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Wireless World

ELECTRONICS, RADIO, TELEVISION

NOVEMBER 1964

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MULLARD 'PANORAMA'

SPECIAL BENEFITS OF 'PANORAMA'

The benefits to be gained from the absence of any protective panel in a television receiver, when a 'Panorama' tube is fitted, are fourfold:

1. Problems of dust accumulation between the faceplate and the panel are eliminated, and the difficulty of maintaining an efficient dust seal between faceplate and panel is removed. There can be no inaccessible dust to spoil the brightness of the picture.

2. The number of reflecting surfaces between the viewer and the picture is minimised. Reflections of internal and external light are thus reduced and picture-contrast accordingly improved.

3. Greater freedom is allowed for cabinet styling since, in conventional receivers, the style is dictated to a large extent by the need to provide the safety screen.

4. Smaller and lighter receivers with better physical stability are possible. Absence of the panel eliminates the space between the faceplate and the screen, and gives a better distribution of weight. (Use of the 'short' electron gun also contributes to shallow cabinets.)

Mullard

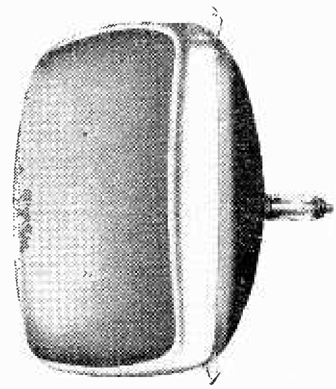
What's new in the new sets

DIRECT VISION PICTURE TUBES

'Panorama' direct vision picture tubes represent a new concept in tube design and have been introduced by Mullard to complement their existing range of 'Radiant Screen' tubes.

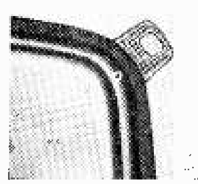
From the earliest days of popular television, research and development by Mullard on picture tubes have produced a steady progression of advancements in design and performance. Tubes with small round screens and narrow deflection angles gave way to larger screens with straighter edges and steadily increasing deflection angles. New screen coatings giving brighter pictures with better contrast were introduced in the 'Radiant Screen' tubes. Improved electron guns gave crisper definition and a more life-like picture quality.

Further improvements in screen shape resulted in the present-day 19 and 23-inch tubes. Use of 110-degree deflection and the 'short' unipotential electron gun enabled these larger tubes to be produced without recourse to a corresponding increase in depth of the tube. All these features have led steadily to better performance and improved receiver design and styling. All are incorporated in the Mullard AW47-91 and AW59-91 'Radiant Screen' tubes and in the 'Panorama' tubes.



Two 'Panorama' tubes, the 19-inch A47-18W and the 23-inch A59-11W, are being introduced. They are the latest outcome of continuous research by Mullard and represent a revolutionary advance in picture tube development. Reinforcement of the picture tube envelope by means of a specially designed metal band fitted around the periphery of the faceplate, has eliminated the need for a separate protective panel in front of the tube and made possible 'direct vision' viewing.

SERVICING BENEFITS



Apart from the time saving due to elimination of any dust problem, service engineers will welcome the unprecedented ease of mounting 'Panorama' tubes by means of the four fixing lugs on the metal band. With 'Radiant Screen' and 'Panorama' Mullard now offers the industry a choice of the two most advanced picture tube techniques.

MVE 5486

14WW—101 FOR FURTHER DETAILS.

Wireless World

VOL 70 NO 11 NOVEMBER 1964

Why Lasers?

THE popular press, and now the enemies of James Bond in the make believe world of the cinema, have invested the laser with an aura of fantasy which is sometimes difficult to dispel. This is partly because many of us have forgotten our classical optics and cannot explain at a moment's notice how light beams are formed and why the spectral purity of laser light should permit concentrations of peak power which are several orders of magnitude above our previous experience. But there is nothing in laser light which invalidates the principle of Huygens or the laws of diffraction.

"Cathode Ray" on p. 563 of this issue explains in his inimitable style the differences between coherent and incoherent light, and those who were able to attend the conference on lasers and their applications, sponsored by the I.E.E., the I.E.E.E. and the I.E.R.E. at Savoy Place last month will also have found their feet on firmer ground after hearing the opening address by Sir Robert Cockburn and listening to the discussions.

Although the scientific interest of lasers is enormous the practical uses, to quote Sir Robert, "... have been somewhat limited (I almost said trivial) ..." One of the reasons is the low continuous or mean pulsed powers available (1 to 100 watts) and the low efficiencies (1% in gaseous and solid lasers, though 30% is obtained, with lower directivity, in gallium arsenide junction lasers). Point-to-point communications through the atmosphere, although offering by present standards almost unlimited channel space, are reliable only to distances of the order of a mile. Machining and welding operations are possible—with the aid of a microscope—and the potentialities of improved radar performance by exploiting the higher frequencies to increase scanning rate without sacrifice of resolution can at present be realized only at short ranges, for lack of power. In this connection Sir Robert pointed out that an input of 10 kW was at present about the minimum requirement for a relatively modest coverage, and that with 1% efficiency the problem of getting rid of the heat was quite considerable.

The world of lasers is Lilliputian or Brobdingnagian according to the way you do your arithmetic. If by Q-switching (building up oscillations at low level in a lossy cavity and then suddenly removing the losses) you can discharge only 1 joule in 10^{-8} seconds, you can write to the papers that you have a 100-megawatt laser and no one can disprove you. With this powerful tool you can do in a day as much damage to a razor blade as a month of ordinary shaving would do, but the damage will be localized to a degree limited only by the quality of the lenses used to concentrate the beam. Concentrations of 10^{11} W/cm^2 have been reported (paper by T. P. Hughes, N.P.L.) in a spot area of 0.1 mm diameter giving, by time-resolved spectroscopy, a surface temperature of 5,000°K with a trapped underlying plasma temperature of at least 20,000°K. Light fluxes up to 10^{14}W/cm^2 are feasible in an area of a square wavelength and this is equivalent to an electric field strength of 300 MV/cm which is comparable with internal atomic field strengths.

So here we have part of the answer to the question we have posed ourselves. The laser is a powerful tool for scientific research. It is already in regular use in laboratories concerned with plasma physics and atomic energy. It has surgical precision and has already been used successfully to coagulate tissue and repair detached retinas in the eye. But it has not yet fulfilled its promise as a means of communication and will not do so until the economics of providing the environment of outer space, or otherwise removing the inimical effects of atmospheric turbulence are shown to be practicable.

Constant Luminance

ITS MEANING, PURPOSE AND PRACTICAL APPLICATION IN COLOUR RECEIVERS

By IAN MACWHIRTER,* A.M.I.E.E.

THE first colour television system which was tried and evaluated in an attempt to convey high-quality tri-chromatic information within the confines of an existing black and white television radio frequency channel, was the R.C.A. dot-sequential system.[1] This system of signal coding, developed during the late 1940's, underwent a series of subtle mathematical changes by the contributor members of the United States National Television System Committee (N.T.S.C.), before it was finally adopted by the Federal Communications Commission, (F.C.C.), as the system by which public colour television broadcasting could begin in the U.S.A. One of the major improvements to the early R.C.A. dot-sequential system, the use of constant luminance sampling, was made following a breakthrough by Loughlin[2] as a result of an analysis of the nature of the R.C.A. dot-sequential signal coupled with the known characteristics of colour vision.

The object of this article is to explain the concept of constant luminance and how the advantages to be gained from its use are obtained. In doing so, it is hoped to show that the transmission of a true luminance signal does not, of itself, necessarily permit complete constant luminance operation. Some details of the "separate luminance" principle are also given to distinguish it from constant luminance.

Band-Sharing

Three of the essential requirements for a broadcast colour television system are:—

(i) The colour signal shall provide a high-quality picture when seen on an unmodified black and white receiver.

(ii) The complete signal shall occupy no greater bandwidth than the existing black and white signal.

(iii) The colour system shall convey sufficient information for a pleasing tri-chromatic reproduction to be obtained on a correctly designed receiver.

The first requirement is satisfied by the transmission of a signal representative of the luminance of objects. The luminance signal E_Y may be derived from a single pick-up tube in a camera whose spectral response curve matches the photoptic visibility curve of the eye (Fig. 1). Alternatively, it may be formed by matrixing from suitable tri-stimulus voltages which are generated by three separate pick-up tubes within a camera as explained later.

The luminance response curve is, in any case, required for normal black and white transmissions (in practice, however, for black and white transmissions, a modified curve is commonly used which is claimed to give a superior rendition of tone values than one might predict from known theory alone). In other words, one component chosen for the colour signal, the luminance signal, could be expected to be indistinguishable from a normal black and white signal.

Having satisfied the first requirement, both in response to the colour spectrum and television signal bandwidth, the second can be satisfied only if band sharing can be used. That is to say, the additional colour information must be transmitted within the same bandwidth as the luminance signal. In colour television systems which have been proposed for public broadcasting, e.g. N.T.S.C. and SECAM, the colour information also carries luminance and the signals are in the form (E_R-E_Y) and (E_B-E_Y), where E_R and E_B are voltages representing two of the tri-chromatic variables E_R (red) and E_B (blue) and E_Y the voltage representative of luminance. The third colour component E_G is discussed later.

The bracketed components, called chrominance signals, do carry luminance since E_Y is written into each one. At the receiver they serve, on the red, green and blue phosphor stimuli, to add to or subtract from the luminance values which would otherwise be established by the signal E_Y. For example, the total red stimulus is controlled by the voltages $(E_R-E_Y)+E_Y=E_R$.

Finally, it is known that when attempting to match colours in a three-colour colorimeter, for very small angles of view it is sufficient to make a match of luminance alone, i.e. the eye appears to become increasingly less conscious of hue, and hue errors, as the field becomes narrower. Using this knowledge, it has been found possible to restrict the bandwidth of these chrominance signals to about one-fifth of the accompanying luminance signal without objectionable loss of sharpness in the colour

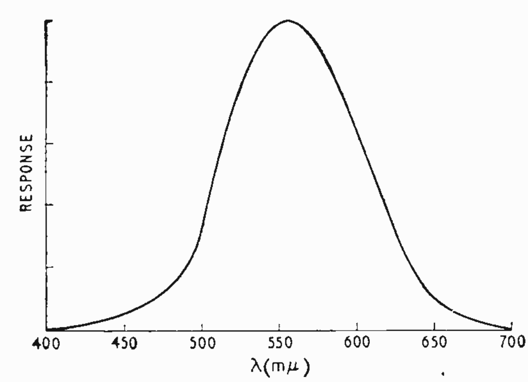


Fig. 1. Photoptic visibility curve of the human eye.

* Thorn-A.E.I. Radio Valves & Tubes Ltd.

picture or any apparent lack of colour in areas of fine detail.

The band-shared spectrum of a colour television signal is represented diagrammatically in Fig. 2.

In addition to being occupied by the luminance signal E_Y, the signal spectrum from d.c. to f_{max} is also occupied by the chrominance components which themselves modulate a sub-carrier F.

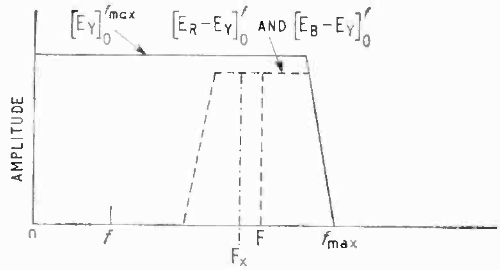


Fig. 2. The spectrum of a band-shared colour television signal.

Composition of the Luminance Signal

When the luminance signal is to be formed by matrixing, the colour camera is required to have pick-up devices which produce tri-stimulus voltages E_R, E_G, and E_B. The spectral response of these devices may be calculated using a knowledge of both the particular phosphors to be used on the display and the particular white point to which the display should be matched under the conditions $E_R=E_G=E_B=1.0$. The luminance signal E_Y may be derived by adding suitable fractions of these tri-stimulus voltages thus:

$$E_Y=lE_R+mE_G+nE_B$$

The coefficients l, m and n, which are, in turn, appropriate to the particular tri-stimulus colour responses, are a direct indication of the relative luminosities (brightness) of the reproducing phosphors. The signal E_Y is, therefore, intentionally proportioned so that it has, in effect, a colour response which matches that of the eye at normal or photoptic brightness levels.

It is important to grasp, therefore, that this photoptic curve is matched equally well by the classical N.T.S.C./SECAM values where

$$E_Y=0.2989\ E_{R1}+0.5865\ E_{G1}+0.1146\ E_{B1}$$

and by a luminance equation recently derived[5] for a colour camera designed to work with the present-day sulphide phosphors and a 9300°K white point when

$$E_Y=0.1903\ E_{R2}+0.7228\ E_{G2}+0.0869\ E_{B2}$$

Normally the second suffixes 1 or 2, which relate to the phosphors and white point assumed, are omitted in technical writings and this can cause confusion as to the reason why there is not a unique set of coefficients to describe the matrixed luminance signal.

If signals of this form are used with display tubes of transfer characteristics (gamma index)γ, then the pre-corrected luminance signal should be

$$E_Y^{\ 1/\gamma}=(lE_R+mE_G+nE_B)^{\ 1/\gamma}.$$

While a signal of this form may be made by matrixing as already discussed, the elegant method is to derive the luminance signal directly from a single pick-up tube. This procedure can be shown to have advantages in the design and maintenance of camera equipment.

Reclaiming the Green Component

Since the chrominance signals chosen for transmission along with the luminance signal are commonly $[E_R-E_Y]_o^f$ and $[E_B-E_Y]_o^f$, it is necessary to re-create $(E_G-E_Y]_o^f$ at the receiver. The square brackets here indicate that these signals are band-width restricted, within the limits of d.c. to a frequency f, compared with the accompanying luminance signal. When the colour difference signals modulate the sub-carrier F, these frequency limits are transposed, e.g., the upper sideband occupies the spectrum F to $(F+f)$.

Now the expression for the luminance signal $E_Y=lE_R+mE_G+nE_B$ may be re-arranged in the form

$$(E_G-E_Y)=-\frac{l}{m}(E_R-E_Y)-\frac{n}{m}(E_B-E_Y)$$

and this equation for deriving the green colour difference component from the two transmitted difference signals will prove to be a keystone to the successful implementation of the constant luminance principle.

At the receiver, as was indicated earlier, the original tri-stimulus voltages may be obtained as follows: for the case of red and for such low frequencies below f where the bandwidth restriction of the red colour difference channel subtracts no energy, the square brackets may be omitted.

$$\therefore\quad E_R=(E_R-E_Y)+E_Y=E_R$$

A more general equation is:—

$$E_R=[E_R-E_Y]_o^f+[E_Y]_f^{max}=[E_R]_o^f+[E_Y]_f^{max}$$

This shows that up to the chrominance cut-off frequency f, the true red signal is available. For higher frequencies corresponding to areas of picture detail between f and f_{max}, control of the display tube is by the luminance component only. This latter is applied, in the case of a three-gun shadow-mask tube, simultaneously to all guns so that fine detail is reproduced as a "black and white", or achromatic modulation upon the background colour. This is in accordance with the observed minimum requirements of the human eye as explained earlier. In the early days of the work of the N.T.S.C., the signal $[E_Y]_f^{max}$ was called the "mixed highs" signal but this term has fallen into disuse.

The Purpose of Constant Luminance

Referring to Fig. 2, it does appear inevitable that as a result of band sharing some unwanted cross-talk between luminance and chrominance components must occur. It is no encouragement to colour television engineers that the presence of the chrominance components causes disinterested viewers [3] of the compatible black and white picture to give typical quality labels of slightly poorer than "just perceptible" for the N.T.S.C. system and "definitely perceptible but not disturbing" for the SECAM system to the impairment of the picture caused by the presence of these chrominance signals.

