such cases, consultants with the appropriate skills can provide a source of the necessary know-how which is accessible to all.
I fear that the very comprehensiveness of the article, covering the range of companies, establishments and styles which it did, will not have given a possible user of consultants much guidance or help in deciding whether a consultant might be of help to him personally, and if so whom to select.
My own views on this are that use of a consultant can be indicated if the development is short term; if internal development effort or appropriate skills are not available, or if long term commitments resulting from the recruitment of new permanent staff are unacceptable.
The question of whom to select to undertake the work is more difficult. Clearly, in addition to personal competence to carry out the work, a consul-
tant should also have adequate equip ment and resources to ensure that time is not wasted as a result of a half-baked approach. But in addition, and equally important, he should have a clear and detailed 'perception of the client's needs and proven ability to deliver results efficiently. This requires much experience of managing development programmes. This is a difficult and absolutely essential art and did not receive, in my view, nearly enough exposure in an article which paid attention to so many other aspects. It is management ability which really distinguishes and singles out the competent contract research and development activity and failures in this same area which have given rise to perhaps the most cogent criticisms of consultants quoted in John Dwyer's article. Effective management and control of projects is of equal importance to technical competence and this fact can hardly be over-emphasised.
R. J. Smith-Saville,

Cambridge Consultants Ltd, Cambridge.

## DOPPLER DISTORTION AGAIN

In the September issue Mr Coleman raises the question of the audibility of Doppler distortion in loudspeakers and feels that it is rather surprising that the listeners should be able to recognise "such subtly camouflaged distortions." I rather take the opposite view: "how can they fail to recognise them?" Reference to the test results quoted on page 67 of the April, 1974 issue show that when using an 8 inch speaker the Doppler distortions are some 16 dB higher than the amplitude intermodulation distortions. As the distortions components arising from 'Doppler and amplitude intermodulation are at exactly the same frequencies it is reasonable to believe that the Doppler distortion will be more significant when the distortions are judged subjectively. I believe that the "dirty reproduction" generally thought to be due to amplitude intermodulation is largely due to Doppler distortion.

It may only be coincidence but if one notes the order of merit for the seven pairs of high quality loudspeakers recently panel tested by Hi-Fi News, it will be seen that there is a significant measure of agreement between the measured Doppler distortions and the listening panel opinion about the quality of reproduced music.

Mr Coleman's suggestion that room effects may make Doppler distortion more obvious is well based. When reproducing two test tones in a room, head positions can be found where the distortions are either very audible or almost inaudible, particularly if one ear
is closed off. However these two tone tests would appear to have little relevance to the reproduction of music, for we then use two ears and are listening to signals which have a continuously changing spectra. If these room effects are of any significance no form of acoustic treatment will have any effect.
Some months ago "Cathode Ray" suggested that Doppler distortion should occur in air, even if separate loudspeakers are used for reproduction of the h.f. and l.f. signals. This is to be expected and during some other investigations we made a check on the effect. The equipment we have available will measure Doppler components having amplitudes down to about 75 dB below the test tones. The h.f. and 1.f. signals were provided by separate oscillators and reproduced by separate speakers but we could find no evidence of Doppler distortion with any of the half dozen speaker configurations we checked. These tests do not prove that it does not exist, but suggest that any Doppler distortion that is produced due to interactions in the air is not of practical significance.
James Moir,
James Moir \& Associates,
Chipperfield,
Herts.

RESISTANCE
COMPARATOR
1 did not have an opportunity to comment at the time of publication in November Letters of Mr Sandman's contribution to the quest for a minimal resistance comparator. Assuming there was not some awful draughting error in reproduction, then I would note that with equal reference $R_{f}$ and unknown resistor $R_{x}$ the diagram shows his output meter would indicate $80 \%$ f.s.d. with normal meter resistance and for $R_{x}=2 R_{f}$ (100\% over-range as covered in the Griffiths-Choi scheme) the output volts would be 20 V and the meter would be trying vainly to read $160 \%$ f.s.d.

Moreover, he makes no provision for backing-off the standing output so that small deviations ( $1 \%$ ) from nominal can be displayed as full scale deflections and even if he had done this he would find that $1 \%$ changes in $R_{x}$ would cause only 0.1 V change in f.s. output. This would entail reducing his meter series resistor to zero and one would then see the full temperature coefficient of the coil resistance with its effect on the calibration, although this would be a small matter compared to the devastating 25 mA that his op-amp (741?) would pump into the meter if the unknown resistor were disconnected. I can not readily imagine the situation in which Mr Sandman's circuit would be useful. D. Griffiths, Imperial College,
London SW7.

## ELECTRODYNAMICALLY INDUCED E.M.F.

I should like to make some comments on the points made by C. S. Evans (Letters, October issue). Taking first the issue of magnetic screening, it is clear that it would not work. I suspect that the reason for the confusion is a misapplication of the concept of lines of flux. Flux lines can move through the screening material - they don't become tangled up in it.

The answer is seen more clearly if the true nature of magnetic screening is recognised. An opposing magnetic field is developed in the screening material and this almost cancels the original field at the wire. When the aircraft is in motion, there is no induced e.m.f. arising from the magnetic field of the screen because the velocity of the conductor relative to the field of the screen is zero. Since the conductor is moving relative to the earth's field an e.m.f. is induced just as if the screen were not present.

The second method suggested wouldn't work either. As Mr Evans correctly states, the insulated wire is in no way shielded from the magnetic field. Therefore an e.m.f. is induced in it, opposing the e.m.f. developed in the wings. The potential difference between the wire ends depends, in a uniform field, only on the distance between the ends of the wires and not on the wing span.
Colin R. Masson,
Darwin College,
Cambridge.

## VANISHING COMPONENT SHOPS

1 was very interested in Mr Pethers's letter in the November 1975 issue. We are $90 \%$ mail order but we do carry out a shop trade during normal business hours. Mr Pethers's criticism of lack of technical staff could, I know, be levelled at us also. On Saturdays we have several competent people to answer queries but during the week we can only supply people who can find what they want in our catalogue and give one of the mail order staff a list of their requirements.

The reasons we cannot supply technical staff all the time are these. The rewards in the retail component trade are so meagre that we could not afford to pay a technical assistant, even in the event of finding anyone suitable (that in itself being an almost impossible task). The last straw was, of course, the multi-rate of VAT, zero, $8 \%$ and $25 \%$, which has turned our job into a nightmare. As I have stated many times, we only carry on because we like our customers and enjoy electronics, and look forward to our reward in a land
where components are solid gold and speaker cabinets packed with cloud! A. Sproxton,

Home Radio (Components) Ltd,
Mitcham, Surrey.

## INSTRUMENT READ-OUT IN BRAILLE

I refer to J. M. Osborne's letter, (July 1975 issue) concerning the possibilities of Braille digital meters. There is little problem concerning the digital decoder for b.c.d. to Braille, and the circuit shown using five s.s.i. chips will do the $j o b$ at about $£ 1$ per digit. In addition the scheme can easily be extended, if desired, to read scale factor, units and decimal point. For higher volumes a p.l.a. would be more compact.


The problem lies in constructing a suitable Braille transducer. The obvious solution of four solenoid/lever arrangements per digit would do the job, but on the grounds of size and power consumption a more sophisticated solution involving perhaps hydraulics would lead to a more satisfactory solution. If any reader has any ideas on the matter or who has knowledge of a commercial Braille readout I would be glad to hear from liim.
S. J. Cahill,

Ulster College, The Northern Ireland Polytechnic, Jordanstown Newtownabbey, Co. Antrim.

# Television tuner design - 4 

## Alignment and use of coupler

by D. C. Read, B.Sc.

These instructions cover all aspects of adjustment and even though most of the operations described are simple to carry out - many are virtually standard practice for equipment of this type - the same degree of detail is maintained throughout because the listed sequence is a step-by-step diary of actions taken and measurements made when liningup two tuners to this design (one with a varicap front end and the other fitted experimentally with vaned-capacitor tuner). Hence, the instructions given are all essential and practical and, if followed closely, should lead to a circuit performance well within tolerance.

1. Connect +24 V to the board and measure the total current delivered which should be about' 140 mA . If the on-board regulator is fitted, the supply to this should be at 30 V nominal.
2. Using an Avometer Model 8, or voltmeter of similar sensitivity, check that the d.c. values measured through the circuit agree with those listed in the Table. Ensure particularly that

- the separate u.h.f. tuner supply feeds taken through $D_{1}$ and $D_{2}, D_{3}$ are both at $12 \pm 0.7 \mathrm{~V}$.
- the tuning voltage supply at the collector of $\mathrm{Tr}_{20}$ is +11 V (within the zener tolerance). If, as mentioned in step 11(a), one of the higher-numbered television channels is required for selection, a supply voltage of +16.5 or +22 may be derived by adding extra zeners to the $D_{12} / D_{13}$ circuit.

3. Set adjustable resistors $R_{17}$ if fitted, $\mathrm{R}_{19}$ and $\mathrm{R}_{83}$ to mid-range; set $\mathrm{R}_{78}$ fully counter-clockwise to give about 15.5 V on the slider.
4. Disconnect one end of $R_{74}$ (connected to the junction of $\mathrm{C}_{65}, \mathrm{R}_{75}$ and $\mathrm{R}_{76}$ ) to disable the a.g.c. circuit - but make sure that the wire link feeding a.g.c. signals to the i.f. stages is in position. Thus the control voltages fed to the u.h.f. tuner module and to the $\mathrm{Tr}_{2} / \mathrm{Tr}_{3}$ i.f. amplifier are basically set by the fixed values of $\mathrm{R}_{85} / \mathrm{R}_{86}$ and $\mathrm{R}_{80} / \mathrm{R}_{81}$. (Subsequently, the full range of gain control is tested by connecting three extra resistors to the a.g.c. circuit so as to form a temporary adjustable divider as shown dotted in Fig. 2.)

For i.f. alignment, start with an a.g.c. feed of +4.5 V as measured at the injection link wire (D in Fig. 2); this gives a gain reduction of about 12 dB as


Fig. 19. Oscilloscope test probes: (left) "twin-legged" (Hewlett Packard) type; (next) home-made version constructed from "standard" probe. The
two other examples - with earth
leads coiled round the probe body - are simpler usable alternatives. Care in the use of probes is necessary in the 30-40 MHz region to avoid measurement error and confusing radiations.


Fig. 20. Sweep response of $\mathrm{Tr}_{9}$ output showing sound i.f. band.
shown in Fig. 13 of Part 2. Note that an increase of a.g.c. voltage results in a decrease of main-circuit gain because the transistor current gain falls as the collector current rises.
5. Disconnect the wire link feeding the incoming signal to $\mathrm{Tr}_{1}$ base. Inject a signal from a sweep generator directly into $\mathrm{Tr}_{1}$ via small ( 2.2 nF ) disc ceramic capacitor connected to the base and with a 75 -ohm terminating resistor across the end of the generator outlet cable having one end soldered to the earth plane of the board. Remove $C_{23}$ from the if. output test pins; feed the output from across $C_{22}$ to the display oscilloscope.

It is essential in this operation that the oscilloscope probe should not cause too much change in circuit constants normally obtaining at the measuring point, neither should it distort the observed response by virtue of added reactance in the output test feed-line itself. Ideally, a two-legged probe such as that provided with Hewlett-Packard equipment - see Fig. 19 (top) - could be used; failing this, it is worthwhile constructing a home-made version, using a standard type of probe without the clip attachment and modified as suggested by the second example shown in Fig. 19. The two other lower examples show probes with their earth leads round the body; these would be usable alternatives.

The injected signal sweep should be from about 29 to 44 MHz , preferably having the swept range set so that the horizontal scale on the oscilloscope graticule fits in with the important frequencies in the anticipated pass band. Typically, this would require having the sound carrier at 33.5 MHz spaced two divisions left of centre and the vision carrier at 39.5 MHz two divisions right of centre. The response oscillograms in part 2 have such a spacing. Set the signal amplitude at $\mathrm{Tr}_{1}$ base to give an output r.f. envelope measured across $\mathrm{C}_{22}$ (at the MC1330P input) of 100 to 120 mV pk-pk.
6. Adjust $L_{4}$ and $L_{6}$ to obtain a response shape which is shown in Fig. 4 of part 2; Fig. 3 was obtained with the sound
trap/sound take-off disconnected. Adjusting $L_{4}$ alters the position of the pass-band (tuned so that the required range of frequencies is covered) and $L_{6}$ controls the in-band shape (set so that the double-humped response is as flat and level as possible).
7. Adjust the $L_{19} / L_{20} / L_{21}$ circuit as follows. The drawing of this inductor assembly (Fig. l5) shows a 2 -mm gap between the upper coil, $\mathrm{L}_{19}$, and the overwound pair, $\mathrm{L}_{20} / \mathrm{L}_{21}$. This is a typical spacing for the required degree of coupling but because it is a critical factor in circuit operation it may need slight alteration, possibly making use of the sliding paper tube arrangement mentioned earlier.

It is essential to use a non-metallic tool (e.g. a filed-down knitting needle). when making core adjustments otherwise there will be a marked change in notch response each time the tool is withdrawn and it will be virtually impossible to achieve the necessary tuning accuracy. For similar reasons, the shielding cans must be fitted and screwed down in position before any adjustment is made.

To start the procedure, initially fit only the lower tuning core. Observing the sweep output as in step 5 , use this core to set the tuning of $L_{20} / L_{21}$ so as to obtain a dip in the response at 33.5 MHz . Now add the second core, screwing it only a short way into the stack until the $\mathrm{L}_{19}$ circuit resonates to give the deep narrow response null which is created by phase cancellation between the two circuits. Next, readjust the lower ( $\mathrm{L}_{20} / \mathrm{L}_{21}$ ) core to find two such null points which, typically, should occur within $3 / 4$ turn of the core and be separated by about 300 to 500 kHz ; this double-null condition shows, in fact, that optimum coupling exists. (If only one null is obtained - which may be quite shallow - this indicates overcoupling, and the coil spacing of 2 mm must be increased slightly; alternatively, a decrease in spacing is desirable if the two null points are more than 500 kHz apart.)
$\mathrm{L}_{20} / \mathrm{L}_{21}$ core is then screwed out to selëct the null at the higher frequency (and which is further identified by having a noticeably steeper slope towards the video pass band as illustrated by the response shapes in Figs 5, 6 and 7 of part 2). Finally, move the two cores simultaneously and very gently together along the former, maintaining the shape of this steep-sided notch, until the null point is set exactly at 33.5 MHz .
8. Loosely couple the signal from the sweep generator to the i.f. injection test point (drawn in Fig. 2 as an unconnected pin to the left of the u.h.f. tuner i.f. output connection) by putting a plastic sleeve over the test point and winding two or three turns of wire round it with one end soldered to the

Test conditions

inner of the generator output co-axial cable. The braid of the cable is connected to the board earth plane, and the generator is terminated by a 75 -ohm resistor between the cable inner and the earth plane.

Replace the link to $\operatorname{Tr}_{1}$ base which was previously disconnected. With the same sweep range and output across $C_{22}$ established in step 5 - preferably adjusting the generator output if necessary to obtain the required 100 to 120 mV pk-pk signal amplitude specified, and leaving the temporary a.g.c. divider variable at the 4.5 V setting given in step 4 - adjust the u.h.f. module tuning slug for pass-band position and the core of $L_{2}$ for band shape to obtain a response as illustrated in Fig. 5 of Part 2. Here, it may be advisable to temporarily tune out the adjacent-vision trap $\left(\mathrm{L}_{1}\right)$ and the adjacent-sound trap $\left(\mathrm{L}_{3}\right)$ so that the effect of these circuits can be disregarded when making the main passband adjustments.

To do this, tune $L_{1}$ well below 31.5 MHz and $\mathrm{L}_{3}$ well above 41.5 MHz ; after making the main-circuit setting, re-tune $L_{1}$ and $L_{3}$ to give notches at these frequencies. Observe that $L_{3}$, particularly, has an effect on the pass-band shape itself; in fact, the $\mathrm{L}_{3}$ response skirt completes the shaping at the 6 dB -down region around the wanted vision carrier at 39.5 MHz . Ideally, the pass-band response here should be symmetrical about the carrier frequency such that the curvature towards the notch is an inverted mirror image of the curvature towards the flat part of the pass-band - the 39.5 MHz point being at the 6 dB -down level; this is, of course, a particular requirement of the vestigial side-band reception system as specified by the idealised curve of Fig. 8.
9. With the sweep test signal still being fed at the same level as in step 8 via the u.h.f. tuner i.f. injection test point coupling, disconnect $C_{44}$ from $\mathrm{Tr}_{9}$ collector and transfer the oscilloscope probe to between $\mathrm{Tr}_{9}$ collector and earth; the oscilloscope must be operated in the a.c.-coupled mode for this next measurement. Remember to re-connect $C_{23}$ across $C_{22}$, and note that the value of the replaced component must be as nearly as possible the same as the probe capacitance, typically 10 pF , if the circuit conditions as indicated on the oscilloscope are to remain undisturbed.

Connect a 20 pF capacitor across $\mathrm{L}_{16}$ (this, added to the 10 pF or so of the probe, gives the tuning capacitance required for $\mathrm{L}_{16}$ ). Fig. 20 shows the response shape which should be obtained at this point. The 33.5 MHz sound carrier is at the centre of a 1 MHz pass-band (only $\pm 50 \mathrm{kHz}$ of this being occupied by the sound-modulation side-bands, the remainder of the bandwidth is needed to develop an a.f.c. voltage of sufficient range). Adjust $\mathrm{L}_{14}$, $L_{15}$ and $L_{16}$ to obtain this response. $L_{14}$ and $L_{15}$ together produce a lop-sided sound i.f. pass-band shape - with


Fig. 21. Sound/a.f.c. discriminator $S$ curve at $\mathrm{L}_{18}$ output test point.


Fig. 22. Insertion test signal (i.t.s.) waveforms as generated (top), and as received showing added distortion (bottom).
greater response at low frequencies which counteracts the slope of the main i.f. circuits which affect the signal taken from $\mathrm{L}_{19}$ and $\mathrm{L}_{20} ; \mathrm{L}_{16}$ fills in the dip around the sound carrier frequency at the pass-band centre. The dip is caused by stray-capacitance top-C coupling of the $\mathrm{L}_{14}$ and $\mathrm{L}_{15}$ circuits. In practice, the response shape in this region is relatively unimportant because signals within the band are clipped by the limiting action of $\mathrm{Tr}_{12}$ and $\mathrm{Tr}_{11}$. The energy transferred by the circuit should be equally distributed on either side of the carrier frequency and this requires that the sides and skirts of the response curve, not the peaks, should be as symmetrical as possible about the carrier point.
10. Remove the added capacitor and the probe. Re-connect $\mathrm{C}_{44}$ to establish the circuit as in Fig. 2. Connect the probe via $R_{103}$ and the test point provided to the junction of $R_{54}$ and pin 6 of the $L_{18}$ assembly to observe the discriminator response curve which ultimately should appear as in Fig. 21. To obtain this shape, first adjust $L_{18}$ to set the S-curve centre exactly at 33.5 MHz ; accurate adjustment here is necessary because it determines the frequency point about which the a.f.c./tuner combination settles. Second, adjust $\mathrm{L}_{17}$ and $\mathrm{L}_{16}$ to obtain a symmetrical S-shape. The high peak-peak voltage of the curve, as illustrated in Fig. 21 ( 15 to 20 V ), is
needed to give a sufficient amplitude of recovered audio - taken from the 100 kHz middle region - and a wide a.f.c. range.