Conversely, the parts of a colour receiver which demodulate the chrominance signals must also receive that part of the luminance signal which is band-shared. The resulting unwanted cross coupling is known as cross colour and produces spurious visible effects on the picture. Similarly, any noise

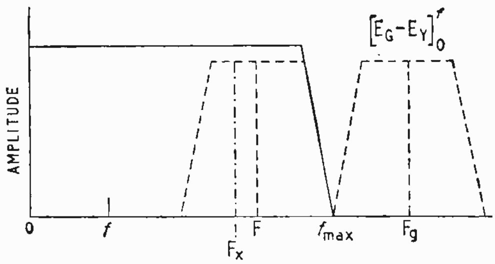


Fig. 3. Spectrum as in Fig. 2 with additional out-of-band $[E_G-E_Y]$ signal.

or interference which may be present in the band-shared spectrum will also produce unwanted spurious effect on the picture, colloquially known as 'parc'. The subjective effects of parc and cross colour are especially noticeable owing to the fact that in the process of demodulation of the chrominance signals, their frequency components are heterodyned down. Whereas band-shared components illustrated in Fig. 2, produce relatively fine-grain low-visibility effects when displayed with the luminance signal in a black and white receiver, following chrominance demodulation in a colour receiver a signal F_x in the vicinity of the subcarrier F is reproduced as $(F - F_x)$. This frequency shifting is desirable for the chrominance components but highly undesirable for the band-shared components of E_Y and noise because low-frequency signals of a given amplitude are, in general, more annoying than high-frequency signals of the same amplitude. By this heterodyning process, the susceptibility to noise and other spurious signals of a band-shared colour television system is inevitably worse than a black and white system of the same nominal channel width f_{max}.

One of the functions of constant luminance operation is to minimise the subjective annoyance of noise in the band-shared spectrum. By constraining the visible effects of noise to perturbations in hue, another feature of colour vision is involved. This is that the eye is less sensitive to changes involving hue alone than is the case when changes in luminance also occur. Loughlin's work[2] showed that provided the relative gains of the three colour difference channels are inversely related to the luminosity coefficients (relative luminosities) of the display's phosphors, then this noise protection feature can be achieved. At the same time, the transmission of all and nothing but luminance information by the luminance channel means that the bandwidth restriction of the chrominance information in no way detracts from the ability of the luminance signal to portray fine colour detail in black and white. These are two of the advantages of constant luminance working. The name "constant luminance" derives from the fact that, provided all phosphor stimuli are operating, the luminance of the display remains constant despite any change of signal level in the chrominance channel of a receiver. This is because the sum of the luminances caused by components in the band-shared channel is designed to be zero.

The old R.C.A. symmetrical dot-sequential signal[1] was called "constant amplitude" to indicate that although band-shared components produced equal *voltages* in the colour difference channels, which added vectorially to zero, the luminance effects did not.

A qualitative account of the mechanism of constant luminance can be given as follows:—

Excitation of each of the three phosphors produces luminance. The observed luminance will be the numerical sum $Y_R + Y_G + Y_B$ (the red, green and blue luminances). Hence, if a signal in the chrominance channel causes, say, an increase in the value of $[E_R - E_Y]'_o$ and $[E_B - E_Y]'_o$ with corresponding increase in red and blue luminance, then since

$$Y_G \equiv E_Y - \frac{l}{m}[E_R - E_Y]'_o - \frac{n}{m}[E_B - E_Y]'_o,$$

it is clear that the green luminance Y_G will decrease. If the proportions are correctly chosen, this decrease in Y_G can be made to compensate exactly for the increases in Y_R and Y_B.

The mathematical relationships which govern the achievement of constant luminance are well known[4] and can also be expressed as follows for an N.T.S.C. type of transmission:

$$l^2 a^2 + n^2 c^2 = m b^2 \quad \text{(i)}$$

where a, b and c are the gains associated with red, green and blue channels of a colour receiver following unity gain demodulators; and before phosphors of (assumed) balanced efficiencies.

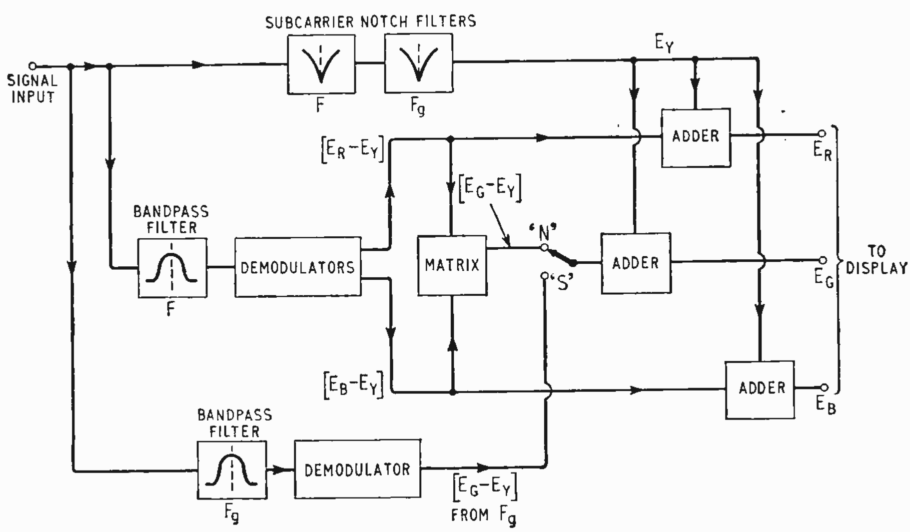


Fig. 4. A colour receiver switchable for 'normal' recovery of $[E_G-E_Y]$ or 'special' recovery.

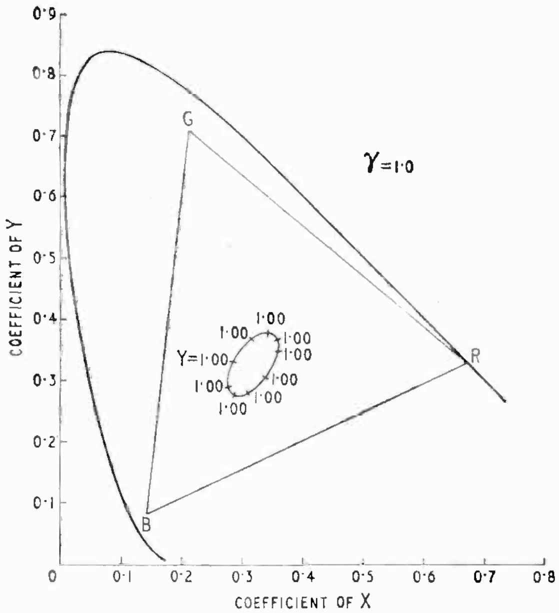


Fig. 5. Effect of a band-shared signal 0.2 sin ($F_x + \psi$) when the intended colour is white, $E_Y = 1.0$.

The Mechanism of Constant Luminance

The application of the theory can be illustrated numerically by calculating one aspect of the noise susceptibility of a receiver designed to receive the colour signal shown in Fig. 3. This will be seen to be similar to the spectrum of Fig. 2, except that the green colour difference component is independently made available on another sub-carrier F_y outside the spectrum of what would be a normal N.T.S.C. or SECAM signal. A receiver (Fig. 4) could be made switchable to be either "NORMAL," i.e. the $[E_G - E_Y]'_o$ component is derived from the other two, or to have a separate out-of-band $[E_G - E_Y]'_o$ detector ("SPECIAL" case).

In the event of a spurious signal of frequency F_x being present in the band-shared spectrum as shown in Fig. 3, and with the green channel selected as for "NORMAL" reception, the visible perturbation caused by F_x can be calculated for an N.T.S.C. type receiver and is shown in Fig. 5. It will be seen that although a cyclical change of hue is observed, depending on the phase of F_x, there is no change in the reproduced luminance. However, if the green channel switch is set to "SPECIAL," i.e. the green phosphor is not excited by signals representative of $(F - F_x)$, then not only do hue variations occur but variations in luminance as well which result in greater subjective annoyance, see Fig. 6.

For this case, although a true luminance signal E_Y is still radiated, there has been a failure of the particular noise protection feature which constant luminance should provide. In other words, the transmission of a true luminance signal does not, of itself, guarantee correct constant luminance receiver operation.

It should be clear from the foregoing that the achievement of constant luminance is essentially a receiver design problem. Specifically, it requires that one of the colour difference signals be derived from the other two, and that these two share the same band. The gains in each of the three colour difference channels should then be precisely related to the luminosity coefficients of the display's phosphors according to equation (i). The transmitted signal is then proportioned to match the requirements of the receiver.

Effects of "Gamma" Pre-correction

Display tubes used for colour television have a transfer characteristic or gamma index of about 2.8. It is necessary, therefore, to apply an inverse pre-correction in the picture originating equipment to correct for this non-linearity.

Both present-practice N.T.S.C. and SECAM signals are gamma pre-corrected in the following way: $E_Y' = lE_R^{1/\gamma} + mE_G^{1/\gamma} + nE_B^{1/\gamma}$ and the two transmitted chrominance components are: $[E_R^{1/\gamma} - E_Y']'_o$ and $[E_B^{1/\gamma} - E_Y']'_o$. (Note that this form of luminance signal can only be formed by matrixing.)

It has been shown [5] that the mathematical analysis which establishes the practical conditions for the achievement of constant luminance receiver operation is similar to, but not identical with, that of the linear case.

If this solution (equation (i)) is used in the receiver design then it follows, by definition, that the luminance signal should provide all the luminance information—and nothing more—for the correct reproduction of luminance and hue in both large and small areas of the picture. The reproduced luminance Y from the luminance signal is obviously proportional to $(E_Y')^\gamma$. Ideally, $(E_Y')^\gamma = E_Y$. This it does for the grey scale when $E_R = E_G = E_B$, but for colours the value of E_Y' is always lower than the ideal luminance signal $E_Y^{1/\gamma}$ where $E_Y^{1/\gamma} = (lE_R + mE_G + nE_B)^{1/\gamma}$. For this latter, the luminance $Y \propto (E_Y^{1/\gamma})^\gamma = E_Y$ which is correct for all colours. The errors in hue and luminance occasioned by the use of present-

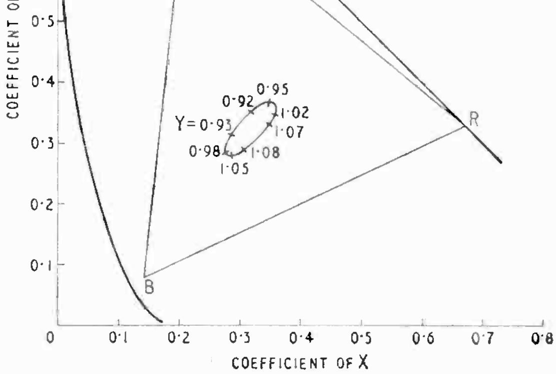


Fig. 6. Effect of a band-shared signal 0·2 sin ($F_x + \psi$) upon the intended colour, system white, but when the green signal is derived from an out-of-band signal (see Fig. 3.)

practice gamma pre-correction are small for pastel colours but rapidly worsen as saturation increases.

For large areas, the errors in luminance are limited to those caused by cross colour and parc only, for small areas which correspond to colour difference frequencies in excess of f (Fig. 2) there are luminance errors in the reproduction of picture detail as well. Unwanted errors in hue in large areas of colour are limited to those caused by cross-colour and parc. The undesired errors are aggravated by the use of display phosphors and white points which are not appropriate to the weighting coefficients assumed in the system. This condition is typical in the operation of colour television receivers today[5].

What can be said at this stage is that it does not appear possible to design an *exact* constant luminance receiver for use with the present practice N.T.S.C. or SECAM signals which gives correct colorimetry. Exact constant luminance receivers require not only more complex circuits than those used at present, but transmitted signals which are different from present practice[6]. Specifically, the complexity arises because of the need for non-linear processing of the signals. This feature is not called for when using present practice signals and in consequence the design of receivers is less costly. Whether the cost of exact constance luminance receivers outweighs the improvement in performance appears to be the subject of some controversy.

Camera Design

The particular formation of the present practice luminance signal E_Y' is the cause of another source of error. When a colour camera using separately scanned pick-up tubes is used to provide the tristimulus voltages E_R, E_G and E_B, then the amplitude of the luminance signal E_Y' is not only low for all coloured objects, as has been explained, but the maximum definition provided by E_Y' is critically dependent upon the accuracy of image registration between the three pick-up tubes. Of course, cameras can be made to provide good registration provided that particular tight tolerances are maintained. However, the problem of design and maintenance of a colour camera would be eased if one of the pick-up tubes generated directly a luminance signal. Proposals for such cameras have been made for some time[6,7] and James has suggested the use of the name "separate luminance" to describe them. "Separate luminance" can be extended to mean a system in which the terminal equipment is particularly designed so that (i) the receiver is "constant luminance" (full noise protection and no low definition chrominance information causing changes of luminance), and (ii) the camera has a "separate luminance" pick-up tube.

Observations

The problem, to which a solution of universal acceptance has not yet been found, is the composing of a specification for a correct constant luminance receiver of no worse stability and with only little more complexity than present practice receivers. Such a receiver would, by definition, provide the viewer with the advantages of both receiving signals from a camera with a "separate luminance" pick-up and having exact adherence to the constant luminance principle.

Such specifications for both the signals and receivers (including the system now attributed to Livingston) were proposed by the various industrial contributors to the N.T.S.C. deliberations during early 1953[8], but at that time the true luminance signal $E_Y^{1/\gamma}$ was derived from matrixing tristimulus voltages.

The subjective benefits of a "separate luminance" pick-up could not, therefore, have been realized. Although the studies of the N.T.S.C. Gamma Sub-Committee were never completed, the gamma specification of the final N.T.S.C. signal was worded so as to allow for the introduction of more advanced forms of gamma correction if their use was desirable.

If a colour receiver designed to work with a present practice N.T.S.C. or SECAM signal is used with a transmission containing a true luminance signal $E_Y^{1/\gamma}$ and either (a) James's colour difference components

$$[E_R^{1/\gamma} - E_Y^{1/\gamma}],\ [E_B^{1/\gamma} - E_Y^{1/\gamma}]$$

or

(b) Livingston's (N.T.S.C. Gamma Sub-Committee case II) components where the colour difference signals are identical with present practice, then hue and luminance errors will occur[6,9]. This results because the value of $E_Y^{1/\gamma}$ is always greater than E_Y' (except for the grey scale) and this will tend to desaturate all colours and reproduce them at higher than intended luminances. There is an exception to this in the case of predominantly green colours with reception from the James signal, when such colours will tend to have higher than intended saturations and luminances.

The introduction of the "separate luminance" camera has reopened this decade-old problem, which is complicated by the use of display phosphors and white points other than those written into the specification for the transmitted signal which are known to aggravate these distortions.

Acknowledgement

The author wishes to thank the management of Thorn-AEI Radio Valves & Tubes for permission to publish this article.

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An Introduction to

MICROWAVE TECHNIQUES

4.—MICROWAVE COMPONENTS AND INSTRUMENTS

By K. E. HANCOCK*

HAVING dealt with microwave power sources in the previous article in this series, we will now pass on to the discussion of various microwave components. There are unfortunately a greater variety of components in use than can be dealt with here. In this article, therefore, we will concentrate on components constructed of waveguide, covering those in common use, and in particular those used in the measurement circuits to be described in the next and final article in this series.

It should be remembered, however, that most of the components described have their counterparts in coaxial line. Many are also made in stripline, a form of construction lending itself to the compact construction of complicated integrated microwave circuits.

Choke Flanges

The first component we will examine is really not a component at all, but a resonant circuit incorporated in many components for reasons shown later.

It is obviously impracticable to solder waveguide

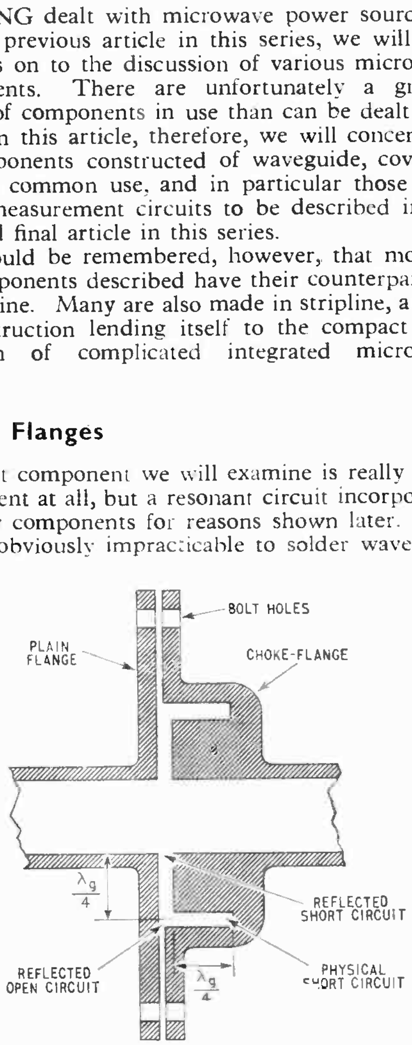

Fig. 1. Cross-section through waveguide coupling showing choke flange.

components together as we would resistors and capacitors. Instead, flat plates called flanges are attached to either end of the waveguide component. Holes in the flanges allow it to be bolted to succeeding components.

Remembering our theory of reflections and standing waves it will be obvious that to prevent reflections due to mis-alignment of the waveguides, the flanges must be exactly at right angles to the centre line of the guide, the bolt holes must be very accurately positioned and finally the complete flange face must be perfectly flat.

These conditions can be fairly easily fulfilled, at any rate to an acceptable limit for low power, but the tight tolerances make it an expensive process. In addition, at high power there is a tendency for arcing to take place over the very small gaps at the flange junctions, particularly if standing waves are present.

To overcome these problems a choke flange is used. This device, which is illustrated in Fig. 1, is essentially a $\lambda_g/2$ short-circuited line built into the flange. It will be remembered from the previous article that a half-wavelength line will reflect its terminating impedance to the input. As the choke is terminated in a short circuit, an effective short circuit will appear at the junction between the two flanges, thus preventing arcing, reflections and leakage. As any particular waveguide size is used over a range of frequencies, the standard choke flange is designed to be most effective at the centre frequency of the range.

As the normal v.s.w.r. obtained with choke flanges is from 1.03 : 1 to 1.05 : 1 they are seldom used on precision low-power test equipment, but are found on most other components. Choke flanges should always be mated with a plane flange.

Ferrite Isolators

With many power sources it is important that the load should be reasonably matched as a high v.s.w.r. can cause frequency pulling, and in some cases cause failure of oscillation. There are several methods of overcoming this, the most efficient being the use of a ferrite isolator. This is a device that passes power in one direction with little loss, whilst having a high loss or attenuation when power is propagated in the opposite direction.

There are at least three types of ferrite isolators using slightly different principles. The commonest type, the resonance isolator, will be dealt with here. The other types the reader may encounter are the Faraday rotation isolator, and the field displacement isolator. Of these less common types the former may be recognized by the fact that the ferrite is mounted in a circular waveguide which is surmounted by a magnet. The latter may be distinguished from the resonance isolator by the fact that

*Canadian Marconi Company.

two slabs of ferrite plus some form of resistive element are used.

A cross-section of a resonance isolator is shown as Fig. 2. This device is basically a slab of ferrite material mounted across the narrow dimension of a waveguide in a strong constant magnetic field. Ferrite is a magnetic material having an extremely high resistivity. A microwave signal can therefore pass through it with very little loss. All magnetic materials have free spinning electrons which give them their magnetic properties. When the ferrite is placed in a strong constant magnetic field, the axis of spin of these electrons will line up with the field. If a microwave field is present in the waveguide,

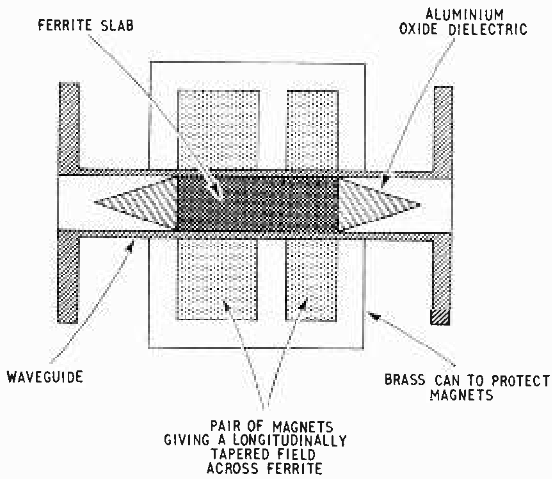

Fig. 2. Cross-section through a broadband ferrite isolator.

the magnetic field component will displace the axis of spin and cause it to precess about the axis of the constant (d.c.) magnetic field at a frequency, proportional to the constant magnetic field, termed the gyromagnetic resonant frequency. If the constant magnetic field is adjusted to make the gyromagnetic resonant frequency equal to the microwave field frequency, the attenuation constant through the ferrite will depend on the direction of propagation of the microwave signal. In one direction the direction of electron spin will allow the ferrite to pass the signal with little loss, whilst in the other direction will greatly attenuate it. As shown in Fig. 2, if a broadband device is required the magnetic field strength is varied over the length of the ferrite, often by using two magnets.

Ferrite devices are fairly new, and are in a continuous state of improvement. However, to give the reader an idea of the present state of the art, performance figures of two isolators currently available are given in the table. Insertion loss refers to loss in the "non-lossy" direction, whilst isolation refers to the loss in the "lossy" direction.

Waveguide Attenuators

As in low-frequency circuits, it is often necessary in microwave work to set or adjust the power level. The microwave equivalents to the common resistor and potentiometer are the fixed and variable attenuators. In these components a resistive film designed

Frequency Range (Mc/s)	Insertion Loss (dB)	Isolation (dB)	V.S.W.R.
5,950– 6,450	0.1	30	1.05
12,400–18,000	1.5	30	1.15

to have little inductance at microwave frequencies is deposited on a backing element designed to give little reflection and mounted in the waveguide so that a portion of the field is absorbed by the resistive element.

Two main types of variable attenuators are used. The transverse attenuator, shown in Fig. 3, typically uses a flat lozenge-shaped element of glass, ceramic or fibreglass. On this is evaporated an extremely thin film of resistive material, often nickel-chromium alloy, only a few microns thick. The resistive element is mounted on driving rods so that it is parallel to the narrow side of the waveguide. The element is moved across the guide, often with a micrometer drive to ensure "resetability." With a well-designed element maximum attenuation is reached when the resistive element is at the centre line of the guide, the attenuation law being approximately co-sinusoidal. The normal maximum attenuation of this type of attenuator is 40 dB with a minimum attenuation of 0.1 dB. A well-designed and manufactured unit can be calibrated to give an accuracy of ±0.02 dB or so. The main disadvantage of the transverse attenuator is the fact that attenuation varies with frequency, requiring a different calibration for each frequency used. It is, however, simple to manufacture and use, and can be made very compact.

The second type of attenuator in common use is the rotary vane attenuator, shown in Fig. 4. In this device the rectangular waveguide is transformed, usually by long tapers, to a circular waveguide. If the transitions are properly designed the H_{11} circular mode, which is very similar to the H_{10} rectangular mode, will be propagated in the circular guide. Let us divide the circular waveguide into three parts, making the centre section rotatable.

An attenuating vane, of similar construction to that

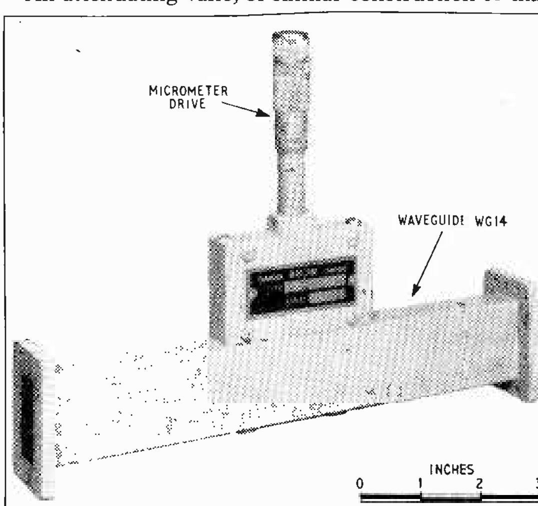

Fig. 3. Transverse attenuator with micrometer adjustment.

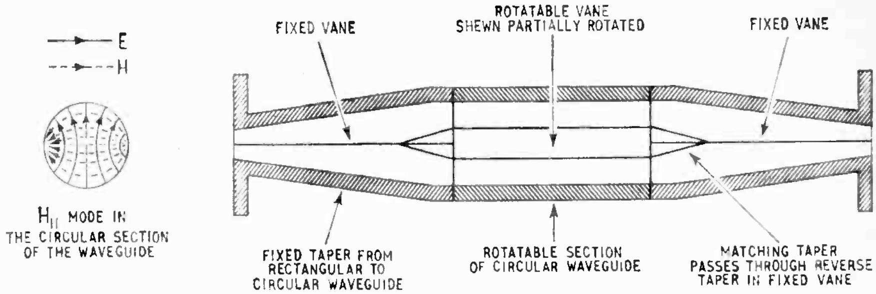

Fig. 4. Schematic diagram of a rotary vane attenuator.

used in the transverse attenuator, is mounted in the first fixed portion of the attenuator, passing exactly through the centre line, and in a perfectly horizontal position. If the H_{11} mode in the circular guide has not been rotated by manufacturing errors in the taper, the electric field will pass through the vane at right angles, and no attenuation will take place. If some rotation has taken place, any horizontal component of the electric field set-up will be greatly attenuated, thus effectively keeping the "E" field vertical at the horizontal centre line.

Considering now the rotatable portion of the attenuator, which contains a similar attenuating vane. When the vane is in a horizontal position no attenuation will take place, when in a vertical position maximum attenuation will occur. It can be readily seen that the attenuation will be a function of the angular movement of the vane. This function, mathematically the secant of the angular displacement, is, for an ideal attenuator, independent of frequency.

The second fixed vane in the output taper will further attenuate the signal by an equal amount, the process being similar, the final field being aligned with the output rectangular waveguide. The attenuation through the component therefore follows a secant squared law, the calibration being simply a secant squared scale. This gives a wide scale at the low end, very suitable for measuring small increments of attenuation. The scale is usually calibrated to about 50 dB, and as mentioned previously is independent of frequency. The main disadvantage of this instrument is its bulk, due to the need for long tapered transitions.

The Standing Wave Detector

To be able to measure the match of a piece of microwave apparatus, we must be able to detect and measure either the standing wave, or both the reflected and incident waves and compare them. Both methods are used, but only the former, which is the simplest and in general the most accurate, will be dealt with here. The other method, the reflectometer technique, is most suitable for swept frequency measurements, and is widely used as a production inspection tool.

As mentioned in the article on microwave theory, if a straight narrow slot is made down the exact centre of the broad wall of a waveguide, the signal will not be affected. This is the basis of the standing wave detector. A short probe is mounted in a slot of this type, and moved longitudinally along it. A current is induced in the probe proportional to the amplitude of the standing wave at the point in question. The probe signal is rectified and displayed on a meter, the difference between the minimum and maximum signals giving a measure of the v.s.w.r.

A line drawing of a standing wave detector is shown as Fig. 5. The ends of the slot are usually tapered as an additional aid in preventing reflections. Great care is taken to keep the slot parallel with the sides as any variation will, of itself, give a variation of output and thus an error in the v.s.w.r. measurement. The probe depth must be kept constant for the same reason. The runners of the probe carriage must therefore be machined accurately flat over the whole distance of travel. Similarly the bearings of the probe carriage rollers must have minimum eccentricity as this will give a cyclic variation of probe depth.

The probe itself consists of the inner and outer sections of a coaxial line. The inner section projects into the waveguide. The probe depth is often variable, although an empirical optimum depth frequently used is $12\frac{1}{2}\%$ of the narrow dimension of the waveguide. The outer sheath of the coaxial probe must be aligned within the inside wall of the waveguide. If this is not done radiation into the slot from the inner conductor can occur, creating a slot wave which will add to the measurement errors. As shown in Fig. 5, the gap between the probe and the slot wall is often packed with resistive material to aid in preventing slot waves and radiation in general.

Very often a microwave crystal rectifier is included in the probe carriage, together with a matching device to equate the impedance of the probe coaxial line with that of waveguide. Movement of the probe carriage along the slotted section is precisely

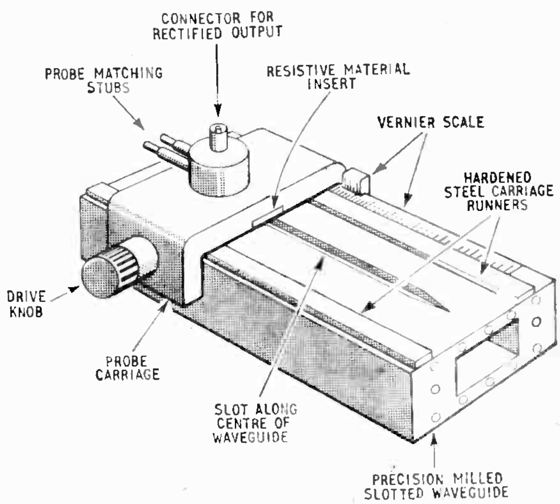

Fig. 5. Standing wave attenuator.

measured by the incorporation of a vernier scale, or a micrometer.

From the foregoing it will be appreciated that a good standing wave detector is a precision instrument, and as it is perhaps the most frequently used instrument in the microwave laboratory, will amply repay any additional care taken with it.

Wavemeters

As in low-frequency work a common requirement is the precise measurement of frequency. This can be accomplished at microwave frequencies by several methods. For example, the accurate measurement of the distance between two minima of a standing wave by a standing wave detector will give half the guide wavelength, from which the frequency may be easily calculated. This can, however, be rather tedious, and the usual method of measurement is with a wavemeter, or, as it is sometimes called, a frequency meter.

This device is quite simply a high "Q" resonant circuit. As mentioned in part three of this series, resonant circuits at microwave frequencies are obtained by the use of shorted lengths of line, either coaxial or waveguide. A half wavelength of line or any multiple thereof, shorted at both ends, will act as a parallel resonant circuit. For reasons that will not be covered here, circular waveguide cavities can be made to have extremely high "Q" factors, thus giving a very sharp resonance. This condition is of course ideal for a wavemeter, where high definition of frequency is required, and most waveguide wavemeters use this form of cavity.

The cavity itself is normally constructed from a highly polished silver-plated cylinder, made to very accurate dimensions. The frequency is varied by making one short circuit a variable plunger driven by an accurate micrometer type drive.