Beware of a possible measurement error which can occur and which might not be detected without advance warning. The error is caused by the phase-shifting effect at l.f. given by many oscilloscope a.c.-coupled X-drive circuits; it does not occur therefore, if the oscilloscope and the generator are d.c.-coupled throughout. If the applied frequency sweep is at a slow rate, and of . saw-tooth form with a fast fly-back stroke, the displayed curve is effectively delayed along the frequency axis by the coupling and will therefore be artificially displaced with respect to either second-trace market pips or the graticule scale, whichever is being used for measurement. This means that although the S-curve may apparently be set with its zero-crossing exactly at 33.5 MHz , the actual setting will be different. In more comprehensive sweep generators. the sweep rate is arranged to be the same in both directions, and this usually results in a double response display showing two $S$-curves which are displaced by an amount which increases as the sweep rate decreases. The technique for correct adjustment using such equipment is to estimate a mean between the two curves and set the circuit accordingly. Failing this, i.e. when using one of the simpler types of sweep generator and a.c.-coupled oscilloscope, set the displayed curve symmetrically about the indicated carrier frequency - accepting any error which this introduces - and subsequently make compensating adjustment when setting up the a.f.c. system as described in step 20. Further small adjustment will be needed in any case at that point to compensate for the offset voltage created by the input biasing current to the SN72741P op-amp.
11. The information originally given in this step was transferred to Part 3 of the article where it appears in the section headed "Tuning-voltage supply arrangements". Also, Fig. 21 became Fig. 17.
12. Disconnect the oscilloscope probe from the discriminator ( $\mathrm{R}_{103}$ ) test point and transfer it to the $I C_{1}$ pin- 4 test point, which is fed via a $1 \mathrm{k} \Omega$ resistor as a protection because, if pin 4 is accidentally earthed, the i.c. would almost certainly be damaged beyond repair. Connect a suitable u.h.f. aerial to the tuner input. The display required is that of the i.ts. (insertion test signal) waveform which is included in the television signal radiated by all stations in the U.K. on lines 19/332 and 20/333 of the field-blanking interval. The top trace of Fig. 22 shows this waveform as generated and the bottom trace is the received signal with distortion added by the transmitter and the tuner.

At this point, i.e. with the a.f.c. system inoperative, it may be necessary to
check the adjustment of the appropriate pre-set control to obtain a satisfactory display. There may also be some difficulty in obtaining a stable trace of the i.t.s. waveform. If so, a suitable triggering signal for the oscilloscope can be obtained from the sync separator circuit formed by $\operatorname{Tr}_{13}$ and $\mathrm{Tr}_{14}$; from the last-mentioned, a feed of inverted mixed sync pulses is made available via $\mathrm{R}_{105}$ which protects $\mathrm{Tr}_{14}$ from accidental damage. If the set-up procedure detailed here is being carried out at a time when a test card transmission is available, there should be no problem in obtaining a stable trace by means of the oscilloscope internal trigger signal.

As the a.g.c. circuit is not working, in addition to the pre-set tuning adjustment just mentioned, it will be necessary to set the tuner gain using the temporary divider circuit installed in step 4. To do this, first check that $R_{78}$ is set to a minimum, ensuring that $\mathrm{Tr}_{17}$ is not conducting, then adjust the divider variable to give about 4 V at the i.f. a.g.c. injection point. It may also be worthwhile adjusting $\mathrm{R}_{83}$, the r.f. a.g.c. pre-set control, to obtain optimum noise performance. Finally, by manually setting the tuning voltage, it should be possible to fit the received sound carrier precisely in the notch created by the $\mathrm{L}_{19} / \mathrm{L}_{20} / \mathrm{L}_{21}$ circuit; if this is not done the displayed waveform will be obscured by a superimposed 6 MHz fuzz.
13. The MC1330P ( $\mathrm{IC}_{1}$ ) tank circuit inductor $L_{7}$ must initially be adjusted to obtain maximum amplitude of the displayed signal and then more precisely for best possible linearity of the overall transfer characteristic by observing the 6 -step staircase waveform of the i.t.s. (Because each stairstep carries a block of colour subcarrier, it is necessary to imagine a sawtooth line running diagonally through each block; and make adjustments until this line is as straight as possible.)

Observe also the $10 \mu$ s rectangular pulse to white level - the so-called luminance bar at the beginning of the line; an added criterion in the adjustment of $L_{7}$ is that overshoots should occur only at the bottom-left and top-right corners of the bar (compare the form of upper and lower traces in Fig. 22). Overshoot at these points is caused by h.f. signal components arriving early with respect to those at l.f. and is also the reason for the ringing in front of the $0.2 \mu \mathrm{~s}(=2 \mathrm{~T})$ sine-squared pulse which follows the bar; the effect is shown in more detail by the expanded waveform in Fig. 23. Correction for this type of distortion is obtained by adjustments to be made in step 15.
14. Transfer the oscilloscope probe to the base of $\mathrm{Tr}_{4}$ and then tune $\mathrm{L}_{8}$ for minimum 6 MHz "fuzz". To do this effectively, it is helpful to de-tune the receiver slightly, using $\mathrm{R}_{93}$ or $\mathrm{R}_{96}$ as appropriate, to temporarily upset the notching conditions mentioned at the


Fig. 23. Detail of $2 T$ and $10 T$ i.t.s. pulses showing effect of gain and delay inequality distortion.
end of $\overline{\text { step }} \overline{1} \overline{1}$; the displayed signal then contains a larger sound-carrier component which makes it easier to check the effect of tuning $L_{8}$.
15. Connect the probe to the output of the video strip, i.e. to the 75 ohm output terminal. This will allow adjustment of the group-delay equalizer. The curves in Fig. 11 show that in this circuit the first stage $\mathrm{L}_{10} / \mathrm{L}_{11}$ corrects for delay difference between the lower and middle video frequencies whereas the $L_{12} / L_{13}$ stage has its main effect at the top end of the band. Adjustment of the two pairs of inductors is best carried out by inspection of the first i.t.s.-line waveforms. Initially, $\mathrm{L}_{10}$ and $\mathrm{L}_{11}$ are tuned for minimum overshoot on the luminance , bar corners (see step 13), and to reduce ringing before and after the 2 T pulse. Figs. 24(a) and 24(b) are expanded i.t.s. waveform traces which, in comparison with Fig. 23, illustrate the improvements in delay response typically effected by these adjustments. Two oscillograms are given - taken with the same tuner tuned to different channels - to show that the transmitted signal itself has differing imperfections such as can be seen in the examples on inspection of the trace shape around the bar corners and on either side of the $2 T$ pulse.

Set the $L_{12} / L_{13}$ stage for best symmetry of the $1 \mu \mathrm{~s}(=10 \mathrm{~T})$ chrominance plus - luminance pulse, between the 2 T pulse and the staircase in the waveform, illustrated in more detail by the expanded traces of Figs. 24(a) and 24(b).
When adjusting for symmetry, ensure that the pulse base is as flat as possible; specifically, $\mathrm{L}_{12}$ and $\mathrm{L}_{13}$ are tuned so that any ripples present in the base line (i.e. departures from the blanking-level line as seen in the Fig. 23 example) are reduced in amplitude, and those remaining are disposed symmetrically about the pulse-base centre, as in Figs. 24(a) and 24(b). Having achieved best symmetry, even better flatness can sometimes be obtained by adjusting the variable resistor in the series-tuned circuit which is optionally connected across $\mathrm{R}_{25}$ in the $\mathrm{Tr}_{4} / \mathrm{Tr}_{5}$ amplifier stage; an improvement of 1 to 2 dB in chro-
minance-luminance gain inequality which shows up as a symmetrical bump or dip in the pulse base - should be possible.
16. It may be necessary to adjust $L_{9}$ to reduce the amount of colour subcarrier second harmonic ( 8.866 MHz ) present in the tuner output. Do this by inspecting the $14 \mu \mathrm{~s}$ block of $100 \%$ amplitude subcarrier (often called the chrominance minibar) which occupies the first half of the second i.t.s. line on lines 20 and 333. Make the subcarrier sit symmetrically about the $50 \%$ luminance pedestal which carries it. Ideally, the block should also extend fully between blanking and white level; however, if there is a difference in circuit amplification between upper and lower halves of the video band - i.e. chrominanceluminance gain inequality - which remains even after the step 14 adjustments, then $L_{9}$ must be set simply for best symmetry of the chrominance minibar. This setting is important because an unfortunate side-effect of the group-delay equalizer will be to increase asymmetry of the chrominance component if this contains an appreciable amount of second harmonic.

Step 16 completes the adjustment of inductors in the circuit. The final pracesses in the line-up procedure are concerned with the a.g.c. and a.f.c. systems.
17. The a.g.c. buffer amplifier, $\mathrm{Tr}_{13}$, should receive a video signal of 4 V pk-pk from the $\mathrm{Tr}_{4} / \mathrm{Tr}_{5}$ stage. If neces-

(a)

(b)

Fig. 24. Distorted 2T and 10T pulses mainly corrected by tuner group-delay equalizer but comparing remaining distortion caused by different transmitter characteristics.
sary, arrange this by means of the temporary a.g.c. divider variable; also, set the blanking-level voltage between 15 and 15.7 V by adjusting $\mathrm{R}_{19}$ in the supply voltage feed to $\mathrm{IC}_{1}$ (MC1330). The requirement here is for $D_{9}$ to slice the picture component just above blanking level. To check that this slicing action is correct, connect the oscilloscope to the $\operatorname{Tr}_{13}$ emitter $\left(\mathrm{R}_{104}\right)$ test point provided and observe the i.t.s. staircase waveform in which slicing should occur between the second and third stairsteps In practice, because the component called for in the $D_{9}$ position has a zener-point spread quoted from 15.2 to 16.9 V - for a current of 1 mA adjustment of $\mathrm{R}_{19}$ will also be necessary to create the required conditions for the zener diode actually being used. As a simplifying alternative, $\mathrm{D}_{9}$ could be omitted and slicing will then be achieved by ordinary potential division in the remaining circuit if $R_{67}$ is changed to $5.6 \mathrm{k} \Omega$. In reception conditions where aircraft flutter is a particular problem, however, the stabilizing effect of the zener is of such benefit that it is worth retaining.
18. Use the oscilloscope to check the timing and size of the sampling pulses fed to the gate of the f.e.t., $\mathrm{Tr}_{16}$. These should be of about $3 \mu s$ duration and should extend between 0 V and the voltage appropriate to video blanking level in the signal appearing at the source; a typical amplitude is 15.5 V .
19. Reconnect $R_{74}$; disconnect "a.g.c. test" divider. The threshold of a.g.c. action is controlled by $\mathrm{Tr}_{17}$ emitter potential, i.e. it depends on the setting of $\mathrm{R}_{78}$. However, the overall requirement here is to obtain optimum r.f./i.f. signal conditions by setting the a.g.c. crossover - using $R_{83}$ to determine $\operatorname{Tr}_{18}$ base bias - so that there is minimum buzz on sound consistent with minimum video noise. The last-mentioned can be observed either by using an oscilloscope display of the tuner output signal waveform, or by viewing appropriate areas of the resulting picture on a television screen. Captions are best used for such inspection because some programme signals, particularly those originating at outside broadcasts, are themselves affected by noise and thus
confuse adjustment. Ref to Fig. 13 in part 2, where comparison of curve (a) with curve (b) gives the conditions which obtain with maximum i.f./r.f. a.g.c. overlap and (b) compared with (c) shows minimum overlap. The practical compromise will, of course, be somewhere between these two limits.
20. To set up the a.f.c. system, first with a.f.c. off and the relevant station-selection buttons pressed in sequence, check that the settings of $\mathrm{R}_{90}, \mathrm{R}_{93}, \mathrm{R}_{96}$ and $\mathrm{R}_{99}$ give as nearly as possible correct tuning for the chosen stations.

Accurate manual tuning is best achieved by connecting the oscilloscope to the MC1330P pin-4 test point via $\mathrm{R}_{102}$ 'and observing each selected transmission, particularly checking for equality of amplitude between the 2 T pulse and luminance bar and for absence of 6 MHz "fuzz". Now ṣwitch a.f.c. on and slightly re-adjust the core of $L_{18}$ if necessary to correct for any change in tuning caused by spurious a.f.c. error signals which may occur (for reasons already discussed in the note to step 10). By checking again the output i.t.s. waveform and setting $\mathrm{L}_{18}$ for minimum 6 MHz component, the sound carrier is accurately positioned at the zero-crossing point of the discriminator $S$ curve. After this re-adjustment of $\mathrm{L}_{18}$ remember that it is most important to fix the core firmly (use p.t.f.e. tape or elastic band): to prevent movement and the resulting disturbance in these critical tuning conditions. Note that, with the a.f.c. switch off, the f.m. de-emphasis timeconstant is incorrect because $R_{62}$ and $R_{61}$ are out of circuit. The effect of this error is too small to be of any consequence and will certainly not be detected in the sound output.

## Sound-only tuner line-up

Line-up of the sound-only tuner is a much simpler process than for the full circuit because there is no need for care in the adjustment of video i.f. pass-band shape, the position of notches or the phase-cancellation point. Thus, after

Fig. 25. Audio feed isolator circuit up to coax link is housed within tuner or receiver; circuit may be live up to isolator. Pins are numbered looking from top.
checking that direct voltages agree with the appropriate values given in the table, the following operations should be sufficient to obtain a satisfactory sound output. It is assumed initially that a signal generator capable of giving a single-frequency output in the 30 to 45 MHz range is available; subsequently, an alternative method is suggested for use in the absence of a generator.

1. Set the circuit gain provisionally by adjusting the a.g.c. test divider variable to give $\mathrm{a}+4.5 \mathrm{~V}$ input at point D in Fig . 2. Panel-mounting variables of $2 \mathrm{k} \Omega$ are not readily available and as this resistor is now required as a permanent control, it could be changed to one of $5 k \Omega$. The two fixed resistors in the divider chain must then also be changed to $10 \mathrm{k} \Omega$ (at the top) and $2.2 \mathrm{k} \Omega$.
2. Loosely couple the signal generator to the u.h.f. module i.f. injection point in the manner already explained in step 8 of the full line-up. Set the frequency of the injected signal as accurately as possible to 33.5 MHz .
3. Connect a voltage-measuring instrument to the $\mathrm{R}_{64}$ test point to inspect the output of the discriminator circuit. Ideally, an oscilloscope or an electronic voltmeter should be used here because it is a high-impedance point and therefore cannot accept additional loading without confusing change in the measured parameters. Failing the ideal, a low-current meter such as an Avo Model 8 or similar can be used, connected between the test point and the tuning-voltage supply - point $F$ in Fig. 2 - but remember that the indicated voltage only represents the true circuit output when it is at or near 0 V , i.e. when the meter is taking no current.
4. Adjust the core of $L_{18}$ to find the sharp. negative and positive voltage peaks which show the limits of the $S$ curve. The voltage peaks will probably not be equal about zero, because the overall amplitude-frequency response of the preceding band-pass circuits has yet to be optimized, and the maximum values obtained will depend on the extra loading involved. If an Avo 8 is being used, the maxima will be in the region of $\pm 4.5$ to 5 V . Having located these

positive and negative peaks, set $\mathrm{L}_{18}$ core so that the discriminator output is at 0 V between them for the injected sound i.f., of 33.5 MHz .
5. Remove the generator coupling and connect an aerial to the tuner. With the a.f.c. switch off, manually set the tuning voltage fed to the u.h.f. module so as to receive a transmission. The discriminator output should remain at about 0V and, as a check on the $L_{18}$ setting just made, the tuning voltage can be varied slightly to sweep this output through to its positive and negative limits. By listening to the incoming programme the i.f. circuits can be adjusted, tuning for maximum sound level. Start with the high-Q coupled pair $L_{14} / L_{15}$, followed by $\mathrm{L}_{17}, \mathrm{~L}_{16}$ and finally the i.f. coil in the ELC1043 module. With these circuits roughly set in this way, final adjustment is made as follows. Reduce the r.f./i.f. gain by increasing the a.g.c. voltage. With the narrow-band sound carrier signal now at lower level, the wide-band circuit noise has greater effect and the i.f. coils can be re-adjusted, this time seeking best signal-tonoise ratio. This occurs when the overall pass-band is as narrow as possible consistent with proper recovery of the modulation.
6. With the circuit response set as described, the a.g.c. voltage is then reduced, allowing the gain to rise with a consequent decrease in the audible noise. The programme sound should
first rise and then stay at the same level after the onset of limiting. The change in a.g.c. is continued until just beyond the point where no further difference in noise level can be detected.. The limiter, of course, maintains a reasonably constant level of audio r.f. carrier against varying r.f./i.f. signals. The optimum degree of limiting is 15 to 18 dB ; if more is used, the effective sound i.f. bandwidth (Fig. 20) increases with a consequent worsening of sig-nal-to-noise ratio. For even greater amounts of limiting, there is also the possibility that energy from the video luminance or chrominance sidebands (see the spectral distribution shown in Fig. 17), or even from the lower adjacent-channel transmission, could intrude into the widened passband and add unwanted background to the programme signal.

The remaining adjustment is the tuning of the adjacent-channel traps, $\mathrm{L}_{1}$ and $L_{3}$, and this ideally requires the use of a signal generator. Fortunately, in most locations, adjacent-channel interference on sound is not a problem and these circuits can normally be omitted.

If a signal generator is not available for establishing the correct tuning point for $L_{18}$ as in steps 3 and 4 above, this can . be achieved by listening to the noise generated in the r.f./i.f. circuits. With no aerial connected and as much gain as necessary, the sound output will be white noise at a low level. As $L_{18}$ core is adjusted, the sound noise level
increases; when at maximum, the noise energy accepted through the i.f. passband is being demodulated along the greatest discriminator slope - the $S$ curve is then positioned with its centre in the middle of the i.f. band. Once the discriminator is set in this way, the rest of the line-up then proceeds as explained in steps 5 and 6.
The l.f. response is maintained to 10 Hz and tuner output-signal frequencies as low as this could be something of an embarrassment if the sound equipment being used is capable of fully responding to them. To prevent problems arising, therefore, it would be worthwhile feeding the audio signal via an op-amp connected as an active high-pass filter cutting off sharply at, say, 20 Hz . If this extra circuit is connected in the pre-amplifier then it would also take care of unwanted high-amplitude l.f. signals from other possible sources, e.g. pick-up handling noises and d.c. thumps from f.m. tuners when station re-selection takes place. Suitable circuit for such filters, including design calculations, have appeared in several recent articles (e.g. Active crossover networks, Wireless World, November, 1974).

## Audio connection through coupler

Picture display equipment can be either of the monitor type such as the Decca model CS2240L or modified conventional receiver. The first represents the
Performance
Video bandwidth see Fig. 12.
Pulse \& bar performance: see Fig. 23.
Chrominance/luminance delay inequality: see
$10 T$ pulse, Fig. 23.
Signal-to-noise ratio see Fig. 25.
Differential gain $3 \%$.
Differential phase: $\pm 3$.

## Performance of sound-only tuner

Harmonic distortion $-50 \mathrm{~dB}(0.3 \%)$
(Figure limited by u.h.f. modulator) Overload margin $\quad 20 \mathrm{~dB}$ over 0.5 V r.m.s. for t.h.d. $-40 \mathrm{~dB}(1 \%)$ Signal-to-noise ratio 50 dB for $\pm 50 \mathrm{kHz}$ Amplitude response 20 Hz to 20 kHz
$\pm 0.5 \mathrm{~dB}$

## Performance using coupler

11 mA standing current ( 50 kl ) pot. at min)

| Harmonic distortion | $-46 \mathrm{~dB}(0.5 \%)$ |
| :--- | :--- |
| Overload margin | 20 dB |
| Signal-to-noise ratio | -50 dB |

Signal-to-noise ratio
Amplitude response
as above
6.5mA standing current ( 50 ks ) pot. at half)

| Harmonic distortion | $-40 \mathrm{~dB}(1 \%)$ |
| :--- | :--- |
| Overload margin | 14 dB |
| Signal-to-noise ratio | -54 dB |

Signal-to-noise ratio $\quad-54 \mathrm{~dB}$

Differential gain results in colour subcarrier amplitude change with luminance level, seem as colour saturation as brightness changes. $14 \%$ is just perceptible. Differential phase results in colour subcarrier phase change with luminance level, seen as hue/colour change. $\pm 25$ is just perceptible for delay-line PAL decorters (which convert phase errors into desaturation). $\pm 6^{\circ}$ is just perceptible for simple PAL decoders, used on some Japanese sets. and shows as a hue-colour change or alternate lines (Hanover bars).

In interpreting the performance figures given, remember that in television productions, sound is inherently at a disadvantage because the picture always takes precedence. Exceptionally , then, the ambient noise level at the source could be as high as -45 to -40 dB with respect to the wanted signal.