Wavemeters are normally made in two configurations. Perhaps the most frequently used is the absorption wavemeter. In this case the wavemeter is coupled directly from the side of a piece of waveguide forming part of the main circuit. Any signal at the frequency for which the wavemeter is tuned will be coupled into the wavemeter and absorbed, giving a sharp dip in the output. A line drawing of this type of wavemeter is shown as Fig. 6.

The main disadvantage of this type of wavemeter is that the sharpness of the dip is adversely affected by a high v.s.w.r. in the circuit, and as the power at the frequency of interest is absorbed, the device obviously cannot be left on tune in the circuit.

The other wavemeter configuration in general use

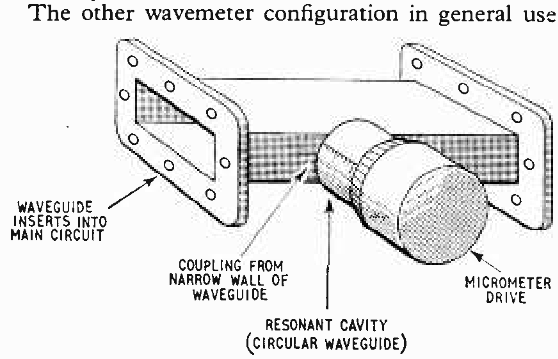

Fig. 6. Absorption wavemeter.

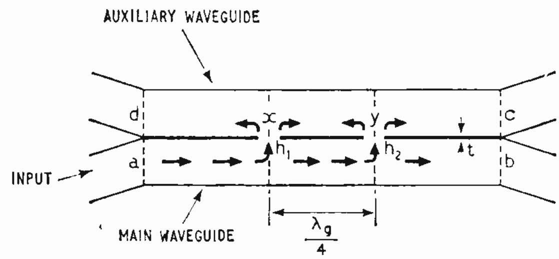

Fig. 7. Schematic diagram of a directional coupler.

is the reaction wavemeter. This device is coupled onto the end of a subsidiary waveguide line, and again absorbs some part of the signal to which it is tuned. Output is taken from the cavity by a probe in the side, that often incorporates a rectifying crystal, and thus provides a d.c. output when the wavemeter is tuned to the incoming frequency. The subsidiary line containing the wavemeter can be decoupled from the main waveguide by means of a directional coupler, as will be explained later. Standing waves in the main line will therefore have no effect on the wavemeter, which has the additional advantage of not affecting the output, so can be left in circuit to indicate any frequency drift. It has the disadvantage of requiring additional circuitry. All waveguide wavemeters can support an infinite number of modes, those supporting a harmonic of the wanted signal being obvious examples. The range of any wavemeter is therefore limited to avoid the ambiguity caused by these modes. In certain cases ingenious mode filters can be built into the cavity to suppress some of the unwanted modes.

Directional Couplers

In many microwave circuits it is required that a sample of the signal is taken from the main circuit. In most cases a portion of the signal from one direction only is required. This may be the incident wave, or the reflected wave depending on the circuit. This can be done quite simply by a device which is not normally used at low frequencies, the directional coupler.

Referring to Fig. 7, we have two waveguides with one broad wall common to both, and two identical holes in the common wall a quarter guide wavelength apart. Consider power fed into point (a). This will propagate down the main waveguide, and an amount proportional to the size of the holes and their position across the guide will be coupled through the holes h_1 and h_2 to excite the auxiliary guide, the remainder of the signal appearing at point (b). The power coupled through each hole will split equally, half travelling toward point (c) and half toward point (d). Consider now the waves and their relative phases at the reference planes x and y. If it is assumed that equal power is coupled through each hole, power will arrive at x from h_1 having travelled a distance $ah_1 + t$, and from h_2 having travelled a distance $ah_1 + \frac{\lambda_g}{4} + t + \frac{\lambda_g}{4}$. It will be seen that the path lengths are different by $\lambda_g/2$ or 180° and as the amplitude of each wave is equal, complete cancellation will take place. No power is therefore propagated in the auxiliary guide toward point (d). Consider next the reference point y. The path length of the wave from h_1 will be $ah_1+t+\frac{\lambda_g}{4}$, whilst that from h_2 is the same and will be $ah_1+\frac{\lambda_g}{4}+t$. The two waves will be in phase and add all the power coupled through the holes propagating toward point (c).

If, as in the example above, point (a) is the input port, the arm of the auxiliary guide ending at point (c) is called the coupling arm, whilst that terminated by (d) is defined as the directional arm. It will be noted that if the wave direction in the main waveguide is reversed, the coupling and directional arms will also be reversed. Thus any output at point (c) will always, in an ideal coupler, originate at point (a), whilst any output at point (d) will always, in any ideal coupler, originate at point (b).

The main parameters of a directional coupler are *coupling* and *directivity*. Other parameters often quoted are main and auxiliary waveguide v.s.w.r. and power handling. Coupling is defined as the ratio of the input power to the power appearing at the coupling arm, when the remaining parts are terminated with a matched load. This parameter is usually quoted in decibels and is given by

$$C = 10 \log_{10} \frac{P_{in}}{P_{(c)}}$$

where P_{in} = input power (watts), $P_{(c)}$ = power at point (c) (watts), C=coupling (dB), and is proportional to the hole dimensions, positioning and number, waveguide dimensions and frequency.

Directivity is defined as the ratio of the power output of the coupling arm to the power output of the directional arm when the remaining arm is terminated with a matched load. Again the parameter is usually quoted in decibels being given by

$$D = 10 \log_{10} \frac{P_{(c)}}{P_{(d)}}$$

where $P_{(d)}$ = power at point (d), and D=directivity (dB). In the ideal case, as complete cancellation occurs in this direction, directivity is infinite. However, for various reasons this does not occur in practice, the normal limit being 50-60 dB.

There are many designs of directional couplers. The holes may be in the broad or narrow wall, round or a variety of other shapes, and the two waveguides may be parallel, at right angles, or at a specific angle dictated by the design. They are all based on the principle of waves adding in one direction and cancelling in the other. The directional coupler shown in Fig. 8 is a three-hole side-wall coupler, designed to act as an incident power monitor. It has a matched load built into its directional arm, the coupling arm having a mount for a rectifying crystal to measure the power in that arm. This is a typical application, and the performance, which is also typical, is coupling 20±0.5 dB, directivity 20 dB minimum, frequency 5950 Mc/s to 6450 Mc/s.

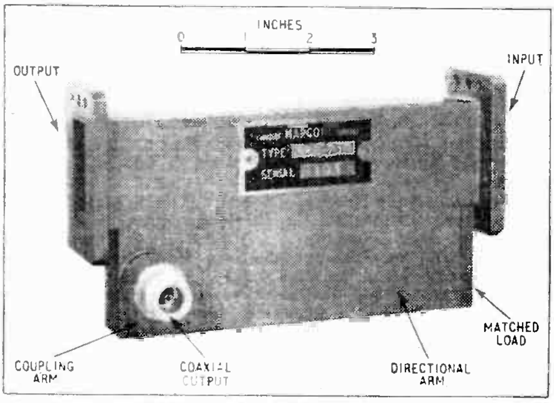

Fig. 8. Three-hole side-wall directional coupler.

Waveguide Loads

As has been mentioned previously, it is often necessary to terminate the end of a waveguide so that all the power is absorbed, and none reflected. It has been found that if the dimensions of the waveguide are changed very gradually, little reflection will occur. Simply tapering the waveguide to a point, however, will merely cause reflections from the point. If instead we insert into a short-circuited length of waveguide a tapered wedge of resistive material, reflections will be small due to the taper, and the power will be gradually absorbed by the resistive material. Any power remaining and reflected from the short circuit will again be greatly attenuated on its return path through the material. A diagram of this type of load is shown in Fig. 9.

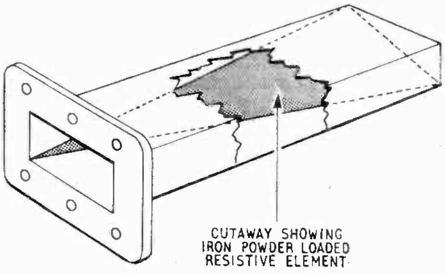

Fig. 9. Waveguide load.

The absorbing material is often a synthetic resin loaded with very fine particles of iron. The taper lengths are usually of the order of five times the guide wavelength, and the units are often finned to increase power dissipation. A standard load will have a v.s.w.r. of 1.01 or 1.02 to 1 over a waveguide bandwidth, although high-power components may have a mismatch as great as 1.15 : 1.

Waveguide-to-coaxial Transitions

As many microwave sources and outputs have coaxial feeders it is necessary to provide a transition between coaxial line and waveguide. For propagation to take place we must provide a suitable link between the field existing in the coaxial line and the field of the mode we wish to propagate in the waveguide. Propagation of the H_{10} mode, which we will consider here, is comparatively easy. A simple extension of the centre conductor of a coaxial line through the centre line of the broad wall of the guide is sufficient for narrow bandwidths. The length of penetration is made a quarter guide wavelength, whilst the waveguide is short circuited a quarter guide wavelength

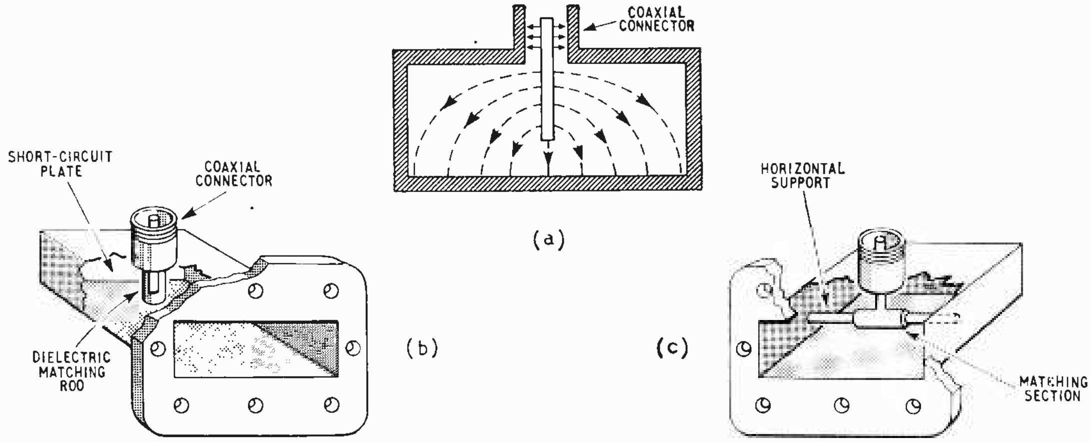

Fig. 10. Waveguide-to-coaxial transitions.

from the probe. A line drawing of this type of transition showing an approximate distribution of the electric field is shown as Fig. 10(a), Figs. 10(b) and 10(c) showing refinements of the basic technique giving a better match over a wider band.

Microwave Detectors

There are several methods of detecting microwave frequencies, but we shall concentrate on the commonest and possibly the simplest, the crystal diode. It has been found that one of the earliest of all detectors, the crystal and catswhisker, or to give it its modern name, the point contact diode, is ideal for the rectification of microwave frequencies. The crystal material is usually silicon and the diodes are normally made to a coaxial configuration for convenience in mounting. The characteristic is approximately square law over a fairly wide power range, but whenever possible should be operated at approximately constant power levels to ensure accuracy of measurements. The diode may be mounted in a coaxial line, which is attached to a waveguide to coaxial transition, or may be mounted directly in a waveguide, usually with a taper or some other form of impedance matching. A typical unit is shown as Fig. 11.

V.S.W.R. Indicator

The output obtained from our waveguide detector, whether it is an individual component, as just described, or part of a standing wave detector or other component, is at a very low level. For convenient use, particularly in measurement circuits, amplification is required. As stable d.c. amplification of low level signals is a rather expensive business, most microwave power sources intended for measurement work include square wave modulation at a few kilocycles. We can therefore use a high-gain, low-frequency amplifier with a meter output as an indicator in microwave measurements. As one of the primary uses of this indicator is the measurement of v.s.w.r., the scale is calibrated in this parameter, and the device called a v.s.w.r. indicator. Full-scale deflection is unity v.s.w.r., and the output obtained from the standing wave maximum is lined up with this reference. The standing wave detector probe is then moved to the minimum amplitude of the standing wave, the v.s.w.r. being read directly from the indicator scale. Modern v.s.w.r. indicators are designed to be quite versatile, and usually incorporate wide-range attenuators, power ratio scales, variable filters for various modulation frequencies, and many other useful additions.

In the next and final article in this series we will cover the more common microwave measurement techniques using the components just described.

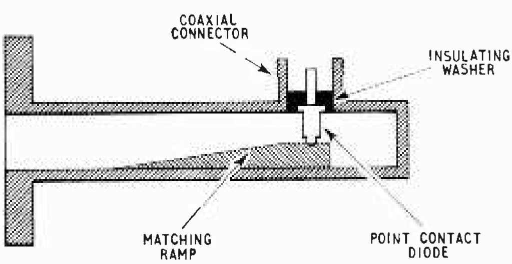

Fig. 11. Schematic drawing of a broad band crystal detector.

CLUB NEWS

Halifax.—A lecture-demonstration on transmitter alignment will be given to members of the Northern Heights Amateur Radio Society by L. M. Dougherty on November 25th. Fortnightly meetings are held at 7.30 at the Sportsman Inn, Ogden.

Heckmondwike.—Single-sideband operation will be discussed by A. W. Walmsley (G3ADQ) at the Spen Valley Amateur Radio Society meeting on November 26th at 7.30 at Heckmondwike Grammar School.

Ipswich.—Meetings of the Ipswich Radio Club are held on the last Wednesday of each month at 7.30 at the Civic College.

Loughton.—Peter Brooks, of the B.B.C., will talk about the Corporation's short-wave broadcasting techniques at the meeting of the Loughton & District Radio Society on November 6th at 7.30 at Loughton Hall, Debden Community Centre, Rectory Lane. The Society is collaborating with members of the British Amateur Television Club to present a closed-circuit television demonstration on behalf of the Television Viewers' Council at the hall during the period November 14th-21st.

Wellingborough.—The November meetings of the Wellingborough Radio Club, which meets each Thursday at 7.45 at Silver Street Club Room, include a lecture on electrochemistry by D. Slater (12th) and another on model aircraft and radio control by D. Britton (26th).

Television Distribution by Wire

2.—MAIN FEATURES OF A PRACTICAL H.F. SYSTEM

By R. I. KINROSS,* M.I.E.E.

IN this article it is only possible to concentrate on matters of general principle and a detailed description of all the items of equipment used in an h.f. system would be of little interest to the general reader. Nevertheless a very brief description of the more important items may not be out of place.

The system to be described is based on nine-pair cable of which six pairs can each carry one TV programme (vision plus sound) and three pairs can each carry one audio programme. (See Table 3.) Thus, during the next few years while we shall probably have only two B.B.C. plus two I.T.A. plus (possibly) one Pay-TV programme, the system will be capable of distributing five television programmes plus four sound programmes. When and if six television programmes have to be distributed, the sound programmes will be reduced to three if we stick to the straightforward principle of one audio channel per pair. However, it is quite feasible to distribute additional sound channels by means of phantom circuits (see Fig. 13) or by means of carrier channels. (Both these methods have been proved satisfactory in many towns.) These are not the most economical methods if 100% of subscribers were to require these extra channels since they require additional equipment in each home. The probability however, is that only a proportion of subscribers would require programmes such as stereo and in this case phantom or carrier would be the most economical overall solution.

Similarly as regards vision programmes on multi-pair systems it is undoubtedly most economical to devote one pair of wires entirely to each television signal, but there is no technical reason why two such signals should not be distributed on the same pair of wires. This has, in fact, been done on a system feeding several hundred thousand subscribers. There is no difficulty in manufacturing network transformers to cover the range 3-20 Mc/s. Thus, in the unlikely event of having to distribute more than six television programmes during the useful life of nine-pair cable, recently erected, additional programmes could be distributed using carriers of about 12 Mc/s and the problems associated with the separating and combining filters for this additional programme would not be anything like as formidable as those associated with separating and combining a total of seven television programmes on one coaxial cable.

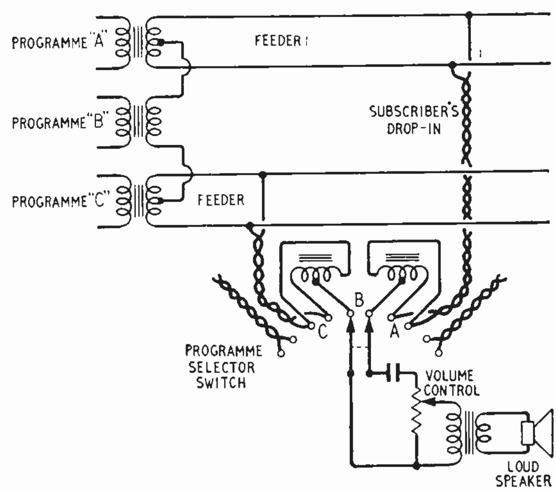

Fig. 13. Additional sound channels can be provided by phantom circuits.

Aerial Site:—Starting at the aerial site and assuming that the best possible aerials and low-noise amplifiers have been installed, we come to the system frequency generating equipment. In the days when only two 405-line programmes had to be distributed, frequency changers using crystal-controlled oscillators were used. The carrier output frequencies of these were at 5 and $8\frac{1}{2}$ Mc/s in order to produce the signals illustrated in Fig. 4. Subsequently, as more television programmes had to be distributed and exact synchronization between carriers became necessary, the outputs of these frequency changers were demodulated and re-modulated on a pair of ring modulators using a common crystal-controlled oscillator as a source of carrier. A more modern solution now becomes possible due to the method developed for reconstituting a colour sub-carrier from a short colour burst. It is now possible to remove most of the modulation from the signals emerging from a frequency changer by the use of a filter and limiter and feed the remaining signal to a comparator which compares its phase with that of another signal applied to it from a crystal-controlled oscillator. A d.c. signal proportional to the phase error is fed to a reactance valve which controls the frequency of the equipment producing the first signal. Several frequency changers can be controlled in this manner with the result that the signals which emerge have their carriers locked both as regards frequency and phase.

A useful refinement at the aerial site is equipment for removing any distortion present in the received signal. This distortion can be due to the transmitting authority or P.O. lines or the receiving site or echos from distant buildings or hills. Considering the reception of a single pulse, most of this distortion can be boiled down to the presence of one or

*Rediffusion Research Ltd.

more unwanted pulses either preceding or following the main wanted pulse. These unwanted pulses may be opposite in direction or in the same direction as the wanted and will vary in amplitude. By designing an adjustable equalizer fitted with a number of delay lines it is possible to add to the received signal a number of pulses equal in amplitude but opposite in direction to the unwanted pulses and so appreciably improve the quality of the received signal. An example of such an improvement is shown in Fig 14. This same equalizer can also be used to introduce a certain amount of predistortion to compensate for the distortion normally present in a wired or aerial receiver due to vestigial sideband reception. Clearly it is more economical to correct the signal once for a town rather than build correctors into each of possibly 10,000 receivers used in the town. This adjustable equalizer is not, however, used for correcting distortion introduced by cable and repeaters: the method of dealing with this is described later.

Trunk Distribution System:—This is the name given to the network used for carrying the sound and vision signals with the least possible distortion from the aerial site (which may be some distance outside the town) to the various areas of the town to be covered. The main considerations for this network are minimum distortion rather than minimum cost, since trunk cables normally comprise only some 6% of the network as a whole. We therefore aim to connect as few repeaters in tandem as possible and to this end choose cable of low attenuation. We also wish to introduce no crossview whatever between signals prior to these reaching the cheap multi-pair feeder cables. Vision pairs should therefore be screened from each other and the natural choice in this case is multi-coaxial cable: one cable per vision programme. The audio programmes are usually distributed on balanced pair cables operating at voltages up to 650 volts a.c. and carrying kilowatts of direct audio power to the various distribution points. These distribution points are usually a mile to $1\frac{1}{2}$ miles apart and it is from these points that the feeders using multi-pair cable radiate outwards to subscribers. No subscribers are ever connected directly to a trunk system.

These trunk cables are undoubtedly bulky and expensive but the modern tendency is to run underground nearly all cables that do not feed subscribers directly. The cost of trenching and making good can be anything from 15s. to 30s. per yard so that cable costs tend to shrink in significance compared with this.

Repeaters have a gain of about 50 dB and are equipped with two outlets: one of $4\frac{1}{2}$ volts for continuing the trunk route and one of 15 volts for energizing the start of a feeder. A fault on a feeder causing a reflection back to the repeater will have no effect on the signal on the trunk route.

The trunk cable and repeaters introduce a small amount of group delay distortion and every second or third repeater this is corrected by introducing a few corrector units of the form shown in Fig. 15. The choice of corrector unit is very simple. A pulse and bar 6 modulated signal is transmitted down the trunk system and corrected for least distortion by adjusting a variable group delay corrector. This consists of a number of switchable filters as illustrated in Fig. 15 and is designed to introduce an adjustable amount of delay to $\frac{1}{2}$ Mc/s intervals over the band 4 to 11 Mc/s. The amount of delay introduced at each frequency is read from the pointers of the knobs which are directly calibrated in millimicroseconds and the appropriate fixed delay equalizers are then permanently connected.

Since the attenuation of cable in dB varies as $\sqrt{\text{frequency}}$, a frequency equalizer, whose value depends on the attenuation of the preceding cable section, is plugged into each tandem repeater.

Feeder Distribution System:—Since it is desirable to keep the number of repeater kiosks down to a minimum the trunk and feeder repeaters are usually combined as described above. Should it sometimes be necessary to start a feeder at some point other

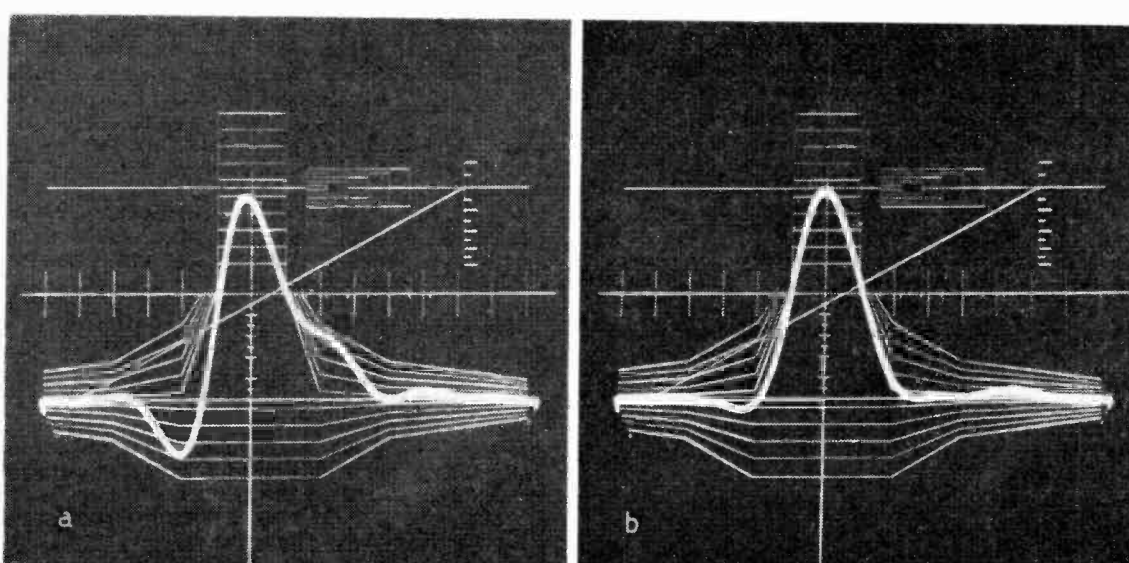

Fig. 14. Correction of pulse by means on an echo equaliser (a) before correction, (b) after correction.

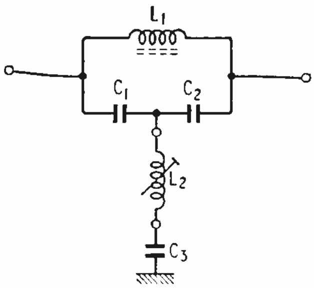

Fig. 15. Group delay corrector unit.

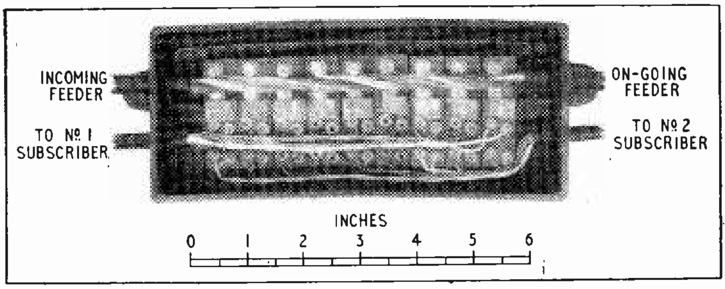

Fig. 16. Subscriber's junction box.

than at a trunk repeater, the input of a simple feeder repeater is bridged across the trunk cable. In theory a feeder repeater can energize some 20,000 yards of feeder but in practice, due to the fact that most feeder distribution areas contain parks, etc., 15,000 route yards is a normal useful coverage achieved. However, this compares very favourably with the figure of only some few hundred yards normally energized by a v.h.f. feeder repeater. Since the average amount of cable per house is some 10 to 15 yards the number of houses per h.f. repeater point vary from 1,000 to 1,500.

The foregoing refers to repeaters using valves capable of delivering some 15 volts of television signal to the start of the feeder. Occasionally it is necessary to feed small pockets of residential areas where it is not worth installing an amplifier of this type. In this case a small transistor amplifier is used, measuring 7in by 1¼in with a gain of 30 dB and an output of one volt. This is capable of energizing about 3,000 route yards of feeder. It requires no mains since it is energized by a nickel-cadmium battery, which is kept charged by means of a 50-volt supply fed to it over a phantom circuit using the feeders for this purpose.

The h.f. system, of course, requires a separate repeater for each programme, but since the number of repeater points on an h.f. system are usually 1/30th of those needed for a v.h.f. system, the total amount spent on repeaters is much less for an h.f. system.

The audio signals are stepped down from the trunk system to 55 volts by means of transformers.

The nine-pair cable, thus energized with vision and audio signals, is usually clipped under the eaves of houses where it is very inconspicuous, yet handy for the connection of a short drop-in cable to the subscriber.

Subscriber's Installation:—This consists of bridging transformers, a drop-in multi-pair cable, a programme selection switch and a wired receiver as illustrated in Figure 5.

It may be asked how it is possible to connect a subscriber quickly and economically on to all the pairs of cables that have been discussed. The solution lies in the design of a very cheap plastic junction box and the development of a method of connecting subscribers to the feeder spur in such a way that it is unnecessary to cut the spur cable or to bare the insulation of the spur cable. Thus the service is not cut off from subscribers receiving their programmes farther down the network and a subscriber can be connected to a multi-pair cable extremely quickly. A subscriber's junction box is illustrated in Fig. 16. Six of the inserts contain a high-frequency transformer which steps the vision voltage on the feeder spur down to a level suitable for the subscriber and at the same time presents an impedance across the feeder spur high enough to cause no reflection. It also transfers the audio signal directly to the subscriber as illustrated in Fig. 5. The remaining three inserts are straight through connections from the feeder spur to the subscriber drop-in and deal with audio signals only. Each of these subscriber inserts is capable of feeding two subscribers, but another version is capable of feeding four subscribers. It is thus possible to feed up to four subscribers with six television programmes and their sound plus a further three sound programmes from this junction box which overall measures 7in × 3in by 1in. The method of connecting the feeder insert to the spur without stripping the insulation of the latter is by means of sharp-toothed phosphor-bronze washers which are placed between the insert and the terminal screw head and these bite through the insulation of the feeder spur. Subscribers are connected by means of a pin and eyelet which strips the insulation from the drop-in wires as it is pushed on.

The programme selection switch consists only of a multi-way two-pole switch which can be made from a single standard switch wafer.

The wired receiver contains two h.f. pentodes for system frequency amplification and from then on follows conventional aerial receiver design. The sound is fed direct to the loudspeaker at 55 volts via a volume control and transformer. Thus the v.h.f. and u.h.f. turrets are eliminated as are also the complications of dealing with both a.m. and f.m. sound which is, of course, necessary on a 405/625 aerial receiver.

If a subscriber already owns an aerial receiver then an "inverter" is connected between the programme selector switch and the receiver. This converts the system frequencies into vacant channels in Band I or Band III. When the aerial receiver wears out and is due for renewal the subscriber finds he can buy or hire the simpler wired receiver for about two-thirds of the price of a new aerial receiver and usually does so. The inverter is then recovered and used in another home.

Colour:—Colour has so far only been mentioned in passing while considering such matters as crossview or changes in attenuation.

The availability of a multi-pair network clearly provides the engineer with many methods of distributing colour information which are not available and could not be used for normal broadcasting. In broadcasting it is essential, whether the system used be PAL, SECAM or N.T.S.C., that the colour information should be contained within the bandwidth allotted to the transmitter for monochrome transmissions. The engineer responsible for designing a wired television system need have no such inhibitions if he feels that a simpler and cheaper overall system can be developed by transmitting the colour information outside the normal monochrome band.

However, to start with, the growth of colour television is likely to be fairly slow and it is important to make certain that signals distributed on networks can be satisfactorily received not only on special simple "wired" receivers but also on normal aerial receivers. Care has been taken, therefore, to ensure that all cable and repeaters installed on a wired system are capable of handling "in band" colour information such as that required for the N.T.S.C. system. In practice, providing care is taken over the linearity of repeaters and group delay correction is introduced in order to maintain a good monochrome picture no difficulty is experienced in distributing an N.T.S.C. type of signal in the h.f. band. These precautions add very little to the overall cost of the system. Later on, if it was thought that, for example, an "out of band" system would result in a material saving in the cost of the colour decoder in each receiver, the information for this could be distributed in addition to the "in band" colour information. This would, of course, slightly increase the cost of the repeaters on the system and would not be done until such a time as sufficient of a cheaper type of wired colour receiver could be connected to the network to warrant this additional extra expense.

Comparison Between H.F. and V.H.F. Systems Comparative Costs

There is undoubtedly room for the growth of both h.f. and v.h.f. systems but it is worth giving some consideration to the conditions which favour the use of one system rather than the other.

Let us start by listing all the items which will cost about the same for either type of system. These are:—

Aerial arrays, low-noise amplifiers, system origination equipment, wayleaves, feeder cable, cable fixings, feeder cable erection labour, trunk system (though the h.f. trunk system is more expensive per 100 yards there is less of it due to the fewer distribution points it has to feed.

The main items costing more for the v.h.f. system are the trunk and feeder repeaters. Usually thirty times as many repeater points are required so that even using broad-band repeaters for v.h.f. this means $7\frac{1}{2}$ times as many repeaters for a system distributing four television programmes.

The items which cost more for the h.f. system are:—

Audio amplifiers, subscribers' drop-in (more complicated junction box and the need for a programme selection switch), inverter (if required).