To test the noise performance of the tuner, both varicap and mechanical versions were measured off-air using locally-generated signals (testing on empty lines 12 and 325 gated from field-blanking interval). The curves in Fig. 26 were plotted from the results of these tests and show that to take full advantage of the tuner's high performance, particularly its overall bandwidth capability, the r.m.s. carrier level at the aerial input should not be less than 1.5 mV . At the other end of the r.f. signal range, the Mullard data sheet on the ELC 1043 shows that an input level of 8 mV gives $1 \%(40 \mathrm{~dB})$ cross-carrier intermodulation, i.e. the spurious 1.57 MHz signal caused by beating of sound and chroma carriers, which results in the annoying herring-bone patterning on the picture; a $1 \%$ level is, incidentally, just perceptible. For even larger inputs, there are the added problems of compression - Mullard quote $30 \%$ for a 15 mV r.m.s. input and u.h.f. local-oscillator pulling.


Fig. 26. Curves show satisfactory noise performance (on a "just perceptible" basis at 38 dB ) is obtained for inputs as low as $700 \mu \mathrm{~V}$ at the aerial socket.
Sound signal-to-noise ratio is better than 50 dB over the whole range.

The narrow window of input levels to be aimed at is between 2 and 5 mV r.m.s. vision carrier. Obviously, proper measuring equipment provides the best means of checking that the optimum conditions obtain, but for home constructors a good indication of signal-to-noise performance can be gained by
ideal, not only because it is specifically designed to accept, and do justice to, a fully-corrected, colour-coded video signal, but because it is powered through a double-wound transformer, thus avoiding the interconnection difficulties associated with the live chassis common in domestic receivers. But the expense involved in buying specialist equipment of this sort makes it a daunting prospect, and the alternative idea of using a standard receiver is an attractive compromise despite the extra work needed in modifying the decoder and in making sure that the installation is safely isolated from the mains.
Assuming that an existing set is to te used, the problem of isolation mainly concerns the tuner outputs, because co-axial isolators for the u.h.f. aerial input are easily obtained - they are fitted to many commercial receivers not only to protect against mains voltages on the chassis but also against aerial static.

Considering first the video output feed, this necessarily requires a direct earth connection to the decoder and display circuits; there is no practical way of avoiding it, mainly because isolators currently available do not have sufficient bandwidth to handle video signals. The simplest solution here is to accept the fact that the tuner could be live and install it and its power supply with the rest of the TV equipment in the cabinet. This option obviously precludes the possibility of using the tuner
comparing the vision noise observed on electronically-generated captions against that on ordinary camera pic-tures. If the caption noise is noticeably less, then it shows that the receiving system is not introducing the majority of the noise.

To obtain a higher input level, better aerial or aerial siting is called for, or even investment in a mast-head amplifier such as the Antiference type MA 102 with PU102 power unit, together costing $£ 8$ in mid-1975. A directional aerial is sometimes needed to reduce multipath effects which give rise to ghost images on the screen; if so, the incoming wanted signal might then be at too great a level and a co-axial attenuator (e.g. the Belling-Lee components with values of $3,6,12$ and 18 dB ) would be required. The caption-noise test can be used to ascertain optimum attenuation; tuner input should be reduced until noise on captions starts to increase, and then raised by 6 to 12 dB so that the applied level is within the recommended range.
Tests on sound signal-to-noise were also done, and the acceptable window of r.f. input level was much greater than for video. The test showed a ratio better than 50 dB over the range $100 \mu \mathrm{~V}-10 \mathrm{mV} \alpha$, and worse than this only when the input sound carrier level was reduced below $100 \mu \mathrm{~V}$ ).


Sound-only tuner with alternative push-button tuning control.
as a sort of remote-control box, but is strongly recommended as the only way to ensure complete safety in operation.

Warning: never uncover the TV set chassis without first disconnecting it from the mains supply. If in any doubt about what is or is not safe practice, have the completed tuner installed by an experienced service engineer or seek advice from Manor Supplies.

The remaining connection to be made is the feed to the audio units and, for this, a reasonably straightforward isolating circuit can be built using a coupler (Fig. 25). These devices, called photocouplers, or optocouplers, provide an additional benefit in that, unlike transformers, they prevent the circulation of mains-frequency earth currents; low-level signal runs are thus better protected against hum interference. The component used in the circuit shown is the Mullard Darlington photocoupler CNY48 which provides input-to-output isolation up to 4000 volts.

The circuit must be installed with the tuner in the receiver cabinet. The BC107 transistor provides a current-source to the infra-red-emitting diode in the coupler. Both devices are fed from the tuner d.c. supply which, as already explained, could have its 0 V rail at the full a.c. mains voltage carried through from the display-circuit earth. The Darlington pair in the coupler package receives d.c. via the audio co-ax inner from the pre-amp power supply, and will be free of mains connection.

Three of the circuit components are added at the pre-amplifier input to provide part of the collector load for the Darlington pair, to couple the audio signal and to keep a charge on the coupling capacitor so that d.c. surges and consequent loudspeaker thumps do not occur each time the TV sound output is selected.

The CNY48 gives a very reasonable performance as shown by the appended figures. The $50 \mathrm{k} \Omega$ variable resistor allows adjustment of the diode standing current and is provided because there is some spread in the operating characteristics of couplers and the setting used will be a compromise between best signal-to-noise and least distortion. There have been reports which suggest possible ageing difficulties with these diodes, so note that the lower the standing current used, the less these ageing effects should be.

## Literafure Received

All published information on the Motorola M6800 microprocessor is now available from Celdis Data Centre. 37;39 Loverock Road, Reading.

Mark 3 Emicron geared motor units a re described in a new publication from Emicron Industrial Devices Lid, 9a Annete Road, London N76ET ... WW401

New and used test equipment is listed and described in a catalogue produced by Electronic Brokers Lid, 49/53 Pancras Road, London NWI 2QB. The catalogue costs 50 p or 11 for overseas readers.

Burr-Brown have sent us a leaflet on the Micromux data acquisition system, which consists of a 16-input, remotely-sited multiplexer and encoder, conımunicating with a central receiver by ordinary twisted pair. The leaflet is obtainable from Burr-Brown Ltd, Permanent House, 17 Exchange Road. Watford WD1 7EB

Automatic inspection of printed-board assemblies is explained and its application in a variety of manufacturing situations examined in a new publication from Marconi Instruments "Getting to the point". Copies can be obtained from the publicity department. Marconi Instruments Ltd. Longacres. St. Albans. Herts. . . . . . . . . . . . WW W 403

We have received from the IEE a list of available publications. including periodicals. conference publications, colloquium digests and many others. Marketing Department, Institution of Electrical Enginpers. PO Box 8, Southgate House. Stevenage. Herts SGl 1 HQ

WW 404

## Circuit Ideas

## MSF receiver using a tone decoder

Using a tuned aerial, preamplifier, and a Signetics tone decoder, type NE567V, the time signals broadcast on 60 kHz from MSF Rugby may be detected. The tone decoder is a phase locked loop device containing an oscillator which, when set to 60 kHz , will lock to an input signal greater than 20 mV r.m.s. at this frequency causing the output to switch rapidly from high to low.

The time signals, which are directly related to the national time and frequency standard, consist of 100 ms interruptions in the r.f. signal each second, with a 500 ms break at the minute mark; the call sign is given twice just before the hour. Depending on the application, receiver design is a compromise between wide bandwidth with accurate resolution of the carrier on/off transition, and narrow bandwidth with good signal to noise ratio. Conventional heterodyne or "straight" receivers with b.f.o. may render the signal audible and the ear can assist in noise discrimination, but this method gives the time signals as pauses in a tone. This circuit produces the opposite situation where the rectified output is inverted and used to switch an audio oscillator so that pips may be heard instead. The tone decoder is well suited to this application because, having adjustable bandwidth, it will accept a large dynamic range of

## Current limited stabilizer with switch-down

The modification suggested by $P$. C. Bury (WW July 1974, page 239) to the circuit originally due to A. E. T. Nye (WW June 1973, page 285) gives foldback characteristics and, as he states, is probably the most generally useful. If, however, a switch-down action is required, it is possible to achieve this by a similar modification in which a p-n-p transistor is connected as shown. This will reduce the output current to a low value once the limit is reached, and will hold it down until the output load resistance rises to several thousand ohms. Action is similar to that of the additional diode except that when the base-emitter diode is turned on by the falling output voltage, the collector of $\mathrm{Tr}_{3}$ draws current from the resistor R which lowers the effective current limit so that a regenerative process suddenly reduces the output voltage to a low value.

If a reset switch is needed a

normally-closed push button switch in the collector of $\mathrm{Tr}_{3}$ may be used, with the advantage that fold-back limiting. remains in action through the $\mathrm{Tr}_{3}$ base-emitter diode during the switch on process.
R. G. T. Bennett,

Christchurch,
New Zealand.
input signals, and avoids the need for tuned r.f. or i.f. stages.

The aerial consists of a $7 \times 3 / 8 i n$. ferrite rod fully wound (about 360 turns) with 26 s.w.g. enamelled copper wire, tuned with a total of about 1900 pF . A 40 turn step-down transformer wound at one end feeds the two stage amplifier. With a 9 V supply the amplifier has a gain of 4700 giving an r.f. output of 400 mV r.m.s. at 140 km from the transmitter. The zener diode improves oscillator stability with voltage and permits the supply to exceed the 9 V maximum of the tone decoder.

Frequency selection of the tone decoder is by means of $R_{1}$ and $C_{1}$ which have values of $16.6 \mathrm{k} \Omega$ and 1000 pF for 60 kHz . The bandwidth may be selected by $C_{2}$ and $C_{3}$, but this is complicated by the input stage limiting of 200 mV , and the fact that, in this
application, the time markers are given when the r.f. signal goes off. With $\mathrm{C}_{2}=10 \mathrm{nF}$ and $\mathrm{C}_{3}=100 \mathrm{nF}$, and an input signal of 400 mV , the output change follows the r.f.-off by 4 ms and the r.f.-on by lms without any output chatter. By reducing the r.f. signal to between 20 and 100 mV , both off and on delays are less than 2 ms . The output may be taken from pin 8 and a suitable load resistor to switch an audio oscillator or drive an l.e.d. as shown.

The complete circuit may be assembled on Veroboard, taking care to minimise coupling between the amplifier and the tone decoder by earthing all unconnected copper strips, and keeping the aerial at least 1 m from the unscreened circuit.
D. A. Bateman,

Crowthorne,
Berks.


## Synthesizer chord-maker

To play chords on a synthesizer, v.c.os run from a common voltage source plus the appropriate off-sets to each oscillator. To do this all the oscillators must be matched to the same voltage/frequency curve. A more satisfactory solution is to use a configuration similar to that shown in the diagram. The advantages are that only a single v.c.o. is necessary in order to get an octave of notes. The circuit is basically a phase-locked loop frequency multiplier with the divide by N section a master divider. Unfortunately the range is limited by the phaselocked loop locking range.
H. A. Thomas,

Hatfield,
Herts.


Select frequencies as required

## Resistance modulator or periodic switch

This circuit can be used either as an a.c. switch which makes and breaks for approximately equal times at an adjustable frequency, or to produce a modulated resistance at an adjustable frequency. The interesting feature is that both the oscillator and the switch can be made using a single CD4007 c.m.o.s. integrated circuit. This i.c. contains two complementary pairs of transistors, and a complementary transistor inverter. Transistors $\mathrm{Tr}_{1}$ and $\mathrm{Tr}_{4}$
are used as a normal inverter. Transistor $\mathrm{Tr}_{5}$ has a $10 \mathrm{k} \Omega$ load and forms a single ended output inverter. Transistors $\mathrm{Tr}_{2}$ and $\mathrm{Tr}_{6}$ are connected back-to-back to form a transmission gate. The input and output of inverter $\operatorname{Tr} 1_{1.4}$ drive the gates of $\mathrm{Tr}_{6}$ and $\mathrm{Tr}_{2}$ respectively to obtain correct operation of the transmission gate, which acts as an a.c. switch. The inverter is also used, in conjunction with the inverter $\mathrm{Tr}_{5}$, to form an astable oscillator whose frequency is controlled by the time constant CR. Because the two inverters are dissimilar and do not necessarily switch
at $50 \%$ of $V_{D D}$ a mark-space ratio of unity cannot be guaranteed; but the prototype, which ran at switched frequencies of 10,30 and 50 Hz had almost equal on and off times. To obtain a modulated resistance characteristic, resistors $R_{1}$ and $R_{2}$ are connected as shown. The resistance seen at the output terminals is then modulated between $\left(R_{1}+R_{2}\right)$ and $R_{1}$, at the frequency of the oscillator.
D. K. Fryer,

Hydraulics Research Station,
Wallingford.


## L.e.d. clip indicator

The circuit seen here was devised as a simple and economical way of detecting the clipping point in audio amplifiers. The values shown are suitable for a 50 W power amplifier. Transistor $\mathrm{Tr}_{1}$ is normally turned hard on, because of the d.c. potential at the output, and $\mathrm{Tr}_{2}$, with the l.e.d., are turned off. When the collector-emitter voltage of the lower output transistor approaches saturation (i.e. it drops below about 5 V ) $\mathrm{Tr}_{1}$ begins to turn off, $\mathrm{C}_{1}$ charges up through $\mathrm{D}_{1}$ and the l.e.d. turns on. The attack time of the circuit is about 3 ms , determined by $R_{3}$ and $C_{1}$, which approximates to the time taken by the ear to first detect clipping distortion in a good amplifier. The decay time of approximately 50 ms is determined by $\mathrm{C}_{1}$ and $\mathrm{R}_{4}$ in parallel with $\beta R_{5}$, where $\beta$ is the current gain of $\mathrm{Tr}_{2}$. 50 ms is long enough to make a single 3ms overload transient visible but provides adequate resolution to detect more severe overloads where several pulses are clipped in rapid succession.
Because the circuit is referenced to the negative power rail, the detection level is substantially independent of fluctuating supply voltages, hence the circuit works equally well for instantaneous, music or continuous signals.
Transistors $\mathrm{Tr}_{1}$ and $\mathrm{Tr}_{2}$ should have a voltage rating of at least 0.5 supply V for dual-rail amplifiers or at least supply $V$ for single-rail amplifiers. With the last mentioned, $\mathrm{Tr}_{2}$ and the top of $\mathrm{R}_{3}$ must be connected to the positive power rail.
If two indicators are used, one for the negative rail as shown plus one referenced to the positive rail using $\mathrm{p} \cdot \mathrm{n}-\mathrm{p}$ transistors, the circuit can be used to detect short-circuited output transistors because one l.e.d. will be permanently on.
J. Dawson \& K. Northover, Cambridge.


## Sequence generator for radio control

This coder generates a sequence of pulses suitable for the blip-and-hold actuators used in radio control. During the stand-by period, the oscillator formed by NAND gate 1 runs contin. uously at a frequency of 0.5 Hz which can be adjusted by altering $C_{1}$ or $R_{1}$, and the four outputs from the 7490 are held at zero by the presence of 5 volts on pins 2 and 3. If push-button $A$ is depressed the voltage on pins 2 and 5 falls to zero and the 7490 is clocked by the negative edge of the oscillator waveform. When a level 1 appears at outputs $A$ and $C$
concurrently, transistor $\mathrm{Tr}_{1}$ will be forward biased and the output of NAND gate 2 will fall to zero. This will stop the oscillator and thus hold outputs A and C at 1 . If the transmitter modulator is driven from output $A$ the signal will consist of two blips of $1: 1$ mark space ratio by a continuous transmission until button $A$ is released. On release of button A all outputs from the 7490 will again fall to zero. By suitable arrangement of switches and gates a multitude of blip-and-hold combinations can be achieved.
G. D. Southern,

Cronton,
Nr Widnes.


# Transmitter power amplifier design-4 

# Conclusion of the practical circuit described in November; also different strip-line constructions 

by W. P. O'Reilly, M.Sc., M.I.E.E.<br>The Plessey Company Ltd

A spectrum analyser is extremely useful when the first ;prototype of a new design is being evaluated. The harmonic content of the output signal and any tendency for spurious or sub-harmonic instabilities under conditions of varying supply voltage, drive level or load mismatch may be monitored.
In the design which has been described stability is ensured by careful supply line filter design and the use of low-Q base return chokes. This ensures that a low source impedance is presented to the transistors at frequencies well below the operating band where the gain is much higher. Due to the inherent inductance of the low-value resistors $R_{4}$ and $R_{6}$ very little gain at v.h.f. is sacrificed.

Considering the amplifier's input/ output characteristic shown in Fig. 5 of the previous article, a saturated power approaching 25 W is obtained for a total current consumption of 3.3 A at 13.6 V . This corresponds to an overall efficiency in excess of $50 \%$. The collector efficiency of the output stage is over $75 \%$ when operating into a matched load. As with most mobile radio transmitters the possibility of a mismatched antenna must be considered. A prototype amplifier has been operated for prolonged periods into both open and short circuit loads at $10,12.5$ and 14.5 V supply without damage. The power transistors will only tolerate these stress levels if they are properly mounted on an adequate heatsink. The maximum safe drive level is 200 mW . Excessive drive could damage the input device and will reduce the tolerance of the amplifier to load mismatch conditions.

At full drive the input v.s.w.r. is typically $1.3: 1$ and, due to the Class B bias of the input stage, the input impedance does not vary rapidly when drive level is reduced as would be the case if a Class $C$ design were employed. Second harmonic output is -35 dB relative to the fundamental, while third and higher harmonics are below -60 dB . These levels should not represent an interference hazard when an antenna matching filter is in use. It is normal to
employ a bandpass filter common to both receive and transmit circuits and several suitable designs have been published. ${ }^{1.2}$
When operated with the correct drive level and supply voltage a power amplifier of this type has a predicted mean - time - between - failure (m.t.b.f.) in excess of 100,000 hours. (More than ten years of continuous operation.) The amplifier is thus essentially free from routine maintenance.
The power amplifier described makes use of microstrip transmission-lines. Other useful v.h.f. components can be made using strip-lines of different constructions and some of these will now be discussed.

## Tri-plate striplines

Fig. 1 shows a transmission line consisting of a three layer sandwich in which
the inner "live" conductor is equally spaced, by means of dielectric sheets, between two conducting earth planes. In this tri-plate structure the field is entirely contained within the dielectric and so, unlike the microstrip transmission lines previously considered, a true transverse electromagnetic (t.e.m.) wave can exist on the tri-plate line. The guided wavelength, $\lambda$, is no longer dependent upon the characteristic impedance of the line, but is simply related to the free-space wavelength, $\lambda_{0}$ by

$$
\begin{equation*}
\lambda / \lambda_{0}=1 / \sqrt{ } \epsilon_{r} \tag{1}
\end{equation*}
$$

where $\epsilon_{r}$ is the relative permittivity of the dielectric. The earth planes above and below a tri-plate stripline act to confine the field. As a result tri-plate is more suitable for filter applications



Fig. 3. Coupled transmission lines.

(a)
(b)

Fig. 4. Tri-plate coupling structures (a) edge coupling (b) broadside coupling.



Fig. 5. Tri-plate $26 d B$ directional coupler (a) construction (b) performance curves.
requiring a high degree of stopband rejection.

The characteristic impedance of a low impedance tri-plate line in a dielectric of relative permittivity $\epsilon_{r}$ is given by reference three and is shown graphically in Fig. 2 for two common dielectric layer thicknesses. Cross-reference to Fig. 7 in Part 2 of this series reveals that tri-plate conductors of a particular width yield a lower characteristic impedance than would be obtained with a similar width microstrip conductor. This and also the shorter guided wavelength in tri-plate structures tend to compensate for the increase in size and weight resulting from the extra layer of dielectric. Both microstrip and tri-plate striplines are used in modern communications equipment, the choice for any particular application being determined by performance requirements.

## Coupled striplines

Many v.h.f. components make use of the phenomenon of backward-wave directional coupling between striplines which are in close proximity to one another. Fig. 3 shows a pair of coupled transmission lines of length $l$. The electrical length of the lines is given by

$$
\begin{equation*}
\theta=2 \pi l / \lambda \tag{2}
\end{equation*}
$$

It can be shown that if all four ports are terminated in the system characteristic impedance, $Z_{o}$ the coupled voltage at port three is given by

$$
\begin{equation*}
V_{3}=V_{1} \cdot \frac{j k \sin \theta}{\sqrt{1-k^{2}} \cos \theta+j \sin \theta} \tag{3}
\end{equation*}
$$

The voltage at the output port on the main line is:-

$$
\begin{equation*}
V_{2}=V_{1} \cdot \frac{\sqrt{1-k^{2}}}{\sqrt{1-k^{2} \cos \theta+j \sin \theta}} \tag{4}
\end{equation*}
$$

In an ideal coupler no output appears at port four, which is termed the isolated port. In practice due to manufacturing imperfections there is always a small output at port four, and the expression $20 \log \left(V_{3} / V_{4}\right)$ is termed the directivity of the coupler (in dB ). Constant $k$ is the coupling factor, generally expressed in dB , and is related to the degree by which the characteristic impedance of one line is affected by the presence of the other. This inter-dependence is characterised in terms of the odd mode and even mode impedances, $Z_{\infty 0}$ and $Z_{o e}$, of the lines. $Z_{o o}$ is the impedance of one line when the adjacent line is at the same potential but $180^{\circ}$ out of phase, and $Z_{o e}$ is the impedance of one line when both lines are at the same potential in phase and magnitude. The coupling factor and system characteristic impedance are related to the even and odd mode impedances by the relationships:-

$$
\begin{array}{r}
k=\frac{Z_{o e}-Z_{o o}}{Z_{o e}+Z_{o o}} \\
Z_{o}=\sqrt{Z_{o e} Z_{o o}} \tag{6}
\end{array}
$$

Once the required values of even and odd mode impedances are known the physical dimensions of a particular line structure may be calculated. ${ }^{4}$
It can be seen from equations (3) and (4) that the coupling reaches a maximum when the length of the coupled lines is an odd multiple of $\lambda / 4$. The phase difference between the outputs of the main line and coupled line is $90^{\circ}$ at all frequencies.
Two commonly used arrangements of tri-plate coupled striplines are shown in Fig. 4. The edge-coupled structure is suitable for coupling factors up to 10 dB . For tighter coupling the spacing, $s$, becomes very small and presents manufacturing difficulties. The broadside coupling structure of Fig. 5 (b) is more suitable for producing close coupling such as in two-way equal power (3dB) dividers.

## Power measurements

It has been seen that a directional coupler may be used to bleed off a sample of the signal flowing in one direction along a line. If a pair of similar couplers are connected with their main lines in series it is possible to monitor both the forward power component and reflected power component on the main line. Such an arrangement used in the output circuit of a transmitter can provide useful information both for tuning an antenna and for controlling the output power to a safe value for the

A member of the Wireless World staff has made available a printed circuit board for the stripline r.f. power amplifier. The board, which measures approximately $23 \times 11 \mathrm{~cm}$ is supplied roller-tinned and drilled. The one-off price is $£ 4.50$ inclusive. All cheques and postal orders should be made payable to M. R. Sagin and sent to 11 Villiers Road, London N.W.2.
particular value of load mismatch present. The net output power is simply the forward power less the reflected power. Thus if the coupling factors are accurately known this arrangement may be used to produce an accurate wattmeter. A dual directional coupler may also be used to measure reflection coefficient or v.s.w.r., since

$$
\begin{equation*}
\rho=\frac{V_{R}}{V_{F}} \text { and v.s.w.r. }=\frac{1+|\rho|}{1-|\rho|} \tag{7}
\end{equation*}
$$

The measurement of r.f. voltage involves the use of specialized equipment and so it is normal to employ diode detectors with the couplers. The d.c. output of the detectors may then be processed to provide $|\rho|$ and v.s.w.r. indications. Coupling factors of between 20 dB and 40 dB are suitable for
this application. Fig. 5 (a) shows the construction of a 23 dB directional coupler suitable for use in the twometre band. The characteristic impedance is 50 ohms. This coupler is manufactured from 0.8 mm ( $1 / 32 \mathrm{inch}$ ) di-clad p.tf.e. - glass material and uses edge coupling. The required dimensions are: conductor width, $w=1.10 \mathrm{~mm}$; conductor spacing, $s=0.70 \mathrm{~mm}$; conductor thickness, $t=.071 \mathrm{~mm}(2 \mathrm{oz} / \mathrm{sq}$. ft .); and length of coupled lines $=$ 370 mm .

The insertion loss of the coupler is approximately 0.2 dB over the frequency band 100 MHz to 200 MHz . The directivity is mainly determined by the dimensional accuracy of the dielectric material and printed conductors. Fig. 5 (b) shows the performance obtained from a prototype coupler made to this design. To modify the design for any other frequency band it is merely necessary to alter the length of the coupled lines using the formula:-

$$
\begin{equation*}
l_{c}=\frac{47.5}{\text { frequency in } \mathrm{MHz}} \text { metres } \tag{8}
\end{equation*}
$$

## Power combining at v.h.f.

When the required output power cannot be obtained from a single transistor the designer must consider combining the outputs of two or more transistors. Most v.h.f. amplifiers having output


$x=0.8 \mathrm{~mm}$ holes for components
$\begin{array}{ll}\odot & =1 \mathrm{~mm} \text { holes for " } 2 \text { "link through connections } \\ \bullet=3.5 \mathrm{~mm} \text { holes for } 0 . c \text { b. mounting } & (T)=11 \mathrm{~mm} \text { noles for transistors }\end{array}$

- $=3.5 \mathrm{~mm}$ holes for P.C D. mounting

Printed-circuit pattern and a view of the amplifier, showing the location of major components for the stripline board. The pattern is approximately half-size- $\dot{\alpha}$ full-size drawing can be obtained from this office.

capability in excess of 100 watts make use of one of the following combining techniques.

Parallel connection. Simply connecting r.f. power transistors directly in parallel is rarely successful. Unless the two devices are nearly identical in input impedance and current gain one transistor will accept more than its share of the drive power. This causes its internal temperature to rise, thus lowering the turn-on voltage, $V_{B E}$, and increasing the conduction angle. As a result the input impedance and collector efficiency fall, causing extra heat dissipation and eventual failure. Extra stress is then imposed upon the surviving transistors so that it is not uncommon for all of the output devices to fail successively in circuits of this type.

When parallel connection is used it advisable to provide a separate matching network to the base of each transistor in order to help stabilise the drive power. If a relatively narrow bandwidth is required then drive stability may be greatly enhanced by designing base matching networks having electrical length $\lambda / 4$. As the input impedance of one device falls the
impedance at the input to its matching network rises and accepts a smaller fraction of the available drive power.
Wilkinson hybrid combiner. The n-way in-phase power divider/combiner devised by Wilkinson ${ }^{5}$ is a narrow band system using quarter wave transmissión lines. Fig. 6 (a) shows a three-way power divider of this type. The ballast resistors, $R_{B}$, provide a high degree of isolation between the output ports at the centre frequency. The isolation and v.s.w.r. fall off to typically 20 dB and $2: 1$ respectively for frequencies $30 \%$ removed from the band centre. Variations on this principle have been devised using multiple sections of transmission. line to obtain wider bandwidths.

Wilkinson divider/combiners may be constructed using stripline techniques. The optimum line characteristic impedance is

$$
\begin{equation*}
Z_{o}=\sqrt{n} \cdot R_{s} \tag{9}
\end{equation*}
$$

where $R_{\mathrm{s}}$ is the system characteristic impedance. If a large number of amplifier modules are to be combined it may be necessary to employ a system characteristic impedance lower than 50 ohms so that the required matching line impedance can be realised in stripline form.
$\mathbf{9 0}^{\circ}$ combiners. Two power transistors or amplifier modules may be combined using 3 dB quadrature couplers as shown in Fig. 6(b). This combining system has the useful property that, provided the input impedance of the two amplifiers are nearly identical, a very low input v.s.w.r. is obtained at port 1. Any reflected power at ports 2 and 3 is routed to the ballast resistor at port 4. By using matching networks which provide increasing v.s.w.r. to lower frequencies in the passband of the amplifier it is possible to remove the inherent gain/frequency slope of the transistors. Pitzalis ${ }^{6}$ has determined element values for gain slope compensation networks of 4,5 and 6 dB per octave.

The output impedance of a 3 dB quadrature coupled amplifier usually has a v.s.w.r. of less than or equal to 1.3 and provides an accurately matched source impedance for an antenna coupling network or output filter. This type of combiner provides typically 20 to 30 dB of isolation between the two amplifiers. Should a fault develop in one amplifier the surviving unit continues to drive a matched load provided the output and ballast ports are accurately terminated.
Using coupled pairs of amplifiers as modules $2^{n}$ amplifiers may be combined via a tree network of 3 dB quadrature couplers as outlined in Fig. 6 (c). Typical combining efficiency of $95 \%$ per step can be obtained using commercially available stripline 3 dB quadrature couplers.
$180^{\circ}$ combiners. Push-pull operation of power transistors at v.h.f. demands accurate control of the phase characteristics of the phase-splitting networks which is difficult to achieve using conventional transformers. Transmis-sion-line transformers provide a means of obtaining the required accuracy at frequencies up to several hundred $\mathrm{MHz}{ }^{7}$ The conventional push-pull configuration does not provide isolation between the power transistors, but $180^{\circ}$ hybrid combiners may be used to obtain the push-pull effect with its associated advantage of greatly reduced second harmonic output. The $180^{\circ}$ combiner is essentially a $90^{\circ}, 3 \mathrm{~dB}$ coupler with the addition of a fixed $90^{\circ}$ phase shifter in series with one output port. Isolation is at least as good as that obtained using quadrature couplers, but insertion loss, phase error and v.s.w.r. are usually significantly greater.

## Future developments

It has not been possible in this brief series to cover all of the r.f. power amplifier design techniques in use today. The last few years have seen great improvements in frequency coverage. and power output capability of transistors. Recently introduced devices are much more rugged than the transistors of three or four years ago, so that now with careful design very high reliability may be achieved in cost-effective
mobile radio systems. The future will see further improvements in power-frequency capability of transistors which are presently at least 10 dB below the theoretical limitation. For broadband applications some manufacturers are including part of the input matching circuitry inside the transistor package, and for some medium power requirements full input and output matching is included. This trend is likely to increase and, as high permeability materials - originally pioneered for microwave hybrid integrated circuits become less expensive and more readily available, complete power amplifier chains may be available on a custom design basis.
There is little reason to doubt that as manufacturing technology conquers present equipment requirements so the necessity for additional communication channel capacity will be Mother to the invention of new modulation processes which will present a challenge to the designers of radio equipment in the years ahead.

## References

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2. Mullard Technical Communication No. 115, July 1972, P. 167.
3. Cohn, S.B. "Problems in Strip Transmission Lines", Trans. IEEE PGMTT3, 1955, p.119.
4. Matthaei, G.L., Young L.' and Jones, E.M.T. "Microwave Filters, Impedance Matching Networks and Coupling Structures", McGraw-Hill, 1969.
5. Wilkinson, E.J. "An N-way Hybrid Power Divider", Trans. IEEE PGMTT, January 1960, pp.116-118.
6. Pitzalis, Jr., O. and Gilson, R. A. "Tables of Impedance Matching Networks which Approximate Prescribed Attenuation", Trans. IEEE Microwave Theory and Techniques, Vol. MTT-19, No. 4, April 1971, pp.381-386. 7. RCA Application Note AN6126.

The author has asked us to point out that a correction is needed to Fig. 7 published in Part. 2 of this series (October issue). The upper curve of $Z_{0}$ vs. strip width has been plotted for air spaced microstrip and not for p.t.f.e. as stated in the text. The following co-ordinates can be used from which the true p.t.f.e. curve can be plotted.

| $Z_{0}($ ohms $)$ | width $(\mathrm{mm})$ |
| :---: | :---: |
| 120 | 0.7 |
| 80 | 2.0 |
| 50 | 4.2 |
| 25 | 11.0 |

Transistors $\mathrm{Tr}_{2}$ and $\mathrm{Tr}_{3}$ in the components list of Part 3 in the November issue should be 2 N 5590 and 2 N 5591 respectively and not 2 N 5990 and 2N5991 as stated. Also Fig. $4(\mathrm{~g})$ shows coaxial braid soldered to earth with a "pigtail" made prior to soldering instead of the braid soldered directly to earth as it should be to avoid degrading the v.s.w.r. of the coaxial to stripline transition.


## LONDON

6th. IEE - "Role of government in support of automation in the UK" by J. H. Major at 17.30 at Savoy Place, WC2.
8th. IEE - 11th Appleton lecture on "Ionospheres of dead stars and living galaxies by Prof. R. L. F. Boyd at 17.30 at Savoy Place, WC2.

9th. IEE - Colloquium on "Solid state transmitters for radar" at 11.00 at Savoy Place, WC2.
13th. IEE/RTS - Colloquium on "Broadcast and wired Teletext systems - Ceefax, Oracle, Viewdata, at 10.00 at Savoy Place, WC2.

14th. IEE/I.Phys. - One-day meeting on ,"Integrated optics" at Imperial College, SW7.
14th. IERE - Colloquium on "Current trends in surface acoustic waves" at 14.00 at 9 Bedford Square, WCl.

14th. IEE - "Electronics as an aid to off-shore operations" by R. Nesbit at 17.30 at Savoy Place, WC2.

15th. IEE - "Underwater remote controlled vehicles" by R. T. Holmes and R. M. Dunbar at 17.30 at Savoy PI., WC2.

15th. IEE - Discussion on "Private telecommunications networks - value for money?" at 17.30 at Savoy Place, WC2.

20th. IEE - Colloquium on "The technology of instrumentation in off-shore applications" at 10.30 at Savoy Place, WC2.
21st. IEE - "Surround sound technology" by Prof. P. B. Fellgett at 17.30 at Savoy Place, WC2.
21st. IERE - Two lectures on "Security systems - intruder alarms" at 18.00 at 9 Bedford Square, WCl.
22nd. IEE - "Antistatic - active and passive methods for removal of static electricity" by Prof. A. W. Bright at 17.30 at Savoy Place, WC2.

22nd. IERE - "Integrated injection logic efficiency and versatility in solid state" by C. S. den Brinker at 18.00 at 9 Bedford Square, WC1.
28th. IEE - Colloquium on "Stability theory and non-linear oscillations" at 10.30 at Savoy Place, WC2.
28th. IERE/IEE Colloquium on "The impact of 1.s.i. components on logic design" at 14.00 at 9 Bedford Square, WC1.
29th. IEE - Colloquium on "Pattern recognition - fact or fiction?" at 14.30 at Savoy Place, WC2.

## ABERDEEN

15th. IEE - "Understanding lasers" by M. Kears at I4.00 at Robert Gordon's Institute of Technology, Schoolhill.

## ARBORFIELD

22nd. IERE/IEE - "Recent developments in electro-chemical energy storage devices" by Prof. P. Dunn at 19.30 at the Lectùre Theatre, School of Electronic Engineering. REME, Arborfield, Berks.

## BELFAST

13th IEE - "The metallurgy and application of solder" by Peter King at 18.30 at Ashby Institute.

## BIRMINGHAM

5th. IEE - "Public radio 'phone services in the UK" by P. J. Linney at 18.00 at Rm SG31, Sumpner Building, University of Aston, Gosta Green.
21st. IEE - "Engineering starts in the schools implications for manpower planning" by G. B. Harrison at 18.15 at Sumpner Building, University of Aston.

22nd. RTS/IERE - "Radio Communications" by Prof. W. Gosling at 19.00 at Birmingham University.
26th. IEE - Faraday lecture on "The entertaining electron" by F. H. Steele in the evening at the Town Hall.

27th. IEE - Faraday lecture on "The entertaining electron" by $F$. H. Steele in the morning, afternoon and evening at the Town Hall.

## BLANDFORD

2Ist. IEE - "Assessment is more difficult than it looks" by P. McVey at 18.30 at School of Signals, Blandford Camp.

## CAMBRIDGE

29th. IERE;IEE - "Tropospheric scatter communications" by'B. S. Skingley at 18.00 at the University Engineering Laboratories, Trumpington Street.

## CARLISLE

13th. IEE - "My Dear Watson" by G. Phillips at 19.00 at Carlisle Technical College, Victoria, Carlisle.

## CHATHAM

21st. IERE - "Managing your P's and Q's" (P = profit, $\mathrm{Q}=$ Quality) by P. Daisley at 19.00 at Medway and Maidstone College of Technology, Maidstone Rcad.

## CHELMSFORD

14th IEE/IERE - "Charge coupled devices" by D. J. Burt at 18.30 at the King Edward VI Grammar School, Broomfield Road.

## CREWE

12th IEE - "Stereophonic and ambisonic reproduction of sound" by Prof. P. B. Fellgett at 19.00 at MANWEB, Crewe.

## DERBY

6th. IEE - "Signal processing using surface acoustic wave and charge coupled devices" by J. H. Collins at 19.00 at Main Lecture Theatre, Derby College of Art and Technology.

## DUBLIN

15th. IEE - "'The RTE Tullamore high power m.f. transmitting station" by M. J. C. Curley and P. Doyle at 18.00 at Physics Theatre, Trinity College.

## DUNDEE

27th. IEE - "Electronic calculators" by Dr Priestland at 19.30 at Ewing Building, The University, Dundee.

## EASTBOURME

20th. IEE/CE1 - "Choice and use of calculators" by D. W. Glover at 18.30 at SEEBoard, Westlords, Willingdon Road.

## EDINBURGH

6th. IEE - "North sea oil telecommunications" by W. N. Lang at 18.00 at SSEB Offices, George Street.
22nd. IEE - "Liquid crystals" by Prof. Davies at 18.00 at King's Building, Edinburgh University, Mayfield Road.

## EVESHAM

12th IEE/IERE - "Air traffic control in the vicinity of a rajor airport" by $S$. Ratcliffe at 19,30 at BBC Club, High Street.

## hatfield

7th. IERE - Colloquium on "Contemporary teaching of electronics" at 10.00 at Hatfield Polytechnic.
14th. IERE - "A look at classical and modern control theory" by P. Murdoch at 19.45 at Hatfield Polytechnic.

## HULL

21st. IEETE - "TEC - The introduction of new course struckures for technician education" by J. W. M. McKain at 1930 at Ernest Bullock Lecture Theatre, College of Technology, Queens Gardens.

## LEEDS

6th. 1EE - "My Dear Watson" by Dr D. N. S. Peach at 18.30 at Leeds University.

## LIVERPOOL

6th IEE - "Electronics in crime prevention" by G. Philips at 18.30 at the Dept. of Electrical Engineering, Liverpool University.
21st. LEE/I Mech.E - "Ergonomics" by P. Crawley at 18.30 at The Polytechnic, Byrom Street.

Meetings continued on page 46


## Outlook costly

The effects of inflation and $25 \%$ VAT continue to hoist the costs of operating an amateur radio station. The Home Office has proposed a $60 \%$ rise in all fees for standard radio licences (other than broadcast reception) with effect from December 1, 1975. This includes all amateur and model radio control licences, taking the basic annual fee for a Class A or B amateur licence in the UK from $£ 3$ to $£ 4.80$. The Home Office states that the income from the former fees no longer covers costs of issue and administration. Combined total of Class $A$ and $B$ licences recently passed the 21,000 mark. The RSGB is similarly proposing to raise its subscriptions by a roughly similar percentage in an attempt to offset the $£ 18,000$ deficit in 1973-74.

Attendance at the recent Leicester amateur exhibition appears to have been a little lower than in 1974 and a general comment was the effect of $25 \%$ VAT on equipment prices.

As an educational and scientific hobby, however, the costs of amateur radio remain very much what the individual enthusiast wishes to make them.

## Satellite plans

With the highly successful and long living Oscars 6 ard 7 still continuing to provide amateurs throughout the world with active transposers for $144 / 28 \mathrm{MHz}$ and $432 / 144 \mathrm{MHz}$ operation, the various "AMSAT" groups in a number of countries, including the UK, are planning ahead for a new phase of operations.

AMSAT-UK, for example, is hoping to provide the first British amateur-built equipment for an Oscar in the form of a $21 / 28 \mathrm{MHz}$ repeater which, if successful, will provide the first "all-h.f." activity in the Oscar series.

It is also hoped that the next Oscar may be replaced in either a synchronous or highly elliptical orbit (i.e. as used in the Russian Molniya satellites) and so would be capable of providing improved coverage over long periods of time, although at the cost of much greater path losses. With a possible launch date
around mid-1978, the next Oscar is expected to be equipped with highpower transponders for the 145.9 and 435.1 MHz bands. German, Canadian and Australian groups are all co-operating with the Americans in planning for this new phase.