In a large town calculations show that most of the above differences cancel out and that there is very little difference in the cost of the two systems.

We are left, therefore, with the main point of the whole exercise, namely the difference in cost between a wired and an aerial receiver.

The latest figures obtainable on this are as follows:—

Average retail price including Purchase Tax of the cheapest models marketed under fifteen different brand names of 405/625-line 19in receivers equipped for u.h.f. reception. (Ref. *Wireless and Electrical Trader* Buyers' Guide, No. 9, 1963)	£70	0	0
Retail price including Purchase Tax of 19in Rediffusion Mark VIII wired receiver completely ready for 405 and 625-line programmes	£49	0	0
Difference	£21	0	0

This difference might be reduced by some £6 if a switchable aerial receiver is bought without a u.h.f. front-end since only v.h.f. is usually needed in a communal aerial system.

From the above it will be clear that in a block of flats in a town without an h.f. system or in an area so small that the number of v.h.f. repeaters is of little consequence, the choice may well be for a v.h.f. system. In a hotel, however, where the owner (as distinct from the owner of a block of flats) would be interested in owning or renting the cheaper wired receiver, the choice will almost certainly be for the h.f. system.

Of course, if the block of flats was in a town already wired for an h.f. system, the arguments are more evenly balanced. The owner wants to have *some* wired system to prevent the erection of private aerials, but he does not wish it to be thought that he is forcing his tenants to buy or hire a certain make of wired receiver. In fact he need have no qualms, since wired receivers are now being made by several manufacturers and marketed through normal retail channels.

It is interesting to note that in certain blocks of flats where both systems have been installed there is now a far greater number of subscribers on the h.f. system.

When introducing a wired television system to a town today, the fact that most of the inhabitants will already own a television set must at first sight appear to be a strong argument in favour of a v.h.f. system. However, the way in which owners of aerial receivers have eventually changed over to wired receivers in towns wired for h.f. a few years ago, coupled with the fact that a large proportion of aerial receivers in current use will not be able to receive B.B.C. 2, has encouraged h.f. system operators to continue introducing their systems into new towns.

With the advent of Pay TV it is perhaps worth mentioning a further point in favour of an h.f. system: the availability of its multi-pair cable considerably reduces the cost and complexity of the equipment needed to control the coin boxes or credit meters in subscribers' homes compared with what is required on a v.h.f. coaxial system.

The New Towns Commission recently decided, after considering very carefully the relative merits of h.f. and v.h.f. systems for Hemel Hempstead and Crawley, to adopt an h.f. system. Similarly the B.B.C. has decided to instal an h.f. rather than a v.h.f. system at the White City.

Acknowledgements.—Acknowledgements are due to the Directors of Rediffusion Limited for permission to publish this article, to cable manufacturers who have co-operated in the work of designing special cables and to the I.E.E. for permission to quote from two of the papers read to the International Television Conference in 1962.

The author would also like to thank his colleagues throughout the Rediffusion Group who have done so much to make the h.f. distribution system successful in large-scale operations.

Correction.—In Fig. 6 in the first part of this article (p. 499, October issue) the left-hand curve should be (c) and the right-hand curve (a).

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COMMERCIAL LITERATURE

Six more **planar epitaxial transistors** have been added to the SGS-Fairchild range. Data sheets on these semiconductors are available from the company's offices at 23 Stonefield Way, Ruislip, Middx. Transistor type 2N3304 is a high-speed device (700 Mc/s) for logic applications with 30 n/sec maximum storage time; 2N3209 is a 20V version of the 2N2894, which is a 550 Mc/s unit intended for fast switching and r.f./i.f. amplification; 2N3120 and 2N3121 are 45V versions of the 2N2927 and 2N2696 which are designed for use as either high-current switches or v.h.f. amplifiers; and 2N3072 and 2N3073 are 60V versions of the 2N3120 and 2N3121, which are also designed for use in v.h.f. amplifiers and high-current switching circuits.
14WW 301 for further details

A brochure describing the German Elac "Studio Series" of **transcription units** has been forwarded to us by the company's United Kingdom agents, The High Fidelity Centre, 61 Woking Street, Dorking, Surrey. Both of the units described are fitted with four-speed motors, and can be operated from any 220 V, 50 c/s or 115 V, 60 c/s supply.
14WW 302 for further details

Mullard Permanent Magnets is the title of a 20-page booklet available from Central Enquiry Handling, Mullard Ltd., Mullard House, Torrington Place, London, W.C.1. The introduction contains information on the properties, and the manufacturing proccesses of Magnadur, Ticonal and Reco magnets. Further sections of the booklet give information on general magnetic theory, applications of permanent magnets, magnetic circuit design, and a table of magnetic symbols.
14WW 303 for further details

Inductive, photoelectric and ultrasonic **proximity switching** is described in a leaflet available from Mec-Test Ltd., 218 Dover Road, Folkestone, Kent. Solid-state switching is featured. Details of their loop system, which is capable of sensing large masses of both ferrous and non-ferrous metals at distances from ½ in to 36 in are given.
14WW 304 for further details

High Frequency Crystal Filters is the title of a 48-page brochure (MQ/108) describing 22 types of filters specially designed for use in mobile radio equipment, with either 12.5, 20, 25 or 50 kc/s channel spacing. Copies of this publication are available from the Quartz Crystal Division of Standard Telephones and Cables Ltd., Edinburgh Way, Harlow, Essex. The characteristics of each device are given in tabular and graphic form.
14WW 305 for further details

The Sophisticated Oscilloscope is the title of a six-page publication written by John Kobbe, manager of the advanced circuitry department of Tektronix Inc, (first published in *Industrial Research*, U.S.A., in March of this year). It briefly describes the evolution of the cathode ray oscilloscope and comments on the present-day limitations and recording facilities. Future developments are also discussed. Copies are available from Tektronix U.K. Ltd., Beaverton House, Station Approach, Harpenden, Herts.
14WW 306 for further details

Avo Transistor Manual.—The second edition of the Avo International Transistor Data Manual is now available, price 35s post free, from Avo Ltd., 92-96 Vauxhall Bridge Road, London, S.W.1. It contains information on approximately 5,200 transistors, and includes a number which are now obsolete. The products of some 90 transistor manufacturers and distributors' products are listed from countries all over the world. Names and addresses of the manufacturers are included, and, where known, the name and address of the United Kingdom subsidiary company or agent is given. A complete cross-reference between Commercial and Services Common Valve (CV) specifications is also included in this 217-page publication.

Leaflets describing the Series 433 **tape recorder counters,** whose features include push button reset, are available from English Numbering Machines Ltd., Queensway, Enfield, Middx. Three- and four-digit counters are described.
14WW 307 for further details

LETTERS TO THE EDITOR

The Editor does not necessarily endorse opinions expressed by his correspondents

Pulse-counter F.M. Receivers

MR. Spencer, in the October issue, is worried about the possible television interference which might be caused by a pulse-counter f.m. receiver such as I described in my letter in the September issue.

He is, of course, quite right to show such concern, and I, too, have been very much conscious of the problem.

Both the original Scroggie design (June 1956), and the more recent crystal-controlled transistor design, operate the local oscillator at about 30 Mc/s to avoid producing interference with the London 45 Mc/s television transmissions. However, I live in the Midlands, where Sutton Coldfield (61.75 Mc/s) is the relevant television transmission, and this is why I decided to operate the local oscillator at about 45 Mc/s. If the local oscillator had been at about 30 Mc/s, its second harmonic could cause interference with television reception, whereas, in this part of the country, 45 Mc/s and its harmonics would seem to be relatively harmless.

Thus, it seemed to me that, if the circuit was to be published, I really ought to produce two designs—a London model and a Provincial model. This is a further reason why, in the event, nothing at all got published!

I did, however, take reasonable care to keep the amount of local oscillator voltage reaching the aerial socket to a minimum, and a rough measurement showed that both the fundamental and the second harmonic were well under 1 mV r.m.s.

It is, perhaps, worth contemplating the fact that the amount of unwanted power radiated by a receiver of this type, which has been well designed, is unlikely to exceed a thousandth of a microwatt. Mr. Scroggie, on page 182 of the April 1958 issue, mentions "the complete absence of a single reported case of interference of any kind by this type of f.m. receiver." Nevertheless, such receivers are basically more likely to cause interference than those with the usual 10.7 Mc/s i.f., and should be designed with all the more care because of this.

My reason for not operating the local oscillator at about 90 Mc/s was mainly that this would have made the circuit more difficult to tune up; with an i.f. as low as 160 kc/s, the r.f. and local oscillator tuned circuits would differ in frequency by only a small fraction of 1%, and the inevitable slight coupling between them would make the tuning adjustments of these circuits undesirably interdependent.

The use of a double superhet was considered, but was rejected on the grounds that it lacked the attractive simplicity of the system finally adopted.

Great Malvern. P. J. BAXANDALL

Transistor Circuit Design

IN the September issue on page 436 in the part dealign with bypassing of the emitter resistor R_E, Mr. Hobbs states that the bypass capacitor is decoupling r_e, and because this is only 6Ω in his example then very large capacitances are necessary if adequate low-frequency performance is desired.

I feel that Mr. Hobbs has overlooked the fact that the statement is true only if the impedance of the base circuit is insignificant. Any practical circuit will have a source impedance of, typically, 2,000Ω feeding into the base, hence r_e will be effectively in series with this resistance, R_s, referred to the emitter circuit, i.e. R_s/β.

Taking Mr. Hobb's value for β as 50 means that the resistance to be decoupled is 46Ω. Thus only one-eighth approximately, of the capacitance Mr. Hobbs indicates would be required for a given low-frequency performance.

Page 145 of "The Junction Transistor," by E. Wolfendale, also covers this aspect of decoupling.

Gerrards Cross, Bucks. B. H. GROSE

The author replies:

Mr. Grose in his letter makes interesting comment, although following upon my heading "Example of a Voltage Amplifier" it was a voltage drive that was being considered. No doubt he has a cascade arrangement in mind where the collector of one stage is coupled to the base of the next and where the collector load forms the source impedance to the second stage. Taking a figure of 2,000Ω for the collector load, the reflected resistance at the emitter would, I agree, be 40Ω and the size of the decoupling capacitor could be made eight times smaller.

Mr. Grose's example is more aptly covered by the succeeding paragraph on page 436 (September issue) where, for a relatively large source resistance, the transistor is treated as a current amplifier with current gain β. I would again maintain that it is wisest not to allow β to influence the overall gain of an amplifier and would say that it often pays to leave the emitter undecoupled. There is a technique, shown in Fig. 4, page 521 of the October issue, using alternative series feedback and shunt feedback stages where an emitter is decoupled straight to a rail and yet the voltage gain is determined solely by resistor values.

Flexibility is perhaps a watchword in transistor circuit design and, in describing a voltage amplifier, I was attempting to be as general (and therefore as helpful?) as possible. That many people think principally in terms of cascade amplifiers with one collector coupled to the next base, I had overlooked.

One attendant drawback of using the source impedance as an element involved in the low-frequency time constant is that the reflected impedance at the emitter is dependent on the current gain β; the low-frequency roll-off will vary with selection of transistor. How much neater it is to place the resistance physically at the emitter. As manufacturers improve their β tolerances, our problems will diminish and the outlook for the future will be rosier.

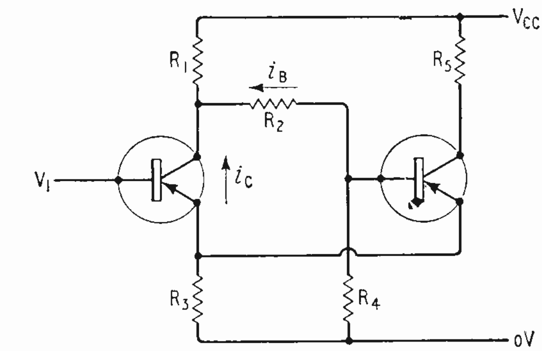
G. P. HOBBS

Trigger Circuit Design

IN his article of August, 1964, Mr. T. D. Towers touched on the basic design of the Schmitt trigger amplifier. With regret I note he has followed the all too common line of dealing with the circuit parameters "piecemeal," and referring those interested to other papers. Most of the literature tends to treat this simple circuit either as a design for the amateur magician (if you do not know, the writer prefers not to divulge!) or else quotes equations requiring a mathematical genius for analysis and solution.

Recently my work has required me to find a quick way to design such circuits capable of being built without undue dependence on device characteristics. Perhaps the method may interest others, as being free of the veil that often surrounds this circuit.

As a first step select the supply voltage V_{CC} to be at least ten times the voltage across the transistor $V_{CE(sat)}$.



Take i_C (as value in data) for $\overline{h_{fe}}$. So that

$$R_3 = \frac{\underline{V_1} - V_{BE(sat)}}{i_C} \text{ and } R_5 = \frac{\underline{V_1}[V_{CC} - \overline{V_1} - V_{CE(sat)}]}{\overline{V_1} \, . \, i_C}$$

Assume " i_B " $\approx \frac{i_C}{2}$ for initial calculation only.

$$R_2 = \frac{V_{CE(sat)}}{\text{“}i_B\text{”}} \text{ and } R_1 = \frac{\overline{V_1} - V_{BE(sat)}}{\text{“}i_B\text{”}}$$

$$\text{Hence } i_B = \frac{\overline{V_1} + V_{CE(sat)}}{R_2 + R_4}$$

$$\text{Giving } R_1 = \frac{V_{CC} - \overline{V_1} - V_{CE(sat)}}{i_C + i_B}$$

At this stage all resistors are specified for controlled backlash ($\triangle V_1$) and a check may be made of minimum gain required.

$$\underline{h_{fe}} = \frac{V_{CC} - \overline{V_1} - V_{CE(sat)}}{R_5\left[\frac{(V_C - \underline{V_1})}{R_1 + R_2} - \frac{\underline{V_1}}{R_1}\right]}$$

The design is not exhaustive but allows a clear circuit requiring little adjustment in practice, while each step is clear in itself to check on variations in one factor where this is required.

London, S.W.11. D. R. BILSTON

[The author's convention is to use bars to indicate levels above and below which the trigger will respond—Ed.]

Thévenin and Norton

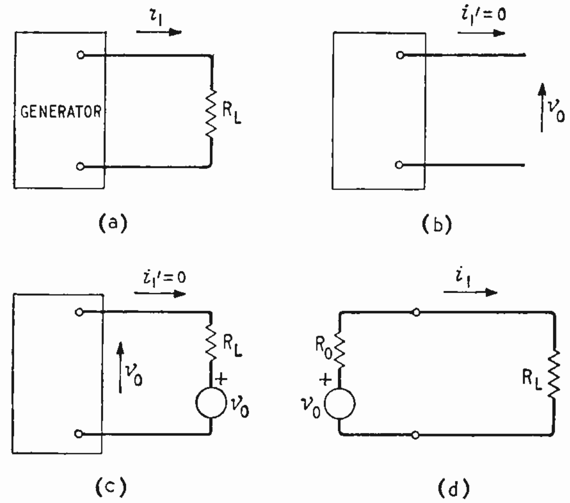
IN your January issue, " Cathode Ray " suggests the existence of a theorem which relates the power loss in a real generator to the power loss in its Thévenin (or Norton) equivalent. Such general relation can easily be shown to exist for purely resistive networks if one is allowed to dig up a little-known theorem due to Pomey (J.-B. Pomey: " Analogies Mécaniques de l'Electricité," Gauthier-Villars, 1921, pp. 50-51), according to which in any linear, lumped parameter and resistive network the sum of the products of the branch e.m.fs. which determine a certain electrical state (I) of the network by the corresponding branch currents of a second state (II) is equal to the sum of the products of the branch currents of state (I) by the corresponding branch e.m.fs. of state (II).