The recent RSGB meeting on AMSAT activities included an interesting account of the building and operation of the Oscar 6 ground control station that 'has been working since last February at the University of Surrey, Guildford, and plays an important role in conserving the state of batteries on board the satellite. An ambitious new Oscar communications and control station is currently being built there involving the use of an ex-ASWE tracking dish which, although weighing several tons, is being installed on the roof of one of the university buildings.

## Television topics

The first slow-scan television convention at Aston University brought over 100 s.s.t.v. enthusiasts to Birmingham in this specialised field of amateur radio. Highlights included the keyboard s.s.t.v. character generators of Richard Thurlow, G3WW and Keith Clarke, G3KRC; four fast-to-slow scan converters; a flying spot scanner built by R. G. Dobdinson, G3RGD and the extremely interesting slow-to-fast converter by Volker Wrasse, DL2RZ which allows s.s.t.v. pictures to be displayed on conventional television receivers thus helping to bridge the gap between slow scan and "normal" television enthusiasts. This German amateur from Keil "stole the show" with frozen s.s.t.v. pictures displayed on 9 and 16 -inch monitors, and Richard Thurlow is one of those who believe that after seeing this form of presentation "s.s.t.v. can never be the same again" though the necessary electronic storage is expensive at present, with a requirement for some 64 type 1404 ics costing some $£ 5$ each.

John Wood, G6AH/T, adds some additional calls to the November list of active 70 cm amateur tv transmitters: G6NOX/T Saffron Walden; G6AMB/T Manchester; G6AAD/T Newcastle-under-Lyne; and F5VA. Several enthusiasts are anxious to establish a recognised "calling frequency" on 144 MHz as a preliminary meeting point for both fast- and slow-scan enthusiasts. Suggestions have included 144.23 and 144.75 MHz . Membership secretary of the British Amateur Television Club is now Brian Summers, G6AJU/T, 13 Church Street, Gainsborough, Lincs (telephone 3940).

## On the bands

A tune across 3.5 MHz on November 1 unexpectedly brought in GB2BP on Platform C in the Forties Field, 110 miles east of Scotland. This was the first amateur radio operation from a North Sea oil rig as part of the celebrations
marking the official start of oil flowing ashore. The size of the rig permitted the use of a vertical half-wave aerial on 3.5 MHz and the international nature of these oil fields was underlined by the fact that the operation on GB2BP was G5APC/W5WC who doubtless learned the oil business in Texas!

The long and intense "tropo" opening over the last weekend in October resulted in enormous numbers of 144 and 432 MHz contacts at distances up to about 1500 km with ducts extending from Sweden in the north to Switzerland in the south; for example Scottish amateurs were able to work into Poland, Czechoslovakia and possibly for the first time Switzerland.

For many years one of the outstanding amateur signals from Angola has been those of Joao Chaves, CR6AI and it was painful to read in the South African Radio-ZS a letter from the 66 -year-old amateur enquiring about possible work there - one of the many amateurs caught up in political turmoil that has greatly reduced the number of amateurs operating in the African continent.

The annual Top Band transatlantic tests, with emphasis on "first-time" contacts across the Atlantic on 1.8 MHz , began this season on November 16 and the other dates are December 21, January 11 and February 8 with times of 0500 to 0730 GMT.

## President 1976

To be formally installed as the 1976 president of the Radio Society of Great Britain on January 23 at Edgbaston, Birmingham is Dr E. J. Allaway, MB, ChB, MRCS, LRCP, G3FKM.

John Allaway is thus one of those whose hobby of amateur radio is something entirely apart from his profession and who has no qualms in showing clearly that he is keenly interested in such traditional amateur radio pursuits as h.f. operation, "dx chasing" and awards and certificates. His year as president is his 30 th year as a member of the Society and his licence dates from 1949. He is a member of the First Class Operators Club, the Certificate Hunters Club and the International Short Wave League.

Dr Allaway seems destined to be a well-liked, well-respected president though it will surely tax his own professional skills to help cure the poor health of the Society's finances which are currently so strained, in spite of a membership which now exceeds 18,000 .

## In brief

American amateurs are facing a marked increase in the difficulties of buying one-off components specified for constructional projects, with firms 'imposing "minimum-billing fees" and claiming that the amateur component market is "too miniscule to warrant stocking specialised small parts."

PAT HAWKER, G3VA

## Current Dumping Strom Ablade <br> Asservissement des étages générateurs de courant Stroomtoelevering



## QUAD

## Now...the most exciting Sinclair kit ever

# The Blac 

 * practical-easily built by anyone in an evening's straightforward assembly.* complete - right down to strap and batteries.
* guaranteed. A correctlyassembled watch is guaranteed for a year. It works as soon as you put the batteries in. On a built watch we guarantee an accuracy within a second a day-but building it yourself you may be able to adjust the trimmer to achieve an accuracy within a second a week. Controlled by a quartz crystal.. powered by two hearing aid batteries ... using bright red LEDs to show hours and minutes and minutes and seconds ...it's also styled in the cool prestige Sinclair fashion: no knobs, no buttons, no flash.
The Black Watch kit is unique, too. It's rational-Sinclair have reduced the separate components to just four.
It's simple-anybody who can use a soldering iron can assemble a Black Watch without difficulty. From opening the kit to wearing the watch is a couple of hours' work.


## The special features of The Black Watch

Smooth, chunky, matt-black case, with black strap. (Black stainlesssteel bracelet available as extrasee order form.)


Large, bright, red display-easily read at night.
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Runs on two hearing-aid batteries (supplied). Change your batteries yourself-no expensive jeweller's service.


## The Black Watch-using the unique Sinclair-designed state-of-the-art IC.

The chip...
The heart of the Black Watch is a unique IC designed by Sinclair and custom-built for them using state-of-the-art technologyintegrated injection logic.
This chip of silicon measures only $3 \mathrm{~mm} \times 3 \mathrm{~mm}$ and contains over 2000 transistors. The circuit includes
a) reference oscillator
b) divider chain
c) decoder circuits
d) display inhibit circuits
e) display driving circuits.

## ... and how it works

A crystal-controlled reference is used to drive a chain of 15 binary dividers which reduce the frequency from $32,768 \mathrm{~Hz}$ to 1 Hz . This accurate signal is then counted into units of seconds, minutes, and hours, and on request the stored information is processed by the decoders and display drivers to feed the four 7 -segment LED displays. When the display is not in operation, special power-saving circuits on the chip reduce current consumption to only a few microamps

The chip is totally designed and manufactured in the UK, and is the first design to incorporate. all circuitry for a digital watch on a single chip.


Complete kit STAS

## The kit contains

1. printed circuit board
2. uniq $\lrcorner \in$ Sinclair-designed IC
3. encepsulated quartz crystal
4. trimmer
5. capacitor
6. LEDdisplay
7. 2 -part case with window in positian
8. batteries
9. battery-clip
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# Electronic circuit calculations simplified 

## 6 - LC circuits

by S. W. Amos, B.Sc., M.I.E.E.

The resonance frequency $f_{o}$ of an LC circuit, whether parallel-connected as in Fig. 1(a) or series-connected as in Fig. (b), is given by the well-known expression

$$
f_{0}=\frac{1}{2 \pi \sqrt{ }(\mathrm{LC})}
$$

In this form the expression is suitable for calculating $f_{o}$ by direct substitution of numerical values for $L$ and $C$. Problems of this type occur: for example it can happen, in designing a circuit, that an inductor and a capacitor are accidentally connected in series or in parallel. It is important then to know the resonance frequency of the combination because the low or high resistance presented by the $L C$ circuit at resonance could have a great effect (beneficial or disadvantageous) on the performance of the circuit.

Knowing $L$ and $C$ we could determine $f_{o}$ quite readily by the use of abacs, but in this series we are concerned only with arithmetical methods such as direct substitution in the expression for $f_{0}$. As an example of such a calculation let us suppose that in an attempt to build a very small medium-wave receiver, a tuning capacitor of 100 pF maximum capacitance is chosen together with an inductor of $800 \mu \mathrm{H}$ (taken from a 465 kHz i.f. transformer possibly). The problem is to calculate the maximum and minimum extremes of the waveband likely to be achieved with this combination. We can estimate the effective total minimum capacitance as 20 pF , made up of contributions from the tuning capacitor, coil selfcapacitance and strays. The maximum capacitance we can take as 110 pF . For the minimum capacitance direct substitution in the expression for resonance frequency gives

$$
\begin{aligned}
f_{0} & =\frac{1}{6.284 \times \sqrt{ }\left(800 \times 10^{-6} \times 20 \times 10^{-12}\right)} \mathrm{Hz} \\
& =1.26 \mathrm{MHz}
\end{aligned}
$$

somewhat short of the 1.6 MHz hoped for. The determination of the other tuning limit is carried out later.

The above calculation was hardly
simple but fortunately needs to be done only rarely: there are simpler methods which do not require direct substitution in the expression for $f_{o}$. These methods depend on two observations which can be deduced from the expression. The first is that if $C$ is fixed, the resonance frequency is inversely proportional to the square root of the inductance. This can also be expressed by saying that the inductance is inversely proportional to the square of the frequency, e.g. if the inductance is quadrupled the resonance frequency is halved. The second observation is that, for a given inductance, the resonance frequency is inversely proportional to the square root of the capacitance: that is to say the capacitance is inversely proportional to the square of the frequency, e.g. if the capacitance is reduced to one quarter of its former value the resonance frequency is doubled.
The simpler method of calculating $f_{0}$ requires a knowledge of the resonance frequency for particular values of $L$ and 'C. A convenient result to know is that if $L=100 \mu \mathrm{H}$ and C $=100 \mathrm{pF}$, the resonance frequency is 1.6 MHz . This can readily be checked by direct substitution in the expression for $f_{0}$.

As a numerical example we will complete the calculation started above: the problem now is to determine the resonance frequency for a circuit in which $L=800 \mu \mathrm{H}$ and $C=110 \mathrm{pF}$. We know that if $L=100 \mu \mathrm{H}$ and $C=100 \mathrm{pF}$ the resonance frequency is 1.6 MHz . Frequency is inversely proportional to


Fig. 1. For given values of $L$ and $C^{1}$ parallel-connected circuits (a) and series-connected circuits (b) have the same resonance freqency.
the square root of inductance and to the square root of capacitance: thus the required value of $f_{o}$ is given by

$$
1.6 \times \sqrt{ }\left(\frac{100}{110} \times \frac{100}{800}\right) \mathrm{MHz}=537 \mathrm{kHz}
$$

very near the required value. But there is difficulty, as shown above, in achieving the desired high-frequency extreme of the medium waveband unless the minimum capacitance in the circuit can be reduced.

The problem of calculating the resonance frequency for given values of $L$ and $C$ is, however, not so frequently encountered as that of calculating the value of $C$ needed to tune a given value of $L$ to a particular frequency (or alternatively of calculating the value of $L$, needed to tune a given value of $C$ to a particular frequency). Such calculations could, of course, be performed by recasting the above expression for $f_{o}$ so as to give an expression for $C$ (or $L$ ) and by substituting numerical values in this. It is, however, simpler to make use of the observation that inductance and capacitance are inversely proportional to the square of resonance frequency.

As a numerical example, suppose we wish to use a medium-wave tuning inductor of $160 \mu \mathrm{H}$ in an i.f. wavetrap in the aerial circuit of a receiver. What value of tuning capacitance is required for resonance at 465 kHz ?

We can begin with the same basic statement as before. If $L=100 \mu \mathrm{H}$ and $C=100 \mathrm{pF}, f_{o}=1.6 \mathrm{MHz}$. Capacitance is inversely proportional to inductance and to the square of the frequency. Therefore the required value of $C$ is given by

$$
C=100 \times \frac{100}{160} \times\left(\frac{1.6}{0.465}\right)^{2}=740 \mathrm{pF}
$$

Reactance method. There is an alternative method of calculating values of $L$ and $C$ which can sometimes be useful. This makes use of the fact that at the resonance frequency of an LC circuit, the inductive and capacitive reactances are equal. Thus if we know or can calculate the reactance of an inductor at the resonance frequency, the required $C$ value is that which has the same value
of reactance at the resonance frequency. To use this method it is necessary to know or to calculate reactance. Calculation is possible by numerical substitution in the expressions

$$
\text { inductive reactance }=2 \pi f L
$$

and

$$
\text { capacitive reactance }=\frac{1}{2 \pi f C}
$$

It is however simpler and more convenient to calculate reactance from a knowledge of its value for a particular inductance (or capacitance) at a particular frequency. Useful results to remember are that an inductor of $100 \mu \mathrm{H}$ has a reactance of 628 ohms at 1 MHz (and that inductive reactance is directly proportional to inductance and to frequency) and that a capacitor of 100 pF has a reactance of 1.6 kilohms at 1 MHz (and that capacitive reactance is inversely proportional to capacitance and to frequency).

The relationship between reactance and resonance is well illustrated in the circuit diagram of Fig. 2. This represents the frequency changer of an f.m. receiver. The transistor is d.c. stabilised by the potential divider $R_{1} R_{2}$ and emitter resistor $R_{3}$. The circuit is basically that of a Colpitts oscillator ( $\mathrm{C}_{3}$ and $C_{4}$ being the two fundamental capacitors and $C_{3}$ being the tuning capacitor). To enable $C_{3}$ to be ganged with other tuning capacitors, the moving vanes of $\mathrm{C}_{3}$ must be earthed. Thus the transistor base must also be earthed. Clearly therefore the emitter cannot also be earthed and the inductor $L_{2}$ is introduced into the emitter circuit to permit r.f. signals at the emitter. An advantage of including $L_{2}$ is that the emitter now becomes a convenient point at which the signal-frequency input can be applied to the transistor. $\mathrm{L}_{3}$ is the primary winding of an i.f. transformer and this cannot be tuned by the usual parallel-connected capacitor because such a capacitor would decou-
ple the collector at r.f., so preventing oscillation. Instead therefore the tuning capacitor $\left(\mathrm{C}_{5}\right)$ is connected between collector and earth via $L_{1}$ which has negligible reactance at the intermediate frequency.

This circuit is a good example of the possibility mentioned earlier, that of forming an $L C$ circuit by accident, for $L_{2}$ and $C_{7}$ constitute an unintentional series-connected resonant pair in the emitter circuit. The negative feedback introduced by these components will seriously reduce the conversion gain of the transistor if $L_{2}$ and $C_{7}$ have any net reactance at the intermediate frequency. Such feedback can be avoided, however, by making $L_{2}$ and $C_{7}$ resonate at the intermediate frequency and the problem is to calculate suitable values for these two components.

First decide a value for $C_{7}$. To minimise feedback the reactance of $C_{7}$ at the oscillation frequency must be small compared with the emitter a.c. resistance of the transistor. A likely value for the emitter a.c. resistance is 25 ohms and thus the reactance should ideally not exceed 2 ohms at 100 MHz . We know that 100 pF has a reactance of 1.6 kilohms at 1 MHz , and thus of 16 ohms at $100 \mathrm{MHz} . \mathrm{C}_{7}$ could hence be 1000 pF which has a reactance of 1.6 ohms at 100 MHz . As $\mathrm{C}_{7}$ must resonate with $L_{2}$ at the intermediate frequency we also need to know its reactance at 10.7 MHz ; this is equal to $1.6 \times 100 / 10.7$ about 15 ohms. Thus the reactance of $\mathrm{L}_{2}$ must also be 15 ohms at 10.7 MHz . We know that an inductance of $100 \mu \mathrm{H}$ has a reactance of 628 ohms at 1 MHz and thus the required value of $L_{2}$ is given by.

$$
\mathrm{L}_{2}=100 \times \frac{15}{628} \times \frac{1}{10.7}=0.23 \mu \mathrm{H}
$$

The reactance of $L_{2}$ at 100 MHz will be nearly ten times the value at 10.7 MHz

Fig. $\overline{2}$. A v.h.f. self-oscillating mixer circuit. The calculations of values for $L_{2}$ and $C_{7}$ are discussed in the text.



Fig. 3. A parallel-tuned circuit damped by a parallel resistor $R$.
and will be thus approximately 150 ohms, many times the emitter a.c. resistance of the transistor and this is the required condition.

Bandwidth of single tuned circuits. Bandwidth is often stated as the range of frequency between the points at which the response is 3 dB down, and for a single tuned circuit the bandwidth is simply related to the $Q$ factor of the inductor thus

$$
\text { bandwidth }=\frac{f_{0}}{Q}
$$

where $f_{o}$ is the centre frequency. For an inductor with $Q=100$ tuned to 465 kHz the bandwidth is 4.65 kHz and if there are three such circuits in the i.f. amplifier of a medium-wave receiver the response is 9 dB down at 4.65 kHz bandwidth i.e. at 2.325 kHz from the centre frequency. This represents a serious loss in high-frequency audio response due to sideband cutting in the i.f. amplifier and the effect may be reduced by damping each tuned circuit by a parallel-connected resistor as shown in Fig. 3. Suppose we decide that the loss of 9 dB can be accepted at 5 kHz . Then the bandwidth of each tuned circuit is 10 kHz and the $Q$ value required for each inductor is 46.5 . What value of parallel resistance is required to reduce the natural $Q$ of 100 effectively to 46.5 ? If we knew the dynamic resistance of the undamped $L C$ circuit at 465 kHz the problem is a simple one of two resistances in parallel. The dynamic resistance $R_{d}$ of an $L C$ circuit is given by

$$
R_{d}=\text { reactance of } L(\text { or } C) \times Q
$$

We are more likely to know the capacitance of $C$ than the inductance of $L$ so, to pursue this numerical example, let us assume that $C$ is 200 pF . This has a reactance of 1700 ohms at 465 kHz and thus the undamped dynamic resistance is $1700 \times 100$, i.e. 170 kilohms. The required, damped, value of dynamic resistance is given by $1700 \times 46.5$ which is approximately 80 kilohms. From Part 1 we can now calculate the required value of damping resistor thus:
product of damped and undamped $R=\frac{\text { dynamic resistances }}{\text { difference between damped and }}$
undamped dynamic resistances
$=\frac{170 \mathrm{k} \times 80 \mathrm{k}}{170 \mathrm{k}-80 \mathrm{k}}$
$=150$ kilohms
In a practical circuit there is no need to add a parallel resistor as indicated in Fig. 3 to achieve the required bandwidth. Bipolar transistors have a low input resistance and this can be used to provide the required damping. For a mean emitter current of 1 mA , a bipolar transistor has typically an input resistance of 2000 ohms. If the base of the transistor is connected to a tapping point on the inductor, as shown in Fig. 4, the required $Q$ value can be achieved by correct positioning of the tapping point. For the example in question we need the effect of a parallel resistance of 150,000 ohms and we have an input resistance of 2000 ohms. The position of the tapping point should be such that there is a turns ratio between the whole coil and that part presented to the transistor of $n: 1$ where

$$
\begin{aligned}
n= & V\left(\frac{150,000}{2,000}\right) \\
& =8.7
\end{aligned}
$$

Thus if the inductor has 150 turns, then the tapping point should be $150 / 8.7$, i.e. 17 turns from one end.
It is alternatively possible to use a capacitive tapping instead of an inductive one. To do this the tuning capacitance of 200 pF is made up of two capacitors in series, the ratio of the two being chosen to give the step-down ratio necessary to provide the required damping. The method of calculating the values of the two capacitors was given in Part 3.
In this calculation it was assumed that there is no contribution to the damping from the collector circuit of the transistor preceding the tuned circuit. This is normally justified because it is common practice to connect the tuned circuit to the preceding transistor via a coupling coil or a tapping point chosen to present the transistor with a resistance equal to its optimum load resistance. The collector a.c. resistance of the transistor is usually very large compared with the optimum load so that damping of the tuned circuit is negligible. The position of the tapping point can be calculated in the following manner.

As shown in Part 1 the optimum load resistance is given by

> | collector voltage swing available |
| :--- |
| collector current swing available |

and for a transistor with a 9 V supply and a mean emitter current of 1 mA is approximately 7000 ohms. ${ }^{*}$ To present the transistor with this value of resistance the position of the collector

[^5]

Fig. 4. Damping of a parallel-tuned circuit by the input resistance of $a$ bipolar transistor.

Fig. 5. Method of matching the input resistance of a transistor to the optimum load of the preceding transistor via an LC circuit and of obtaining a desired working $Q$ value.

tapping point must be such that the turns ratio between the winding presented to the collector circuit and that presented to the base circuit of the following transistor is $n: 1$ where

$$
n=\sqrt{ }\left(\frac{1000}{2000}\right)=\sqrt{ } 3.5=1.87
$$

The base winding has 17 turns and thus the collector winding must have $17 \times 1.87$, i.e. 32 turns. Thus the circuit can take the form shown simplified in Fig. 5. The collector circuit could alternatively be connected to a separate winding of 32 turns closely coupled to the tuning inductor.

There is an alternative solution to the problem of obtaining a required $Q$ value using a given value of damping resistance. This is to connect the resistor across the tuned circuit, as shown in Fig. 3, and to choose the values of $L$ and $C$ so that:
(a) they tune to the required centre frequency
(b) the required $Q$ value is obtained. We can illustrate the method by a numerical example on the i.f. circuit for a 625 -line television receiver. The standard centre frequency is 36.5 MHz and the bandwidth (to include the vision and accompanying sound channel) is required to be 6 MHz . Thus we have

$$
Q=\frac{\text { centre frequency }}{\text { bandwidth }}=\frac{36.5}{6}=6.1
$$

Practical undamped $Q$ values are likely to be very large compared with this. Therefore practical values of undamped dynamic resistance are likely to be very large compared with the damped value. When a high-value resistor is connected
in parallel with a low-value resistor, the net resistance is slightly less than that of the low-value resistor and, with little error, can be taken as equal to the low-value resistor. Thus the damped value of the dynamic resistance can be taken as equal to the external damping resistor, i.e. 2000 ohms. Now we know that
dynamic resistance $=$ inductive or capacitive reactance $\times Q$
and this gives
inductive or capacitive reactance


Substituting the numerical values appropriate to our example inductive or capacitive reactance

$$
\begin{aligned}
& =\frac{2000}{6.1} \\
& =328 \mathrm{ohms}
\end{aligned}
$$

The problem now is to determine the $L$ and $C$ values which have a reactance of 328 ohms at 36.5 MHz . The method of doing this has already been described; the results are:

$$
\begin{aligned}
L & =1.43 \mu \mathrm{H} \\
\mathrm{C} & =13.3 \mathrm{pF} .
\end{aligned}
$$

Bandwidth of two coupled circuits. A better approximation to the ideal square-topped frequency response can be obtained from two similar coupled $L C$ circuits tuned to the same frequency and many receivers incorporate such pairs of $L C$ circuits in the i.f. amplifier. The shape of the frequency response of such a pair is determined by the degree of coupling between them and this is measured by the coupling coefficient $k$, defined as $M / L$ if the coupling is


Fig. 6. Illustrating the response of a pair of overcoupled tuned circuits.

Fig. 7. Two simple methods of coupling two tuned circuits (a) by shunt capacitance and (b) by series capacitance.
achieved, as in i.f. transformers, by mutual inductance between the coils. For a signal injected into one of the tuned circuits, maximum response is obtained from the other when $k=1 / Q$, known as critical coupling. The frequency response is also the flattest for critical coupling. If $k$ is less than critical the response is low and is single-peaked. If $k$ is greater than critical the response has two peaks of equal height, the frequency separation between them being approximately $k f_{o}$. Values of $k$ slightly greater than critical are often employed in i.f. circuits and for such values the passband can be taken as $\sqrt{ } 2 k f_{o}$ as indicated in Fig. 6.

There are many ways of coupling coils to achieve the response of Fig. 6, and for a number of these there is a simple expression for the coupling coefficient so that it is easy to achieve a desired value of passband. Two coupling methods are shown in Fig. 7: at (a) coupling is achieved by a common shunt capacitor $\mathrm{C}_{1}$ and at (b) by a common series capacitor $\mathrm{C}_{2}$. For circuit (a) the coupling coefficient is given approximately by

$$
k=\frac{C}{C_{1}}
$$

and for circuit (b)

$$
k=\frac{C_{2}}{C}
$$

As a numerical example we will consider the design of a pair of coupled
coils for use in the i.f. stage of an f.m. receiver. The standard centre frequency is 10.7 MHz and the bandwidth required is 200 kHz . Thus from Fig. 6 we have

$$
k=\frac{\text { bandwidth }}{\sqrt{ } 2 f_{0}}=\frac{0.2}{\sqrt{ } 2 \times 10.7}=0.013
$$

If we assume the tuning capacitors to be 200 pF then we have, for shunt-capacitance coupling

$$
\begin{aligned}
C_{1}=\frac{C}{k} & =\frac{200}{0.013} \mathrm{pF} \\
& =0.015 \mu \mathrm{~F}
\end{aligned}
$$

and for series-capacitance coupling

$$
\begin{aligned}
\mathrm{C}_{2} & =k \mathrm{C}=0.013 \times 200 \mathrm{pF} \\
& =2.6 \mathrm{pF} .
\end{aligned}
$$

shape of the response curve achieved by either of these two methods of coupling depends on the $Q$ value for the inductors. For $k=0.013$ coupling is critical if $Q=1 / k=75$. For this value of $Q$ therefore the response is flat near $f_{0}$ but the required bandwidth is not achieved. The $Q$ values must exceed 75 to give the bandwidth wanted and the response then has two peaks. The height of the peaks should not exceed the centre-frequency response by too great an amount but the difference will be less than 1 dB provided $Q$ does not exceed 110:

## MF predictions

Following its first signs of life in September last the new sunspot cycle should become clearly established this month. Sunspot number will rise with increasing rate during 1976 and may be as high as 50 by the end of the year.
In retrospect of 1975 it was most noticeable that transfer of international circuits from h.f. to satellite systems over the past decade had drastically reduced co-channel interference problems previously encountered during a sunspot minimum year. However did we manage with about 300 unstable channels for the whole world?




## Linear c.m.o.s. circuits

# This article gives background to the linear use of c.m.o.s. circuits which appear in set 27 of Circards. 

by J. Carruthers, J. H. Evans, J. Kinsler and P. Williams.

Paisley College of Technology

By considering the various families of logic circuits one can deduce what parameters, designed-in to optimize digital operation, result in adverse performance if the circuits are used in the analogue mode. The most interesting family that is available at a low cost is complementary symmetry metal oxide semiconductor logic or c.m.o.s. By adding an extra diffusion to the processing steps needed for a single type of device, Fig. 1, the complementary type can be produced. Thus the p-channel device is obtained by diffusing $\mathrm{p}^{+}$ source and drain into the n-type substrate; a deeper p-type well is diffused to contain the $n$-channel device.
In addition to the conducting channels between the respective sources and drains, there are a number of p-n junctions. Though these are normally reverse-biased by the selected operating mode, their parasitic effects cannot always be ignored. Other p-n junctions are deliberately introduced at the inputs to prevent damage to the thin oxide films by high voltages that might be produced electrostatically or due to external transients.
The conductivity of the channels is increased by a forward bias on the gates. For an n-channel device this corresponds to a gate voltage positive with respect to the source. The conductivity tends to zero for zero gate-source voltage because the doping of the channel is such that there are virtually zero current carriers available. At a particular value of gate voltage, called the threshold voltage, charges induced in the channel by the resulting field allow conduction to commence. No gate current flows because of the insulation provided by the oxide layer. Drain-source conductivity continues to increase with rising gate-source voltage, while the current flow depends on both $V_{\mathrm{gs}}$ and $V_{\mathrm{ds}}$. At low values of $V_{\mathrm{ds}}$, the slope resistance is reasonably linear and is controlled by $V_{g s}$. Ultimately, the output current reaches a limiting value again set by $V_{g s}$ i.e. the output slope resistance becomes very high making the device suitable for various forms of constant-current circuits.
The basic inverter stage is shown in


Fig. 1. Structure of c.m.o.s. inverter.


Fig. 3. Transfer function of, c.m.o.s. inverter.

Fig. 2 and the transfer function in Fig. 3. When the input voltage is low, there is insufficient forward bias on the n -channel device to bring it above the threshold of conduction. The reverse is true for the p-channel device which has a high conductivity i.e. the output is virtually equal to $+V_{\mathrm{s}}$. Conversely a very positive input voltage pushes the output close to zero. When lightly loaded the output swings to within $1 \%$ of the supply voltage. At voltages in the region of $+V_{s} / 2$ both transistors are conducting. This region is traversed rapidly when the inverter is used as a logic level inverter, but the transient current pulse as shown in Fig. 4 is then the only significant contribution to. current drain, as one or other of the transistors is non-conducting in each of the logic states. (At high pulse rates power consumption is associated with the multiple charge/discharge of internal and load capacitances.)

For analogue systems this central region of the characteristic is of great interest. The slope is moderately high, corresponding to a voltage gain of the order -10 to -100 with the unusual property that the gain is greater at lower values of supply voltage (both transconductance and output conductance fall, the last-mentioned more rapidly than the first).
Although not designed for high d.c. stability, temperature dependence of the transfer function is small. Recently c.m.o.s. i.cs have been produced designed specifically for analogue applications; though the configurations may be identical to particular logic circuits, the processing is optimized for linear operation.
Because each m.o.s. transistor can be used as a voltage-controlled resistor, packages containing several devices offer interesting possibilities. The drain-source resistance characteristic remains approximately linear for small values of reverse voltage and current, making a.c. operation feasible. The devices on a common chip will be similar and will be subject to the same temperature variations. Thus, operated from a common gate-source voltage supply, they offer closely-matched resistance characteristics for use in amplifiers etc.
An interesting extension of this technique is possible even where single devices are not accessible separately. By grounding the positive supply line of a package containing multiple inverters, the p-channel devices are kept out of conduction provided the p.d. applied between output and ground of each device is small. Thus a hex buffer inverter i.c. can be used as a set of six matched $n$-channel f.e.t.s.
Any inverting amplifier can be used with n.f.b. to give a see-saw amplifier of reduced but well-defined voltage gain. As only a single gain-stage is involved external compensation against highfrequency stability is not required even

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## New circuit book

"Circuit design - 1, Collected Circards" brings together the first ten sets of Circards, introductory articles to each of the subjects, and ten pages of additional circuits. The hardback A4 book contains 168 pages, in which 120 cards are rearranged so that each is laid out on one page. A brief introduction precedes the articles, which were previously published in Wireless World, and each of the ten subjects is followed by an up-dating page. Corrections have been incorporated where appropriate. "Circuit designs" is obtainable through leading bookstalls at $£ 10$ per copy. In case of difficulty order direct by sending remittance for $£ 10.40$ (includes postage and packaging) to the address given later, making cheques payable to IPC Business Press Ltd.

Topics covered so far in Circards are: 1 active filters
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8 astable multivibrator circuits
9 optoelectronics: devices and uses
10 micropower circuits
11 basic logic gates
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17 c.d.as - signal generation
18 c.d.as - measurement and detection
19 monostable circuits
20 transistor pairs
21 voltage to frequency converters
22 amplitude modulators
23 reference circuits
24 voltage regulators
25 RC oscillators-1
26 RC oscillators-2
27 Linear c.m.o.s.-1
28 Linear c.m.o.s.-2

Circards are a unique way of collating and presenting data about circuits in a compact and easily retrievable way. The sets of $203 \times 127 \mathrm{~mm}(8 \times 5 \mathrm{in})$ doublesided cards are designed for easy filing in standard boxes and for easy access at the desk or at the bench, where transparent plastic wallets keep the cards in good condition.
Each card normally describes operation of a selected circuit, gives measured performance data and graphs, component values and ranges, circuit limitations and modifications to alter performance. Suggestions for further reading are included together with cross references to related circuits. The Circard concept was outlined more fully in the October 1972 issue of Wireless World, pp.469/70.
with $100 \%$ negative feedback. The bandwidth is not particularly high though well above the audio-frequency range, with some gain to beyond 1 MHz . High values of resistors may be used without loading effects due to amplifier input impedance, though the resulting RC time constants due to strays can further reduce the bandwidth.

A recent example of the way in which device technologies change forces a warning note at this point. To improve the performance of gates/inverters for their main functions in purely digital applications, some manufacturers pro-
duce "fully-buffered" versions. These might contain three inverters in cascade to perform the function previously using a single inverter. The improved response is welcome to users with critical requirements in the digital field; those wishing to adapt them for analogue circuits, as decribed in Circards 'Set 27, would find high-frequency instability resulting from the much-increased loop gain and multiple phaseshifts. Remember to check the characteristics of any device or circuit rather than assume that similar titles guarantee identical performance.


## Wire wrapping tool

The model EW-7D-48U is an electrically driven wire wrapping and unwrapping tool. Particularly designed for use in telecommunications, the motor operates from a $48-50 \mathrm{~V}$ d.c. power source. Conversion from the wrapping to unwrapping mode is achieved through the operation of a switch in the handle of the tool.

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WW 301 for further details

## Band-reject filters

Telonic has introduced six tunable band-rejection filters which will cover the $50-1000 \mathrm{MHz}$ frequency spectrum in one-octave bands.

Designated the Series TTR, the new filters are tuned by a single knob, which has direct frequency readout marked on the dial. The filters are suited for


WW 301 for further details
electromagnetic compatibility testing, or for many other laboratory functions. Each filter is tunable over a one-octave tuning range (up to 1 GHz ) and offers an insertion loss of less than ldB. The passband v.s.w.r. is less than 1.5 over the entire tuning range, the notch bandwidth being $4-6 \mathrm{MHz}$ wide at the 50 dB bandwidth and a minimum of 100 kHz at the 1 dB bandwidth. Ultimate attenuation is claimed to be a minimum of 50 dB with typical values of $60-70 \mathrm{~dB}$. The power rating is 50 W . Telonic Altair UK, 2 Castle Hill Terrace, Maidenhead, Berks.
WW 302 for further details

## Watch-timer

Designed to check the accuracy of all types of wrist-watch, the Quartz-timer II is fitted with three types of probe to effect a suitable signal coupling to the watch. It will also provide a power supply for electric watches under test, should this be required. The measuring amplifier can be bandwidth-limited by selecting the appropriate mode switches, and measuring periods of 2,4 , 6 and 12 seconds are also available.

A two-figure digital display with plus or minus error signals is provided, together with an analogue meter which is switchable to read the power supply voltage or average current consumption, and which can also be used to measure the signal level obtained from the probe. Greimer Electronic (GB) Ltd, 182 Upper Richmond Rd, Putney, London SW15 2SH.
WW $\mathbf{3 0 3}$ for further details

## Soldering pistol

The Kager soldering iron pistol features the automatic application of solder to the iron tip, in measured quantities. The tips used in the range are nonscaling and recent trials by the manufacturer


WW 302 for further details

have produced the claim that the life of each tip is well in excess of 10,000 joints. The Kager 3000 range of pistol irons is available in various wattages and voltages and an additional option of electronic temperature control is offered. Paxk Star Engineering Ltd, 62, High St, Croydon CR9 2UT, Surrey.

## WW $\mathbf{3 0 4}$ for further details.

## Op-amp tester

The THETA operational amplifier tester type 255 is a semi-automatic system for the measurement of the characteristics of d.c. operational amplifiers. It will measure bias current, offset voltage, supply current and the transfer function of the device under test, with a minimum of adjustments, and may be used either for laboratory quantitative measurements, for quality assurance tests or for production limit testing.

The instrument is powered by $110 / 240 \mathrm{~V}$ mains and provides the stabilized supplies, waveforms and solidstate switching needed to make the tests. The 255 is designed to be used in conjunction with a directly-coupled oscilloscope for display purposes.

Plug-in modules are used to interface the device to be tested with the tester, these carrying any frequency compensation or other networks necessary for correct operation. Expander modules are also available for 24,28 and 36 -pin packages. THETA, PO Box 10, Martock, Somerset, TAl2 6LT.

WW 305 for further details

## R.f test system

The models 9600 /9650 are two versions of a radio-frequency tracking sweeper-analyser, designed for operation in the frequency range from 1 MHz to 350 MHz . The unit incorporates a sweep generator and tracking spectrum


WW 304 for further details


WW 306 for further details


WW $\mathbf{3 0 7}$ for further details


WW 308 for further details


WW 309 for further details
analyser together with a display unit, reflection-coefficient bridge, comparator and gain and loss measuring attenuators.

The two versions of the system available are for $75 \Omega$ operation (Model 9600 ) and $50 \Omega$ operation (Model 9650). Both units have an 80 dB dynamic range, a sensitivity of $1 \mu \mathrm{~V}$ and a sweep-system level accuracy of $\pm 0.25 \mathrm{~dB}$ over the whole frequency range. Additional features include automatic phase lock, a 50 dB bridge balance for return loss measurements, a 120 dB range for isolation and crosstalk and provision for up to six plug-in crystal markers. Texscan Instruments Ltd, 1 North Bridge Rd, Berkhamsted, Herts.

WW $\mathbf{3 0 6}$ for further details

## Panel indicator

Details have been received of a frontmounting panel indicator, from H. F. Collison-Goodwell. These are miniature tungsten filament or neon bulbs, encapsulated in a clear plastic holder which is designed to clip directly into a 6 mm hole. Types available include $5 \mathrm{~V}, 60 \mathrm{~mA} ; 12 \mathrm{~V}, 30 \mathrm{~mA} ; 28 \mathrm{~V}, 40 \mathrm{~mA}$ and $120-240 \mathrm{~V}$ with integral resistor. A variety of colours are also offered and the lamps are fitted with insulated and tinned leads. H. F. Collison-Goodwell Ltd, Coleshill, Birmingham B46 3BL.

WW 307 for further details

## Mini drills

The Expo Titan and Reliant are miniature 12 V d.c. drills suitable for printed circuit boards, jewellery, model making, etc. The first mentioned measures $114 \times 41 \mathrm{~mm}$ and is supplied with four collets of up to 3 mm capacity, while the Reliant measures $76 \times 34 \mathrm{~mm}$ and is supplied with three collets of up to 2.4 mm capacity. With the aid of a tool kit ( 20 or 40 piece) the drills can polish, brush, burr, saw and grind almost any material at speeds up to $9000 \mathrm{rev} / \mathrm{min}$. Both drills can be fitted to a multipurpose drill stand which can also be used as a miniature lathe or horizontal bench stand. For operation from 240 V a.c. supplies, a mains power unit is available. These accessories can be obtained with the drills in various kit options from A. D. Bayliss \& Son Ltd, Pfera Works, Redmarley, Gloucester GL19 3 JU

WW 308 for further details

## IMPATT oscillator

The ATO range of c.w. millimetre wavelength IMPATT signal sources from Plessey has now been extended to include the frequency band $60-90 \mathrm{Ghz}$. These devices are designated the ATO 260 series and three basic models provide 20,50 and 100 mW c.w. Designed
for fixed frequency operation, the sources can be supplied to any specified frequency in the range, with small adjustments, typically $0.5 \%$, made possible by means of a tuning screw.

Standard units are provided with a UG387/U output waveguide flange and an SMA connector for bias. Typical bias requirements are 20 V at 100 mA . Plessey Optoelectronics and Microwave, Wood Burcote Way, Towcester, Northants.

WW 309 for further details

## Scope probe kit

The latest addition to the Interprobe range of oscilloscope passive probe kits, is the triple-function Interprobe-3, giving $\times 10$ attenuation up to $80 \mathrm{MHz}, \times 1$ attenuation up to 15 MHz and an earth reference check selected by a threeposition switch built into the probe body.

Features include a very flexible cable, a fixed earth-lead, a sprung hook tip and an input capacitance range of $15-50 \mathrm{pF}$. Intertek, 15 Ennerdale Avenue, Dunstable, Beds, LU6 3AR.

## WW 310 for further details

## Non-flam capacitors

Designed to conform with B415, Clause 20 (Fire prevention in TV receivers), this range of plastic encapsulated, metallised film capacitors will self-extinguish in one second after being subjected to the BS flame test.

The capacitance range is from $0.01-4.7 \mu \mathrm{~F}$ in working voltages from 100 V to 200 V , a.c. and tolerance ranges $\pm 5,10$ and $20 \%$.
The capacitors also carry all the BS and PO approvals granted to the PMT range. ITT Components, Capacitor Division, Brixham Rd, Paignton, Devon.