The theorem suggested by " Cathode Ray " can be proved as follows:

Consider a generator (a) in the accompanying diagram, which contains only independent e.m.fs. of any waveform and ideal resistors. Let us consider for the time being the generator and its load as a single network having b branches numbered in such a way that the load (passive) branch, R_L, is branch No. 1. Let e_k and i_k be the series e.m.f. and current typical of branch No. k. State (I), corresponding to (a), will be given by the following array of branch e.m.fs. and currents:

$$\left.\begin{array}{llll} 0, & e_2, & \ldots\ldots\ldots & e_b \\ i_1, & i_2 & \ldots\ldots\ldots & i_b \end{array}\right\} \quad \text{(I)}$$

Since the network is purely resistive, the instantaneous



power, p_g, lost internally in the loaded generator will be given by the sum of the instantaneous powers supplied by the individual branch e.m.fs. less the instantaneous power dissipated at the load, i.e.:

$$p_g = e_2 i_2 + \ldots . + e_b i_b - i_1^2 R_L \quad \ldots \quad \ldots \quad (1)$$

If we disconnect the load, the new situation will be as pictured in (b), where v_o is the open circuit voltage of the generator. As far as the branch currents are concerned, no change will be detected if we substitute the network represented in (c) for the one represented in (b). For (c) we have then the state (II) given as follows:

$$\left.\begin{array}{llll} -v_o, & e_2, & \ldots\ldots\ldots & e_b \\ 0, & i'_2, & \ldots\ldots\ldots & i'_b \end{array}\right\} \quad \text{(II)}$$

where i'_k represents the new value taken by branch current No. k in the generator network when its output terminals are open-circuited. Let p'_g be the instantaneous power loss in the generator under such condition. We have

$$p'_g = e_2 i'_2 + \ldots . + e_b i'_b \quad . \quad \ldots \quad \ldots \quad (2)$$

Referring now to (d), which shows the Thévenin equivalent of the real generator, let p''_g stand for the instantaneous power loss in the equivalent resistor, R_o, so that

$$p''_g = v_o i_1 - i_1^2 R_L \quad . \quad \ldots \quad \ldots \quad \ldots \quad (3)$$

From (1) and (3) it is possible to write

$$p_g - p''_g = -v_o i_1 + e_2 i_2 + \ldots . + e_b i_b \quad . \quad \ldots \quad (4)$$

We now apply Pomey's theorem to states (I) and (II), thus obtaining

$$e_2 i'_2 + \ldots . + e_b i'_b = - v_o i_1 + e_2 i_2 + \ldots . + e_b i_b \; . \; (5)$$

By substitution in (4) we have

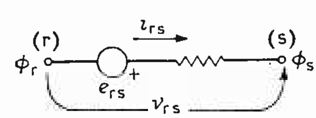
$$p_g - p''_g = e_2 i'_2 + \ldots . + e_b i'_b \, , \quad \ldots \quad \ldots \quad (6)$$

which, by comparison with (2) yields

$$p_g - p''_g = p'_g \, , \quad \ldots \quad \ldots \quad \ldots \quad \ldots \quad (7)$$

which proves the following theorem: *The amount by which the instantaneous power loss in a purely resistive generator exceeds that in its Thévenin equivalent is constant with load resistance and equal to its instantaneous no-load loss.*

Of course, this theorem can be expressed in terms of the Norton equivalent. It would appear also that nothing fundamental in the proof given above would oppose the extension of it to networks with reactive elements provided the term " instantaneous power loss " is interpreted as the instantaneous power dissipated in the resistive elements plus the instantaneous rate of change of the energy stored in the inductive and capacitive elements.



(e)

The resistive case may be proved by reference to a typical branch joining nodes r and s, as shown in (e), let v_{rs} be the potential rise from r to s, and i_{rs} be the current flowing from r to s, for a certain excitation of the network; let v'_{rs} and i'_{rs} be the corresponding quantities for a second excitation of the given network. We shall first show that

$$\sum_{r,s} v_{rs} i'_{rs} = 0 = \sum_{r,s} v'_{rs} i_{rs} \quad \ldots \quad (8)$$

If ϕ_r and ϕ_s denote, respectively, the potentials of nodes r and s in the first state and ϕ'_r and ϕ_s represent the corresponding quantities in the second state, we can write

$$v_{rs} i'_{rs} = (\phi_s - \phi_r)\, i'_{rs}$$
$$v'_{rs} i_{rs} = (\phi'_s - \phi'_r)\, i_{rs} \quad .$$

Now

$$\sum_{r,s} v_{rs} i'_{rs} = \sum_{r,s} \phi_s i'_{rs} - \sum_{r,s} \phi_r i'_{rs}$$
$$= \sum_{s} \left[\phi_s \sum_{r} i'_{rs}\right] - \sum_{r} \left[\phi_r \sum_{s} i'_{rs}\right] .$$

By Kirchhoff's current law, $\sum_{r} i'_{rs} = \sum_{s} i'_{rs} = 0$, so that $\sum_{r,s} v_{rs} i'_{rs} = 0$. By the same reasoning we get $\sum_{r,s} v'_{rs} i_{rs} = 0$.

All this shows that (8) is true, thus generalizing a theorem by Tellegen (cf.—G. Newstead: "General Circuit Theory," Methuen, 1959, pp. 30-31).

We now observe that

$$v_{rs} = e_{rs} - i_{rs} R_{rs}$$
$$v'_{rs} = e'_{rs} - i'_{rs} R_{rs} \quad .$$

By substituting in (8) we obtain

$$\sum_{r,s} (e_{rs} - i_{rs} R_{rs})\, i'_{rs} = \sum_{r,s} (e'_{rs} - i'_{rs} R_{rs})\, i_{rs} \quad .$$

Hence

$$\sum_{r,s} e_{rs} i'_{rs} - \sum_{r,s} e'_{rs} i_{rs} = \sum_{r,s} (i_{rs} i'_{rs} - i'_{rs} i_{rs})\, R_{rs} = 0 ,$$

which shows that Pomey's theorem is true for the resistive case, since the above implies

$$\sum_{r,s} e_{rs} i'_{rs} = \sum_{r,s} e'_{rs} i_{rs} \quad \ldots \quad (9)$$

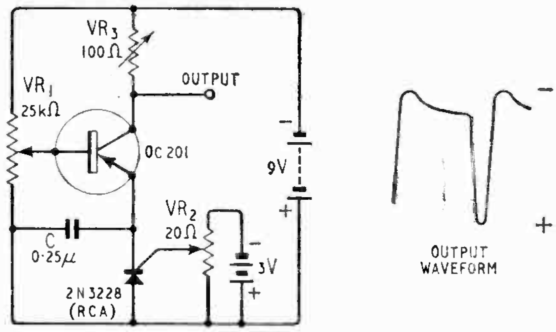
Cuba. J. ALTSHULER
Faculty of Technology, University of Havana

Thyristor Oscillator

ALTHOUGH there have been many articles describing the use of thyristors as protection elements in power supplies in your pages, their use in oscillator circuits seems to have been rather neglected. To remedy this, I would like to describe a circuit I have designed, using a thyristor, which produces an output waveform like that from a blocking oscillator, but with a sharper spike. The oscillator is self starting and does not require a blocking transformer.

The *modus operandi* is as follows. The initial conditions are such that the thyristor is in its high resistance state and the base potential of the transistor is set by VR_1 so that the transistor conducts thus charging capacitor C. The gate of the thyristor is biased from a low impedance source, i.e. VR_2, so that the thyristor "fires" and goes into its low resistance state before the transistor has been cut off by the rising emitter potential. The capacitor then discharges in two stages, thus producing a droop before the main discharge. The reason for this is not yet clear but is probably due to the transistor impedance at this stage in the cycle. As the capacitor undergoes its main discharge through the thyristor, the emitter potential of the transistor falls thus causing the transistor to conduct once again. At the same time the voltage across the thyristor has fallen to such an extent that the thyristor reverts back to its original high resistance state.

An output can be obtained by inserting a load resistance in the collector lead to the transistor. This resistance must be less than 80 ohms otherwise the circuit will not oscillate.



Frequency control is obtained by varying VR_1 and VR_2 within limits. This seems to have little effect on the width of the spike but alters the time interval between succesive spikes. Taking the base more negative and/or taking the gate more positive reduces this time thus causing an increase in frequency.

Applications of this circuit are numerous. By replacing the OC201 by a power transistor and the load resistor by a step-up transformer a d.c. to a.c. converter is possible. This would be very easily stabilized as a feedback voltage could alter the operating frequency directly. An e.h.t. generator would also be possible. Perhaps the most obvious application would be as a replacement to the blocking oscillator itself, thus doing away with one objectionable component, the transformer.

The device could also be used as a pulse code and possibly an f.m. modulator, the modulating voltage being applied to the base of the transistor.

Melton Mowbray. J. SLOMKOWSKI

Double Negative

IN your issue of September, 1964, I read a small joke article on page 447, entitled: "Ici on parle le double-talk," written by "Vector."

I have read this small insert with a smile. However, I do not think that "Vector" is conscious about putting the different levels of attenuators and amplifiers correctly. He is writing about an attenuator having a level of minus 12 dB, which I think is pushing very hard, because the whole article is dealing with a big noise around the visitor and if he adds an attenuator, having a rate of −12 dB, I would think that the whole level on the ear of the poor visitor will be very high, for, according to standard rules, negative numbers of dB markings are always amplified markings, whereas positive numbers are attenuated markings.

Due to the fact, that the denomination dB, as well as Np is by way of principle a logarithmic relation dealing only with fractions, that means:—attenuation. I suppose, however, that "Vector" is quite aware of this fact and this misplacing of negative and positive sign is part of a joke, or even part of a mistake, which slipped into his pen

Baar (ZG), Switzerland. O. STÜRZINGER

The author replies:

Herr Stürzinger is correct in his second supposition that the intrusive minus sign was a slip of the pen. I wish I could blame it on the typesetters but, alas, I find it appears on my original draft and was not properly corrected. The original sentence contained "−12 dB" in its correct sense, was subsequently crossed out and rewritten and the wretched expression sneaked in while I wasn't looking and exercised squatters' rights. It does at least prove (if proof be needed) the prophetic mettle of the Editor, who, in his commentary in the September issue wrote of this abashed contributor ". . . but we would not like to say that his sense of direction is always infallible."

"VECTOR"

COHERENCE

By "CATHODE RAY"

NOW that lasers are coming more and more into the forefront of our art, the words "coherent" and "incoherent" are inevitably heard. They are usually explained quite briefly, because their meaning is thought to be simple and there are so many much more involved matters to be covered in any treatise on lasers. In the hurry to get on to these, it may be that ideas about coherence are not always as coherent as they ought to be.

Although light and radio waves are identical except for frequency (as Hertz in 1884 went to great trouble to prove by experimental demonstration), technologically they were until recently quite different. Only the late "Free Grid" reminded us occasionally that in claiming a monopoly for signalling by wireless our Postmaster General logically should control such things as the winking lights on our motor cars. Even when television came in and radio engineers had to learn about light, it wasn't as an extension of their signalling waves to still shorter wavelengths, but as a new raw material for programmes, analogous to sound. The reason for this lack of interest in light as a solution of the frequency channel congestion problem was that all known sources produced what was, from a radio engineer's point of view, mere noise. In a word, it was incoherent.

Imagine how restricted radio and allied engineering would be if oscillators had never been invented! One just can, of course, signal with radio noise sources, such as sparks, as was in fact done in the pioneering days. But it is a crude business. Instead of a nice clean continuous carrier wave to work on, itself occupying practically no frequency-band width at all, it is already randomly modulated, so occupying an unpleasantly wide band. There is difficulty enough finding sufficient communication channels without that handicap. So although ordinary light *can* be used for signalling, as for example in lighthouses, or even modulated by speech ("talking over a beam of light"), its incoherent waves are unsuitable for anything much more refined than simple on/off modulation, which is fantastically inefficient as regards frequency-band utilization.

To see why light is incoherent we shall have to consider how it comes into existence. There is no such thing as continuous light waves, even in a laser. Light comes only in small bursts, called photons, which are emitted when atoms give up some of their energy. Although the words "atomic energy" conjure up ideas of devastating bombs or, at least, vast power stations, even the light from a burning match is energy given up by its atoms. Your ability to read these printed words is due to the fact that the atoms of paper, when stimulated by light, radiate energy over the whole visible range of frequency, whereas the atoms of printers' ink hardly do so at all.

It is the essence of the quantum theory, which came in with this century and is now firmly established, that energy cannot be infinitely divided, any more than matter. The amount of energy in each of its indivisible units (quanta) is directly proportional to the frequency of the waves by which it is radiated. So the range of quantum sizes is as enormous as the range of frequency; say from 10^1 c/s for very low radio frequencies to 10^{22} for cosmic rays. But they are all very small by ordinary standards because the relating factor—Planck's constant, h— is only 6.625×10^{-34} joule-sec:

$$\delta E = hf \quad \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots \quad (1)$$

Here δE denotes a quantum or least possible energy difference at frequency f c/s. If the energy of an atom fell by 4×10^{-19} joule, for example, this energy would be radiated as a photon at a frequency of $4 \times 10^{-19}/(6.6 \times 10^{-34})$—just over 600 MMc/s, which is interpreted by the eye as blue-green light. Fig. 1 shows the relationships between frequency, wavelength, colour and photon energy.

The ways in which atoms absorb and release photons are subject to strict quantum rules. For simplicity let us consider the hydrogen atom, which consists simply of one proton as nucleus and one electron, which can be regarded as a satellite of the nucleus. Just as more rocket energy must be applied to an earth satellite to put it into a higher orbit, so extra energy must be delivered to an electron to get it into an orbit farther from the nucleus. When an earth satellite descends because of gravity, some of its energy is given up; enough, in fact, to destroy it. When an electron "falls" towards the nucleus it gives up its energy as radiation. The vital difference, however—a result of the vast difference in scale between visible objects and electrons—is that an atom's electron must always be in one of a series of orbits or states, which are quite simply related to the series of whole num-

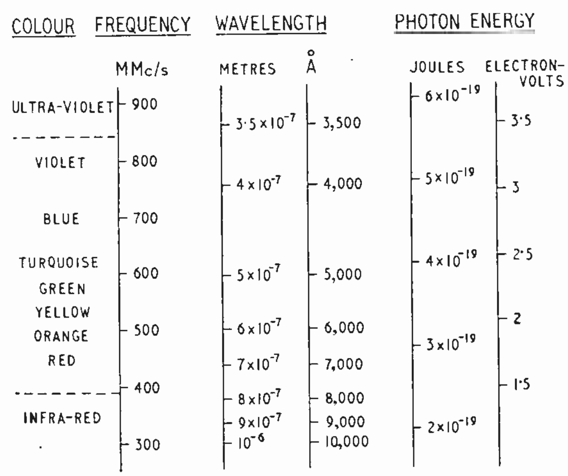


Fig. 1: Scales connecting colour, frequency, wavelength and photon energy of light. 1 angstrom (Å) = 10^{-10} metre.

bers: 1, 2, 3, etc. The number 1 corresponds to the lowest or ground state, to which the electron gravitates when it is free from all outside influences, such as heat, light or electricity. When applied, these influences are capable of imparting energy and raising the electron to one of the higher states, from which it usually drops back at once, with release of the energy.