WW 311 for further details

## Low-current dry-reed relays

The ARME series of dry-reed relays are sealed in a d.i.l. 14-pin package, suitable for socket or p.c.b. mounting. Satisfactory operation is claimed from smallsignal sources, up to a maximum rating of 10 VA . The initial contact resistance is typically $200 \mathrm{~m} \Omega$ and the contact life expectancy is said to be 100 million operations at low signal levels, to 5 million operations at the maximum rated load.

Response time for the relays is typically $500 \mu \mathrm{~s}$, the coil drawing 13 mA from a standard 5 V logic supply rail. The standard version is designated Form A and has a single contact, normally open. Other variants are offered featuring electrostatic shielding between the coil and the capsule and/or an integrated diode which is cathode
connected to the coil, the anode being connected to an otherwise unused pin. Elliot Relays, 70 Dudden Hill Lane, London NW 10 IDJ.
WW 312 for further details

## Pyrometer

A pyrometer has been developed to check the temperature of soldering iron bits. It uses an iron-constantan thermocouple in conjunction with a moving coil meter, incorporating automatic cold junction compensation. The meter scale ranges from zero to $500^{\circ} \mathrm{C}$ with a scale mark every $20^{\circ} \mathrm{C}$. The pyrometers are supplied with a calibration certificate having been tested against the standard e.m.f. of iron-constantan to BS4937, part 3. Light Soldering Developments Ltd, 97-99 Gloucester Road, Croydon CR0 2DH.
WW 313 for further details

## Capacitance meter

The DCM302 is a portable capacitance meter ranging over six decades of measurement from 1999 pF to $199.9 \mu \mathrm{~F}$ with a resolution of $0.05 \%$ being given by the $3^{1 / 2}$ digit display.

The only control is a touch-pad used to start measurement, turn off occurring automatically after about 10 s . Power is derived from two $9 \mathrm{~V}, \mathrm{PP} 3$ batteries giving an operating life of 6,000 measurements. The l.e.d. display has 7.5 mm high characters, full scale range being indicated by a separate l.e.d. Price is $£ 89$ direct from the manufacturers. Aim Cambridge Ltd, Nuffield Road Industrial Estate, St. Ives, Huntingdon, Cambs.

WW 314 for further details

## Sweep generator

The ED3 is a sweep generator covering the frequency range from 20 kHz to 70 MHz in ten overlapping bands. Sockets are fitted for external sweep control, an external marker generator, an external r.f. generator and d.c. level shift. Outputs include the video signal, sync and the time base signal.

An internal marker generator provides for a selection of up to seven crystal derived frequency combs from 10 kHz to 10 MHz . Aspen Electronics Ltd, 18a High St, Northwood, Middlesex HA6 1BN.

WW 315 for further details

## Cycle counter/timer

Capable of counting at rates up to 5 kHz , the Koyo KCD 5 H is a counter timer, with pre-setting obtained from a five digit thumbwheel indicator.

Signal sources may be from a variety of transducers and the counter can be


WW 310 for further details


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WW 315 for further details
reset from an external source. Two types of output are available, a 50 ms pulse or a d.c. level, either of which appears when the count equals the thumbwheel setting.

Noise immunity is improved by filtering and the supply voltage is mains derived ( $220 / 240 \mathrm{~V}, 50 / 60 \mathrm{~Hz}$ ). Tempatron Ltd, 5 Loverock Rd, Reading, Berks.

## WW 316 for further details

## Jacketed flat cable

Thomas \& Betts have introduced a jacketed flat cable as an addition to the Ansley flat cable and connector system. The cable has been developed for external cabinet-to-cabinet. wiring requiring low crosstalk characteristics, such as in high-speed data processing equipment.

To terminate the cable, a small length of outer jacket is removed and the insulated core of the cable is fitted into a standard Ansley insulation-piercing connector. Termination is effected using either a bench or hand type Ansley tool.

The inner construction of the cable is ribbed, making location of the connector simple and every fifth conductor is colour coded as a method of identifying conductors. The number one conductor is identified by a dark-blue strip, which also provides a "polarity" check. The cable is available in 10 to 50 wire sizes and has a UL temperature rating of $75^{\circ} \mathrm{C}$ and a voltage rating of 90 V . Thomas \& Betts Ltd, Sedgewick Rd, Luton LU4 9DT.

WW 317 for further details

## Solid Stafe Devices

Names of suppliers of devices in this section are given in abbreviation after each entry and in full at the end of the section.

## Darlington transistors

The RCA 2 N 6534/5/6/7 are a range of monolithic n-p-n silicon Darlington power transistors produced in TO-66 packages. The devices are made using the double epitaxial construction to produce good forward and reverse second breakdown characteristics; they are suited to power applications at low and medium frequencies.

Since they have gains up to 1000 at 5A, the devices can be operated directly from integrated circuit outputs without
the use of a pre-driver. Versions are available for 80,100 and 120 V all with power dissipation of 36 W . They can handle a continuous collector current of 8 A , with peak collector current up to 15A.

RCA

## P.s.u. transistor

The latest addition to the Mullard range of transistors for use in high-frequency switched-mode power supplies operating directly from a "rectified mains" input, is the type BUX82. It is intended for use in 400 W push-pull or 100 W to 200W single-ended circuits. The BUX82, in common with the other types in the range will accommodate the $10 \%$ voltage variations found on mains supplies.

To meet these requirements, the device has an open base, collectoremitter rating of 400 V and a collector breakdown rating of $800 \mathrm{~V}\left(\mathrm{~V}_{\mathrm{BE}}=0\right)$. The d.c. and peak collector current ratings are 5 A and 8 A respectively.

The switching characteristics are designed to minimize switching losses and to facilitate operation in the $25-50 \mathrm{kHz}$ region.

Mullard

## Microprocessor

The Am2901 is the first of a family of low-power Schottky circuits planned by Advanced Micro-Devices for use in a variety of applications, including micro-programmed computers. This particular device is a four-bit microprogrammable data chip containing an eight-function arithmetic logic unit, a two-port 16 -word scratch pad memory, an additional accumulator register and shifting and control logic. This circuit can execute a total of 12992 instructions and performs a read-modify-write cycle in less than 100 ns .

Advanced

## Decoder/driver

The CD4511 is a b.c.d. decoder/driver with an integral latch on a single chip. The decoding functions and the latch use the c.m.o.s. process and the output devices are bipolar. This enables the device to supply up to 25 mA to drive a display segment. The CD4511 provides the functions of a 4-bit storage latch, a b.c.d.-to-seven-segment decoder and seven high-current drivers. Lamp test, blanking and latch inputs are provided for display testing, turning off or brightness modulation of the display and storing a b.c.d. code respectively.

The CD4511 is available in either 16-pin epoxy-B d.i.l. or 16 -pin ceramic packages.

National

## GaAs light source

The type GL-504 lamp is designed to produce a power dissipation of 140 mW
at room temperatures and is a hermeti-cally-sealed unit in a metal can, fitted with a glass window. The peak emission wavelength is $2 \AA$ and when used with a pulse circuit delivering a pulse width of $10 \mu \mathrm{~s}$, repeated at 100 Hz , will produce an output capable of detection over a distance of 100 metres.

The device measures 4.6 mm dia, by 6.6 mm high, including the glass window. Price is $£ 2$ to $£ 3$ each according to quantity. (Plus VAT at 8\%)

Photain

## High voltage op-amp

The HA2-2645-5 operates from dual supply rails between 10 V and 40 V (up to 80 V total) and incorporates a temperature sensing circuit for output current limiting. Slew rate is $5 \mathrm{~V} / \mu \mathrm{s}$ in the unity-gain configuration, 23 kHz full power bandwidth is 23 kHz and the device provides a typical 12 mA output current. Up to plus and minus 35 V can be produced across a $5 \mathrm{k} \Omega$ load.

Input characteristics for the internal-ly-compensated op-amp include 40 nA maximum bias and offset currents over the full temperature range from $0^{\circ} \mathrm{C}$ to $+75^{\circ} \mathrm{C}$, a $40 \mathrm{M} \Omega$ minimum input resistance and plus and minus 35 V common mode range.

GDS

## 8k r.a.m. board

Motorola have announced the introduction of an 8 k r.a.m. card comprising two 4 k byte blocks of dynamic random access memory. This memory is for use with their M6800 evaluation and development systems and features a standby battery power supply enabling the contents of the memory to be retained for more than a week from a 12 V battery of 4.5Ah capacity.

The memory card is designed to provide a temporary store for the assembly of programmes, prior to loading where high speed loading arrangements such as a floppy disc with direct memory access, or a high speed cassette reader are not available.

Motorola

## Suppliers

RCA Ltd, Solid State Europe, Sun-bury-on-Thames, Middlesex.
Mullard Ltd, Mullard House, Torrington Place, London WC1.
Advanced Microdevices Inc. 901 , Thompson Place, Sunnyvale, California 94086, USA.
National Semiconductor UK Ltd, 19 Coldington Rd, Bedford MK40 3LF.
Photain Controls Ltd, Unit 18, Hangar 3, The Aerodrome, Ford, Sussex.
GDS Sales Ltd, Michaelmas House, Salt Hill, Bath Rd, Slough, Bucks.
Motorola Semiconductor Products Division, York House, Empire Way, Wembley, Middx.

## Conferences \& Exhibitions

## LONDON

Mar. 23-25
Imperial College
CAD 76 (Computer Aided Design)
(Tom McGowran, CAD 76, IPC Science \& Technology Press, 32 High Street, Guildford, Surrey GUl 3EW)

## Mar. 30-31

Savoy Place
Small Electrical Machines
(IEE Conference Department, Savoy Place, London WC2R 0BL)

Apr. 13-15
Grosvenor House
The All-Electronics Show
(The All-Electronics Show, 34-36 High Street, Saffron Walden, Essex)

May 3-7
Savoy Place
Subscriber Loops and Services
(IEE Conference Department, Savoy Place, London WC2R 0BL)

May 10-14
Royal Lancaster Hotel
7th IMEKO Congress
(The Institute of Measurement and Control, 20 Peel Street, London W8 7PD)

## June 22-25

Savoy Place
On-line Operation and Optimisation of Transmission and Distribution Systems
(IEE Conference Department, Savoy Place, London WC2R 0BL)

July 5-8
Savoy Place
Automobile Electronics
(IEE Conference Department, Savoy Place, London WC2R 0BL)

Sept. 1-3
Savoy Place
Advances in Magnetic Materials and their Applications
(IEE Conference Department, Savoy Place, London WC2R 0BL)

Sept. 16-18
Savoy Place
Optical Fibre Communication
(IEE Conference Department, Savoy Place, London WC2R 0BL)

Sept. 20-24 Grosvenor House
International Broadcasting Convention
(IEE Conference Department, Savoy Place, London WC2R 0BL)

Nov. 9-12
Savoy Place
Millimetric Waveguide Systems
(IEE Conference Department, Savoy Place, London WC2R 0BL)

## Nov 22-25 Savoy Place

The Future of Aircraft All-weather Operations
(IEE Conference Department, Savoy Place, London WC2R 0BL)

Nov. 30-Dec. 2
Mount Royal Hotel
Electronic Displays '76
Network, 84 High Street, Newport Pagnell, Bucks MK16 8EG)

## AVIEMORE

Sept. 20-24 Coylumbridge Hotel
Impact of New Technologies in Signal Processing (IEE. Savoy Place. London WC2R 0BL)

## BIRMINGHAM

May 3-7
National Exhibition Centre
IEA-Electrex
(Industrial and Trades Fairs Ltd.. Radcliffe House, Blenheim Court, Solihull, West Midlands, B91 2BG)

May 23-31 National Exhibition Centre HEDA followed by Sound \& Vision '76
(Montbuild Ltd., 11 Manchester Square, London W1M 5AB)

July 20-22 University of Birmingham
Video and Data Recording
(IERE, 8-9 Bedford Square, London, WCIB 3RG)

## BRIGHTON

Mar. 29-30
University of Sussex
Materials and Processing Effects in Semiconductor Devices
(The Institute of Physics, 47 Belgrave Square, London SWIX 8QX)

June 8-11
Metropole Convention Centre
Communications '76
(IEE Conference Department, Savoy Place, London WC2R 0BL)

## BRISTOL

July 12-16
University of Bristol
Liquid Metals
(The Institute of Physics, 47 Belgrave Square, London SWIX 8QX)

## CAMBRIDGE

Apr. 17-19
Churchill College
Quantimet Image Analysis Techniques
(Miss E. J. Jones, Image Analysing Computers Ltd., Moat House, Melbourne, Royston, Herts SG8 6EJ)

June 28-July $2 \quad$ University of Cambridge
IERE Golden Jubilee Convention
(IERE, 8-9 Bedford Square, London WCIB 3RG)

## EDINBURGH

Sept. 19-22 University of Edinburgh
Gallium Arsenide and Related Compounds
(Institute of Physics, 47 Belgrave Square, London SWIX 8QX)

## MANCHESTER

Jan. 5-7
University of Manchester
Solid State Physics
(The Institute of Physics, 47 Belgrave Square, London SWIX 8QX)

## SOUTHAMPTON

## Apr. 6-8

Southampton University
Applications of Electronics in Medicine
(IERE, 8-9 Bedford Square, London WC1B 3RG)

## STONELEIGH

Feb. 24-26
National Agricultural Centre
Audio/Visual in the Midlands
(D. L. Ford Exhibitions (1968) Ltd., National Agricultural Centre, Stoneleigh, Warwickshire)

## SWANSEA

Sept. 7-9
Gas Discharges
(IEE Conference Department. Savoy Place, London WC2R 0BL)

## YORK

Sept. 21-23
Optics
(The Institute of Physics, 47 Belgrave Square, London SWIX 8QX)

## OVERSEAS

Jan. 20-22
Las Vegas
Reliability and Maintainability
(The American Society of Mechanical Engineers, 345E 47th Street, New York, N.Y.10017, U.S.A.)

Feb. 18-20
Salzburg
Software Engineering for Telecommunication Switching Systems
(IEE Conference Department, Savoy Place. London WC2R OBL)

Mar. 2-5
Zurich
53rd AES Convention
(Audio Engineering Society, 60 East 42nd Street New York, N.Y.10017, U.S.A.)

Mar. 8-14
Paris
Festival Intermational du Son
(S.D.S.A., 20 rue Hamelin, 75116, Paris, France)

Apr. 5-10
Paris
Paris Components Show
(Société pour la Diffusion des Sciences et des Arts, 20 rue Hamelin, F75116 Paris, France.

Apr. 22-26
Budapest
Acoustics
(Optical, Acoustical and Filmtechnical Society, H-1061 Budapest, Anker Koz 1, Hungary)

Apr. 28-May 6
Hanover

## Hanover Fair

(Deutsche Messe- und Ausstellungs-AG, P.O. Box 283, 78 Riddlesdown Road, Purley, Surrey CR2 1 YB )

May 1-9
Hanover
Ith German Aerospace Show
(Deutsche Messe- und Ausstellungs-AG. P.O. Box 283, 78 Riddlesdown Road, Purley. Surrey CR2 1 YB)

May 4-7
Los Angeles
54th AES Convention
(Audio Engineering Society, 60 East 42 nd Street, New York, N.Y.10017, U.S.A.)

May 11-24
Boston
Electro 76
(IEEE, 3600 Wilshire Boulevard, Los Angeles, California 90010 , U.S.A.)

May 22-26
Brussels
Euromation (Instrumentation and Automation)
(BIRA, Jan Van Rijswijcklaan 58, B-2000 Antwerp. Belgium)

June 2-4
Atlantic City
30th Annual Frequency Control Symposium
(Department of the Army, Headquarters United States Army Electronics Command, Fort Monmouth', New Jersey 07703, U.S.A.)

## June 15-18

Basle
Medex 76
(Sekretariat Medex 76. Postfach, CH-4021 Basle, Switzerland)

## Aug. 3-6

Toronto
Computer Communication
(Dr Pramode K. Verma, Programme Chairman, ICCC-76, P.O. Box 365, Station 'A', Ottawa, Ontario, Canada KIN 8V3)

Aug. 27-Sept. 5
Amsterdam
Firato 76
(RAI, Europaplein. Amsterdam, The Netherlands)
Sept. 6-11
Namur
International Congress of Cybernetics
(Association Iriternationale Cybernetics, Palais des
Expositions, Flace Andre Rijckmans, BM-5000 Namur. Belgium)

Sept. 7-10 Geneva
European Conference on Circuit Theory and Design
(Istituto Elettrotecnico Nazionale Galileo Ferraris,
Corso Massime d'Azeglio. 42, 10125 Torino, Italy)

Sept. 14-17
Rome
Sixth European Microwave Conference 76
(Prof. Ing. P. de Santis, Conference Chairman. Selenia S.p.A.. Via Tiburtina Km. 12.400, 00131 Rome, Italy)

Sept. 13-15
Eindhoven
Magnetic Bubbles
(P. F. Bongers, Philips Research Laboratories. Eindhoven. The Netherlands)

Oct. 5-8
Geneva
3rd European Electro-Optics Conference and Exhibition
(Mack-Brooks Exhibitions Ltd., 62-64 Victoria Street, St. Albans, Herts AL7 3XT)

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by "Vector"

## "CHRISTMAS IS COMING (or, The Haunted Power Station)

In days of yore it was the pleasant custom for journals to produce what were invariably described as Bumper Xmas Numbers. Large dollops of snow adorned the title on the front cover and holly ran riot everywhere. Inside, normally staid contributors let their hair down and a jolly time was had by all.
With a deferential tug of the forelock I suggested to the Editor that he might revive the practice and combine amusement with instruction. Perhaps Cathode Ray, I said, who has in time past clarified so many matters that perplexed us, might be persuaded to direct his genius to unravelling that Law, known to all electronics engineers by a name which Sir won't let me use. You know - the one which states "That if anything can possibly go wrong it assuredly will, and at the most inconvenient time". And as another feature, I said, why not include the scenario of a pantomime? Heaven knows, the electronics industry writes its own, year after year, complete with cast . . .
At this point the Editor was stricken with apoplexy and had to be revived by his retinue of ravishing secretaries. I won't touch on his subsequent diatribe, which, shorn of its adjectival content, was to the effect that he didn't agree with me.

Tell you what, fellow traditionalists. Christmas Eve is the time for gathering around the log fire and telling ghost stories, so leave what follows until Dec. 24 and then read it.

All right, then. Everybody comfy around the oil-fired radiator? Good. Well then, you may remember that in the November ' 74 issue I wrote about the odd experiences of Jim, a friend of mine, while investigating an old woollen mill site in the West Country, where his forebears lived. In all, he's now made three visits on successive annual holidays. On the first occasion
he removed a piece of stone from the ruins; this was subsequently psychometrised under experimental conditions and in due course an astonishingly accurate description of the site was given. This mentioned " $a$ strong atmosphere of decay" and described it as "a place of sinister happenings", with death by drowning as a specific instance.
Jim recalled that his mother had told him of the drowning of a little girl in the mill leat and had said that the cottage was haunted. On one occasion she'd been taken, as a small child, to visit her grandparents at the mill cottage. One afternoon, she and her mother had been alone in the house when they heard footsteps cross the bedroom above and descend the stairs. A pause, and then the latch of the stair door lifted - at which her mother snatched her up and ran out of the house.
On his second visit Jim, more or less by chance, found that the granite wall surrounding the site was markedly radioactive - not to anywhere near a dangerous degree but very much higher than the norm for the area: As he wasn't able toget into the site on that occasion, the question of whether the wall represented the peak of radiation or whether a higher source lay inside had to remain unanswered.
This old mill site was, as long ago as 1891, acquired by an electricity supply undertaking, which installed a waterdriven generator. The original plant has long since disappeared but the Central Electricity Generating Board still have a water-driven generator there; it's almost certainly the smallest CEGB station in the country. This occupies only a part of the old mill site; the rest being mostly wilderness, but it does explain the granite wall and the need to keep people out.
On his first visit Jim had obtained permission from the SWEB area representative, who, he'd been astonished to find, was a school-fellow of almost half-a-century ago. On the second visit, circumstances forbade a prior arrangement, but this year, on making application, he found that his old school acquaintance had retired and that the succeeding SWEB representative was the son of another old school-fellow! Coincidence number three occurred on the site when, in conversation, he found that in 1930 the power station had been acquired by an electrical firm in his home-town, 250 miles away.

As to the gamma radiation tests on the third visit, nothing sensational materialised. The highest reading was obtained from the ruined wall of the cottage; no peaks were obtained from the ground itself, which is not surprising as I believe that even a few feet of earth will severely attenuate gamma radiation. Possibly then, a local quarry may be the "hot" source, rather than the site itself.

So much for facts as registered by an electronic instrument, but now for the relative quicksands of the unknown. On this occasion the party visiting the site was eleven strong and included the SWEB representative, a professor of electrical engineering and an electronics research engineer, the last-mentioned being a colleague of Jim's.
In the course of the tests it was noticed that the West Highland terrier belonging to Jim's colleague was avoiding the cottage area, although happily trotting around the rest of the site. On several occasions her master stood by the cottage wall and called her; normally obedient, each time she'd approach to within 20 yards, then slow and stop; her tail would droop, she'd exhibit signs of distress and on no account would she come nearer.
Sinister happenings? Well, there was the drowning in the leat. Then there's the grisly circumstance related by an elderly villager. His father had told him that in the woollen mill days there'd been a girl who'd transferred her affections from her intended to one of the mill operatives; the rejected suitor thereupon consulted a "wise woman". who'd cast a spell for him. A couple of days later the mill-worker was caught in the machinery and was so thoroughly dismembered that (to quote the narrator) "they brought him out in one of the big wicker baskets used in the mill".
Another story, which has achieved print, concerns two ladies from London, on holiday in the area. They $\mathrm{cam}_{3}$ across a cottage which looked so delightful that they called, with intent to try to stay there if possible. The door was opened by a pretty little girl, who invited them into the parlour to await her mother. In the room was a white cat curled up on a chair and a black bird in a cage; both of them seemed to regard the visitors with hostility. When the lady of the house arrived she was apologetic, but no rooms were available at the moment. Perhaps later in the year?
In the Autumn the two ladies returned. The cottage had gone; all that was left was a pile of rubble, overgrown with trees and bushes.

Returning to the firm ground of technology, you might be interested to know that the electrical plant of 1891 provided a public supply at $2 \mathrm{kV}, 100 \mathrm{~Hz}$, with stepdown transformers at each drop-out point. One wonders what problems were encountered with line surges caused by resonance and how they were dealt with. The pioneers of high-voltage a.c. transmission lines were S.Z. de Ferranti and Professor J.A. ("diode") Fleming; in 1890 they were just about the only two people in the country with any theoretical knowledge and practical experience of the matter, so it is at least feasible that either or both might have had some part in the enterprise. Perhaps their ghosts haunt the site also!

## The Dymar 1680 portable frequency counter



## Yoưre working miles from base, there's no power handy, and enough AM about to drive you crazy...

You don't have to have problems in order to appreciate the Dymar 1680 frequency counter. But if they arise you'll know you've got a friend.

At home or away, the 1680 offers a frequency range of from 30 Hz to 600 MHz with exceptionally high sensitivity right across the range.

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In the field, the 1680 really comes into its own. Operating from AC mains supply or its own rechargeable batteries, it provides genuine portability at only 5.5 l bs weight and $7.2 \times 10.6 \mathrm{x}$ r.gin dimensions.

Want to know more? Use the Reader Reply Service or contact Dymar direct.

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| BC 11074 | 0.130 | 0.163 | 2N1132 | 0.240 | 0.300 |
| BC 107 B | 0.140 | 0.175 | 2N 2129 | 0.240 | 0.300 |
| BC 108 | 0.090 | 0.113 | 2N 2218 A | 0.220 | 0.275 |
| BC 108A | 0.130 | 0.163 | 2N 2219 | 0.220 | 0.275 |
| BC 108B | 0.130 | 0.163 | 2N 2219 A | 0.220 | 0.275 |
| BC 108C | 0.140 | 0.175 | 2N 2221 | 0.180 | 0.225 |
| BC 109 | 0.090 | 0.113 | 2N 2221A | 0.210 | 0.263 |
| BC 1098 | 0.140 | 0.175 | 2N 2222 | 0.200 | 0.250 |
| BC 109C | 0.140 | 0.175 | 2N 2222 A | 0.250 | 0.313 |
| BC 184(K) | 0.120 | 0.150 | 2N 2904 | 0.190 | 0.238 |
| BC 212A(K) | 0.110 | 0.138 | 2 N 2905 A | 0.230 | 0.288 |
| BC 212B(K) | 0110 | 0.138 | 2N 2906 | 0.170 | 0.213 |
| BC 213C(K) | 0. 110 | 0.138 | 2N 2906A | 0.170 | 0.213 |
| BC $214 \mathrm{~B}(\mathrm{~K})$ | 0.110 | 0.138 | 2N 2907 | 0.220 | 0.275 |
| BCY 71 | 0.220 | 0.275 | 2N 2907A | 0.240 | 0.300 |
| BFY 50 | 0.200 | 0.250 | 2N 3053 | 0.180 | 0.225 |
| BFY 51 | 0.200 | 0.250 | 2 N 4037 | 0.250 | 0.313 |
| BD 131A | 0.360 | 0.450 | 1N 4001 | 0.050 | 0.054 |
| BD 135 | 0.360 | 0.450 | 1N 4002 | 0.065 | 0.070 |
| BD 136 | 0.396 | 0.495 | 1 N 4003 | 0.070 | 0.076 |
| BD 137 | 0.432 | 0.540 | 1 N 4004 | 0.075 | 0.081 |
| BD 138 | 0.450 | 0.563 | 1 N 4005 | 0.080 | 0.086 |
| BD 139 | 0.495 | 0.619 | 1N 4006 | 0.085 | 0.092 |
| 2N 929 | 0230 | 0.288 | 1N 4007 | 0.090 0.040 | 0.097 0.050 |

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        and mute cicuits 
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Set of transistors. diodes.
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        crrcuits for mounting on pack 
        gined front end module. coil
        filter
5 Fibreglass printed circuit board for
6 Stereo decoder of
        cermet preset for decoder
7 Set of transistors LED. integrated
        circuit for decoder
8 Set of components for channe
        selector switch module including
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        push-button swutches. knobs. LEDs
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        E2.152.60
```

        9 Function switch. 10 turn tuning
        potentiometer. knobs
        10 Frequency meter, meter driomet
        10 Frequency meter, meter drive components.
        fibreglass printed circuit board
        12 Set of capacitors, rectifiers, voltagescrean printed facia panel, acrylicsilk screen printed tuning indicatorpanel insert. internal screen. fixingDarts. etc.
    $$
\begin{aligned}
& \text { struction notes (free with complete } \\
& \text { kit) }
\end{aligned}
$$

16 Teak cabinet
One each of packs 1-16 inclusive are required for complete stereo FM tuner
Total cost of individually purchased packs $£ 7685$

## STEREO FM TUNER KIT

In the April and May issues of Wireless World there was published a novel design for an f.m tuner which combines consistent high performance with the elimanation of the critical setting-up procedure equired by 100 many earlien tuners. This original circuit has been developed further and is used as the basis for our new slimline unit The front end is a ready built pre-aligned module which then feeds an amplifier driven screened three section ceramic filter leading to an integrated circuit five-stage limiting amplifier providing excellent a m. ejection. This is followed by a single coil integrated balanced emodulator from which the sipe col batak. Tenperatur mer ompona en-turn tuning potentiometer or by a choice of six preset push-button controls Each of the presel controls can be adjusted on the fron panel with the settings being indicated by six LED lamps behind an acrylic silk screen printed bacia panel insert. Additional circuitry includes temperature compensated AFC restricted to less than station spacing, inter-station muting. a single-lamp LED tuning indicator and a linear sicale frequency meter The stereo decoder. built on a separate board, is based on a well-proven integrated circuit phase-locked-loop o which has been added active filters to remove sub-carrie harmonies and 'birdes' The power supply, 10 ensure station holding stability. uses an integratec cincuit voltage regulator which is powered via a low-hum field specially designed TOROIDAL TRANSFORMER.

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## MORE KITS ON NEXT PAGE!

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| £2500 | E7 80 | ¢1515 | \$4.85 | \&1035 | 1345 |
| $\underline{10.90}$ | 1725 | 1645 | $\pm 4.50$ | ¢495 | 1325 |
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| $\pm 2320$ | E985 | \& 4.25 | L630 | t945 | 2405 |

## *********************************

## 75W AMPLIFIER KIT

In Hi-Fi News there was published by Mr Linsley-Hood a series of four articles (November 1972-February 1973) and a subsequent follow-up article (April 1974) on a design for an amplifier of exceptional performance which has as its principal feature an ability to supply from a direct coupled fully protected output stage. power in excess of 75 watts whilst maintaining distortion at less than $0.01 \%$ even at very low power levels. The power amplifier is complemented by a pre-amplifier based on a discrete component operational amplifier referred to as the Liniac which is employed in the two most critical points of the system, namely the equalization stage and tone control stage, positions where most conventional designs run out of gain at the extremes of the frequency spectrum Unusual features of the design are the variable transition frequencies of the tone controls and the variable slope of the scratch filter. There is a choice of four inputs. two equalized and two linear, each having independently adjustable signal level. The attractive slimline unit pictured has been made practical by highly compact PCBs and a specially designed Toroidal transformer.

Hi-Fi News Linsley-Hood 75W/Channel Amplifier Mk III Version (modifications as per Hi-Fi News April 1974)


```
Pack Fibreglase printed-circuit board
    for power a mp
2 Sat of resistors. capacitors, pre.se1s
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3 Set of emmiconductors for power
    amp. (now using 80
4 Pa, of 2 drilled, finnied heat amkm
        bralase printed-circult board
b set or pre:amp
Set of low nolse rasistora. capacitors
pre-sets for pre-amp
        of low nolse high gain samicon
        ductors for pre.amp
Set ductors for preamp
Set of potentiometeril
Sat of 4 wush-button switches
Torodel transormar complete
        with magnetic screen/housing primary:
        0.117.234 V.secondarins:
f0.85
co.85
〔1.70
c6.50
C8.50
C0.80
c2.70
{2.70
¢2.40
¢205
$3.70
63.70
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Set of low nolso, high gain
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As can be seen from the photograph the tone control unit is very slim (only ${ }^{1 \frac{1}{2}{ }^{\prime \prime}}$ from
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An essential and critical component in a high-quality speaker system is the crossover unit conventionally comprising of a series of passive networks which unfortunately. though introducing reactive impedances between the amplifier and the speakers, result in the loss of the advantage of high solution to this problem and renders the speakers prone to overshoots and resonances. An elegant solution to this problem. described by D. C. Read in Wireless World. involves the use of a series of active filters splitting the output of the pre-amplifier into three channels, of closely defined bandwidth. each of which is fed to the appropriate speaker by its own power amplifier. A design for a suitable 20 -watt amplifier, based on a proven Texas circuit, was also described by Mr Read. The printed-circuit board for this has been designed such that three amplifiers may be stacked and mounted together on a common heat sink to achieve a conveniently compact module

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| 104 | 20 | 4.95 | B |
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| 106 | 4.0 | 7.98 | 1 |
| 107 | 6.0 | 12.71 | 1 |
| 118 | 8.0 | 13.63 | 1 |
| 119 | 10.0 | 17.75 | B |


| \& P | Rof. No. | SECO | $\begin{aligned} & \text { DARY } T \\ & 20-25 \end{aligned}$ | S |
| :---: | :---: | :---: | :---: | :---: |
|  |  | Amps | f | $\mathrm{Pg}_{\mathbf{p}} \mathrm{P}$ |
| P | 112 | 0.5 | ¢.90 | 58 |
| 5 | 79 | 1.0 | 2.52 | 72 |
| 12 | 3 | 20 | 3.77 | 72 |
| 25 | 20 | 3.0 | 4.70 | 85 |
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| 26B | $8 \frac{3}{4 \prime}$ | $5 \frac{3}{4}^{\prime \prime}$ | $81^{\prime \prime}$ | 5.62 | 6.10 | 0.95 |
| 27 A | 1214" | $7 \frac{1}{2}^{\prime \prime}$ | $5 \frac{1}{2}^{\prime \prime}$ | 5.75 | 6.35 | 0.95 |
| 27B | 1214* | $7 \frac{1}{2}^{\prime \prime}$ | 8' | 6.35 | 6.95 | 0.95 |
| 28A | $14^{\prime \prime}$ | $10 \frac{1}{2}^{\prime \prime}$ | $6 \frac{1}{2}^{\prime \prime}$ | 6.95 | 7.55 | - |
| 28B | $14^{\prime \prime}$ | $10 \frac{1}{2}{ }^{\prime \prime}$ | $8 \frac{1}{2}^{\prime \prime}$ | 7.55 | 8.15 | - |
| 29A | $10^{\prime \prime}$ | $4^{\prime \prime}$ | 6" | 4.85 | 5.33 | 0.95 |
| $29 B$ | $10^{\prime \prime}$ | $4^{\prime \prime}$ | $8^{\prime \prime}$ | 5.15 | 5.63 | 0.95 |
| 30A | 12" | 5" | $6^{\prime \prime}$ | 5.25 | 5.85 | 0.95 |
| 30B | 12" | 5" | 8" | 5.56 | 6.16 | 0.95 |
| 31 A | $14^{\prime \prime}$ | 6" | $6^{\prime \prime}$ | 5.75 | 6.35 | 0.95 |
| 31 B | 14" | 6" | $8^{\prime \prime}$ | 6.05 | 6.65 | 0.95 |
| 61 | $15 \frac{1}{2}{ }^{\prime \prime}$ | $7 \frac{1}{2}^{\prime \prime}$ | 919 $\frac{1}{2}^{\prime \prime}$ | - | 8.75 | - |
| 62 | $17 \frac{1^{\prime \prime}}{}{ }^{\prime \prime}$ | $8 \frac{1}{2}{ }^{\prime \prime}$ | 9191 | - | 10.15 | - |
| 63 | $16 \frac{1}{2}{ }^{\prime \prime}$ | $9 \frac{1}{2}^{\prime \prime}$ | 919 ${ }^{\prime \prime}$ | - | 10.15 | - |
| 64 | $15 \frac{1}{2}{ }^{\prime \prime}$ | $7 \frac{1}{2}^{\prime \prime}$ | 12192 | - | 10.15 | - |
| 65 | $17 \frac{1}{2}{ }^{\prime \prime}$ | $8 \frac{1}{2}^{\prime \prime}$ | 1219 ${ }^{\prime \prime}$ | - | 11.60 | - |
| 66 | $16 \frac{1}{2}{ }^{\prime \prime}$ | $9 \frac{1}{2}^{\prime \prime}$ | 121 $\frac{1}{\prime \prime}^{\prime \prime}$ | - | 11.60 | - |

Types 21, 22, 23 and 24 are finished in olive green hammertone with front panels in light straw gloss enamel. Fitted with ventilated rear panels only. No louvres in the base.

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It mears ultra linear; UL also represents inlimited quality of sound: Celestion experience nakes possible a prestige product at_an ordinary cost.
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## S-2020TA STEREO TUNER/AMPLIFIER KIT

## NEW PRODUCT

A high-quality push-button FM Varicap Stereo Tuner combined with a 20 W r.m.s.


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## per channel Stereo Amplifier.

Brief Spec. Amplifier: Low field Toroidal transformer, Mag. input. Tape In/Out facility (for noise reduction unit. etc). THD less than $0.1 \%$ at 20 W into 8 ohms. All sockets, fuses, etc, are PC mounted for ease of assembly. Tuner section: uses Mullard LP1186 module requiring no RF alignment, ceramic IF. INTERSTATION MUTE, and phase-locked IC stereo decoder. LED tuning and stereo indicators. Tuning range $88-104 \mathrm{MHz} .30 \mathrm{~dB}$ mono $\mathrm{S} / \mathrm{N} @ 1.8 \mu \mathrm{~V}$.THD typ. $0.4 \%$.

PRICE: $£ 47.95+99 p$ p $\&$ + VAT.


## NELSON-JONES STEREO FM TUNER

A very high performance tuner with dual gate MOSFET RF and Mixer front end, triple gang varicap tuning, and dual ceramic filter/dual IC IF amp.

Brief Spec. Tuning range $88-104 \mathrm{MHz}$. 20dB monv quieting @ $0.75 \mu \mathrm{~V}$. Image rejection-70dB. IF rejection- 85 dB . THD typically $0.4 \%$.
IC stabilized PSU and LED tuning indicators. Push-button tuning and AFC unit. Choice of either mono or stereo with a choice of stereo decoders.

> PRICE: Mono $£ 25.46+85$ p p\&p+VAT;
> With Portus-Haywood Decoder $£ 31.96+85 p$ p\&p+VAT;
> With ICPL Decoder $£ 29.73+85$ p p\&p+VAT.

## NEW PRODUCT

## S-2020A AMPLIFIER KIT

Developed in our laboratories from the highly successful "TEXAN"
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Typ. Spec. $20+20 \mathrm{~W}$. r.m.s. into 8 -ohm load at less than $0.1 \%$ THD. Mag. PU input $\mathrm{S} / \mathrm{N} 60 \mathrm{~dB}$. Radio input $\mathrm{S} / \mathrm{N} 72 \mathrm{~dB}$. Headphone output. Tape In/Out facility (for noise reduction unit, etc). Toroidal mains transformer.

PRICE: £29.95 + 99p p\&p+VAT.


## STEREO MODULE TUNER

A low-cost Stereo Tuner based on the Mullard LP1186 RF module requiring no alignment. The IF comprises a ceramic filter and highperformance IC. Variable INTERSTATION MUTE. PLL stereo decoder IC.
Typ. Spec. Sens. $30 \mathrm{~dB} \mathrm{~S} / \mathrm{N}$ mono @ $1.8 \mu \mathrm{~V}$. Tuning range $88-104 \mathrm{MHz}$. LED sig. strength indicator. LED Stereo indicator. THD typically $0.4 \%$.

PRICE: Stereo $£ 27.80+85$ p p\&p + VAT. Mono $£ 24.70+85$ p p\&p + VAT
ALL THE ABOVE KITS ARE SUPPLIED COMPLETE WITH ALL METALWORK, SOCKETS, FUSES, NUTS AND BOLTS, KNOBS, FRONT PANELS, SOLID MAHOGANY CABINETS AND COMPREHENSIVE INSTRUCTIONS.

## SUB ASSEMBLIES

## BASIC NELSON-JONES TUNER

Supplied as a printed circuit board with all components and screening box to build a varicap tuner module. Performance spec as above for complete N -J Tuner. For suitable stereo decoders see below. (Illustrated without screening box.)

PRICE: $£ 12.88+25 p$ p $\&$ p VAT.

## BASIC MODULE TUNER

Supplied as a printed circuit board with all components and screened Mullard LP1186, to build a mono or stereo tuner module. Performance spec as above for Stereo Module Tuner complete kit.

PRICE: Mono £13.00 + 25p p\&p + VAT: Stereo $£ 15.00+25 p p \& p+$ VAT
PORTUS-HAYWOOD PHASE-LOCKED STEREO DECODER
Mk II version of this design (WW Sept. 1970). The lowest distortion phase-locked stereo decoder kit available (Typ. $0.05 \%$ @ N-J Tuner O/P level). Separation 40 dB up to 15 KHz .
Complete kit comprises PCB and all components, inc. stereo LED.
PRICE: $£ 7.68+25 p$ p\&p+VAT.

## PHASE-LOCKED IC DECODER KIT

Integrated circuit phase-locked stereo decoder based on the MC1310. THD typically $0.3 \%$. Separation $40 \mathrm{~dB} @ 1 \mathrm{KHz}$.
PRICE: $\{4.27+20 p$ p\&p+VAT.
PUSH-BUTTON UNIT
The six-position push-button unit used in our tuners and tuner/amp. Each track has the required diode law for stability of tuning. There are approx. 40 turns on each button and there are six separate moving pointers. An AFC disable switch is incorporated with each button. The unit is finished in black with red pointers.

PRICE: $£ 3.30+20 p \mathrm{p} \& \mathrm{p}+$ VAT


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NOISE FIGURE METER TYPE II3A (Magnetic AB, Sweden): Complete with Noise Source 121 and 122. £I25. Carr. E1.
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| 2， 3442 | 140 | ${ }^{\text {AC }} 152$ | 049. | 8С309С | 0.20 | WJ490 | 1.05 | BY126 | 0.12 |
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DEPARTMENT OF HISTORY AND COMMUNICATION STUDIES

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