The biggest single energy step is from 1 to 2. When converted to frequency by equation (1) it turns out to be 2,469 MMc/s, far up in the ultra-violet band. This is more energetic than any visible light, which is incapable of raising hydrogen atoms from their ground state. However, an electric discharge through a tube of rarified hydrogen can do it.

Multiple Frequencies

Although an electron normally drops back too smartly for incoming radiation to catch it in an upper state and raise it farther, the drop-back is not necessarily to the ground state. For example, an electron raised to orbit 3 may drop to 2. The energy difference between 3 and 2 is several times smaller than between 3 and 1 or even 2 and 1; its frequency is 457 MMc/s, which is red light. There are many other possible frequencies, some of them still lower, in the infra-red.

Most atoms are relatively complicated and have many more possible electron states and therefore possible frequencies of radiation due to transitions from higher to lower. Furthermore, most atoms are united with other atoms to form molecules, and any changes in molecular structure involve energy differences. Some chemical changes yield products having more internal energy than before, so energy must be supplied to bring about such changes; for example, by applying heat. Particularly interesting are the chemical changes wrought in photographic film by brief exposure to light, which supplies the needed energy. Other chemical changes, such as those commonly known as burning, yield products with lower total internal energy, and we gratefully receive the surplus as heat and light. Changes within the nuclei of atoms provide energy much more intensely, but the general principle is the same. However, let us disregard nuclear sources of light, as fortunately rare, and confine our attention to emission of light due to electronic re-arrangements.

Atoms or molecules that are widely spaced, as in low-pressure gases, radiate on spot frequencies within the light band. But just as the natural frequency of a number of identical resonators spreads out into a band if the resonators are closely coupled together, so if atoms which have the same radiating frequency are brought close together (as in solids or even high-pressure gases) this frequency spreads out into a band. Metal filaments supplied electrically with heat energy provide light over the whole visible band, though more strongly at the red end, as photographers know. The sun, being at a higher temperature, fills out the blue end much better, giving a whiter light. White paper, exposed to white light, responds at practically all frequencies, so we get white re-radiation. Coloured materials are those whose molecules respond selectively to light of particular frequencies, like tuned circuits emphasizing particular frequencies present in "white noise."

From the foregoing brief summary of the science of light one might conclude that provided atoms were kept well apart, as in low-pressure gas, exciting them would yield light at exactly one or more spot frequencies, just as an unmodulated signal generator provides pure single-frequency radio-frequency waves. Even the fairly high-pressure sodium vapour discharge tubes used in many areas for street lighting give light on so narrow a wave-band that it is seen as the characteristic yellow that so upsets complexions and other colour schemes. Although their bandwidth is certainly small in relation to their "carrier" frequency, by radio standards it is enormous. With the utmost care to confine energy changes to a particular transition in atoms too far apart to influence one another's radiation frequencies, the narrowest bandwidth is of the order of 400 Mc/s!

Even in proportion to the "carrier" frequency, this is at least several orders of magnitude worse than can be done with an unmodulated radio signal generator. One reason is the "granular" nature of light emission, by short bursts instead of continuously, as we have already noted. This is, in effect, modulation, with its inevitable sidebands. The

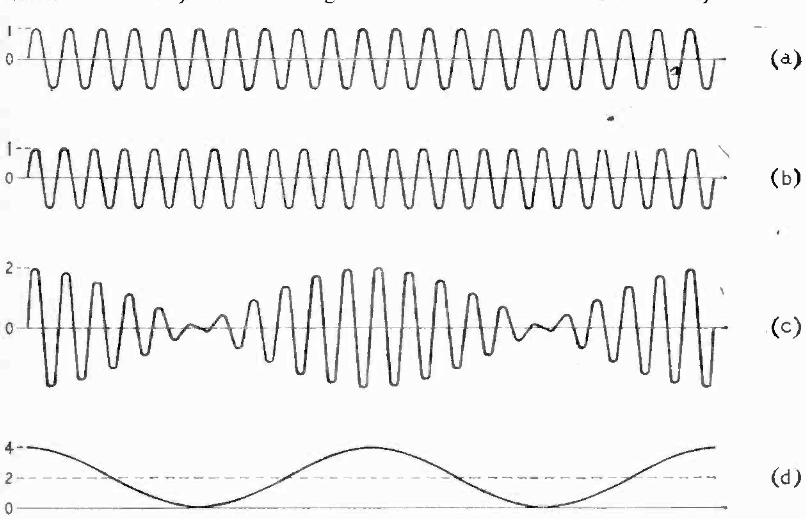


Fig. 2. Two wave-trains (a) and (b), of equal amplitude but slightly different frequency, add together to give a periodically varying ('beating') wave-train, (c). Its power varies as at (d) to give an average twice that of (a) or (b) alone.

shorter the wave-trains, the wider the band they occupy. Denoting the frequency spread by δf and the duration of each pulse of radiation by t, we have

$$\delta f \simeq \frac{1}{t} \quad \ldots\ldots\ldots\ldots\ldots \quad (2)$$

If 400 Mc/s is taken as a figure for δf, t is of the order of $\frac{1}{4} \times 10^{-8}$ sec, so each photon comprises something like one million cycles (taking f as 400 MMc/s), totalling 75 cm. in a vacuum. In fact, however, only a minority of the observed band-spread is due to this cause, and the wave-trains are much longer—of the order of 3×10^{-8} sec., equal to more than 10^7 cycles, occupying perhaps 10 metres.

Doppler Shift

The major cause of band-spread is Doppler effect. We might not have expected this to come into the present subject, unless we happened to remember that photons are discharged from bases that are perpetually on the move owing to heat. In fact, one could say that this motion *is* heat. The drop in audio frequency due to the change in relative velocity between source and hearer when a racing car passes is familiar enough, so I hope there is not need to explain Doppler effect further. The higher the temperature, the faster the atoms or molecules move, and the greater are the Doppler shifts of frequency (above and below normal) in the light waves radiated. So I should mention that the 400 Mc/s band width cited above is typical at ordinary room temperature.

Because the wave-trains of light are emitted at completely random intervals, the phase relationships between them are random. Every momentary phase relationship between any two is equally likely. And because of the Doppler frequency differences the phase relationships go through every possible value continuously with time, as shown in Fig. 2, where (a) and (b) represent two simultaneous trains. These alternately reinforce and cancel one another, giving the combined result shown at (c). When there is full reinforcement (phase difference, 0°) the amplitude is twice that of one alone; when cancellation is complete (phase difference, 180°), zero. Halfway between, the difference is 90° and the resultant amplitude therefore $\sqrt{2}$ times of either alone. The average amplitude is obviously less than double that of each, but what counts is the *power*, which is proportional to the square of the amplitude. In the same three phase relationships it is, respectively, 4, 0 and 2; and the average (d) is 2, compared with 1 for a single wave-train. With n atoms or molecules the total power is of course n times that from one alone.

Fig. 2 shows how two wave-trains of slightly unequal frequency, which are observed at a fixed point, alternatively reinforce and cancel one another progressively with *time*. A very similar phenomenon called wave interference occurs in *space*. (This should not be confused with interference in its commonly used sense of intrusions spoiling one's broadcast programme.) Here the two wave-trains have the same frequency, and the phase differences are caused by differences in distance from the sources of the waves. And so one gets interference patterns, such as you can make by dropping two pebbles a short distance apart into still water. We radio people are familiar with this, because it is employed to obtain directional radiation, by assembling a number of wave sources (unit aerials) into an array devised so that their separate outputs reinforce one another in the desired direction and (as nearly as practicable) cancel everywhere else.

When we study light we are told that its waves, too, produce interference effects, which are often explained in the same way by showing two trains of sine waves interacting. Photographs of interference patterns are provided as evidence. Since light is to us a subsidiary subject that we have to pass at a modest level in order to be allowed to get on to radio engineering or the like, in our haste we can easily get the impression that interference of light is the same as interference of radio waves except for frequency, overlooking the fact that sources of light are incoherent. Sine waves in the diagrams are an excessive simplification, and consequently the phenomenon of light interference is much more restricted than in radio.

Once it has been grasped that light is incoherent, the wonder is not so much that interference ever

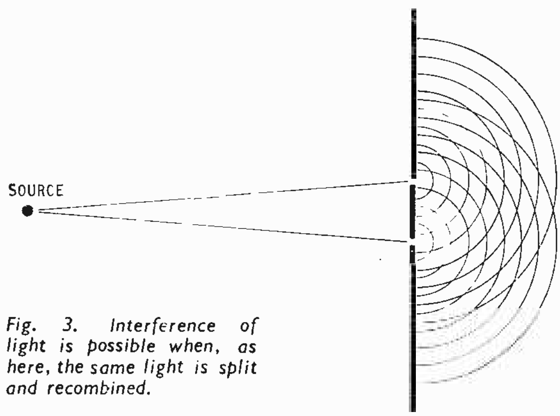


Fig. 3. Interference of light is possible when, as here, the same light is split and recombined.

fails as that it is ever possible. One might think that random noise, with its completely irregular and non-repeating waveform, could not show regular progressive phase displacements. And one would be right. But even white light is confined to a frequency band of about one octave (2 to 1 ratio), enormous though that may be to us in cycles per second. We have already seen that a band-width of 400 Mc/s (which in optics would be described as very monochromatic light) is so narrow at such frequencies that it does approximate to a sine-wave train over a considerable number of wavelengths. So light from a single source can interfere with itself by what in radio we call multi-path propagation. If the light is made to pass through two very small holes equidistant from the source, as in Fig. 3, each of the holes acts as a source of light having identical waveform, and interference patterns can be obtained.

Even so-called monochromatic light doesn't maintain its frequency constant enough to give clear interference effects with path differences exceeding a moderate number of centimetres, corresponding to time differences of the order of 10^{-10} sec. White light is obviously much more restricted still, but light of its constituent colours does interfere over a few wavelengths to produce the well-known colour fringes that plague photographic transparency enthusiasts.

Light from independent sources, including different points on the same source, doesn't maintain even an approximately constant phase relationship at any one point, so interference effects are absent—a fact that is evidence of its incoherent nature.

White light, as we have just seen, is like "white noise" within an octave band, having no definite frequency; or all frequencies in that band, if you prefer to look at it that way. In a high-Q tuned circuit, random noise is selectively amplified, giving it a predominant frequency, just as white light can be made monochromatic (of one particular colour and frequency) by passing it through a suitable filter. By amplifying "monochromatic" radio noise and feeding a sufficient fraction of it back to the tuned circuit, the Q of the circuit can be made effectively infinite and we get a continuous wave-train of constant frequency. We call this device an oscillator. It should not be forgotten that oscillators work by amplifying the noise that inevitably exists in circuits.

Amplification and Oscillation

Until the invention of the laser there was no device for amplifying incoherent or "noise" light sufficiently to obtain coherent oscillation, even for brief periods of time. Even now, most of the several kinds of laser can yield only brief bursts of oscillation. I won't go into details of lasers, since these have been previously described*. It will suit our present purpose best to consider only the gas laser, which is capable of continuous output and phenomenally constant frequency.

In essence it consists of a tube of rarified gases, which are kept in a state of continuous excitation by a v.h.f. voltage applied between two electrodes. The effect is to put large numbers of the gas molecules into a state described as metastable; that is, one that lasts for what compared with a light-frequency cycle is a very long time (of the order of a millisecond) before dropping back and emitting light. However, those in this state drop back instantly if they are stimulated by light of exactly the frequency they emit. "Instantly" means instantly, so literally that the stimulated light falls into phase with the stimulating light. If you imagine a light wave, due to one molecule, sweeping along the tube, as it goes it recruits vast numbers of others which build up to a mighty army, all in step.

The ends of the tubes are designed to reflect something like 98% of the light, which about-turns and gathers more reinforcements from molecules that have been excited in the meantime. There is another reflection at the far end, which is spaced an exact number of half-wavelengths from the first, so that successive sweeps are all in phase. Light of that particular frequency and travelling along the tube is therefore strongly built up; light of other frequencies tends to cancel out, and light in other directions passes out of the tube at once instead of building up by multiple reflection.

Because reflection is not complete, some of the light emerges at one or both ends, and as the emissions of all the gas molecules are in phase, both in time and space, the output is coherent. That is to say, the separate waveforms not only coincide in time so as to reinforce one another; they are in the same phase along their whole front.

Although the similarity of this device to our electronic oscillators may not be obvious, both rely on feedback which is in such a phase as to reinforce an already existing oscillation, preventing it from dying out. And just as insufficient feedback coupling or excessive circuit losses prevent an oscillator from oscillating, so a laser will refuse to lase unless the reflected light waves are in the right phase and of sufficient intensity, and there are enough gas molecules in the metastable state.

As I said, gas lasers are by far the best for coherence, and with sufficient care in construction and operation can be made to provide light within a frequency band of about 1 c/s. Since the frequency is 4.74×10^{14} c/s, this means within better than 1 in 10^{14}! I have heard a recognizable beat note obtained by mixing the direct output from such a laser with a ray reflected from a surface moving towards or away from the source. To obtain a note of 1,000 c/s the relative motion needs to be at the rate of 500 wavelengths per second; that is to say, about one third of a millimetre per second! No wonder the beat note sounded slightly jittery, when every 0.0003 mm/s variation in this rate is enough to shift the frequency 1 c/s.

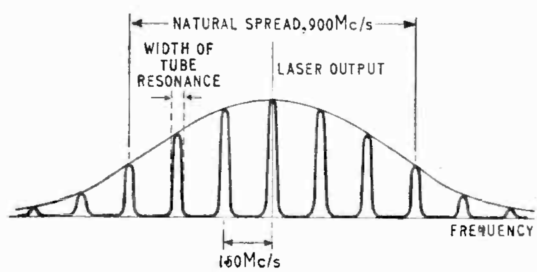
One can go on from this to deduce quite easily that the distance apart of the reflecting ends of the laser tube (normally 1 metre) has only to vary 0.0003 mm, or about one part in 3 million, to shift the output frequency 1 c/s. Hence the care in construction and operation mentioned above. For one thing, the temperature of the apparatus obviously has to be kept extremely constant. And the reflecting surfaces must be flat and parallel with truly optical accuracy.

It is interesting to consider the various steps by which the gas laser reduces the frequency bandwidth to such narrow limits as can be achieved.

First, the gas is at low pressure, so that broadening due to interaction of molecules is negligible. The main cause is Doppler effect caused by the molecules flying about at high speed. Owing to this and the quantum effect, their output is distributed over a band about 450 Mc/s each side of centre (which is 474 MMc/s).

Now, as we saw, the number of half-wavelengths between the reflecting surfaces must be a whole number if there is to be reinforcement by the reflected light. The wavelength at 474 MMc/s being about 6.33×10^{-7} metre, the number of half-wavelengths in 1 metre is about 3.16 million, so a shift

Fig. 4. Analysis of the frequency band of a gas laser.



*For example, by Aubrey Harris in *Wireless World*, Aug. and Sept. 1963.

of one half-wavelength is equivalent to a frequency shift of $(4.74 \times ^{14})/(3.16 \times 10^6)$, or about 150 Mc/s. So there are several different wavelengths and frequencies, within the 900 Mc/s band generated by the gas molecules, that fulfil the regenerative requirement. They are represented by the peaks in Fig. 4.

These peaks represent the frequencies at which there is optical end-to-end resonance of the laser tube. Their very appreciable width indicates that the Q, although very high by radio standards, is not infinite, so without the regeneration of laser action there would not be continuous oscillation. This takes place at the peaks, of course, and a 1 c/s width (or even the considerably wider band of a less preçise gas laser) is much too narrow to show on the same frequency scale. Oscillation can take place at more than one of these peaks, but there is unlikely to be sufficient overall gain for it to do so at the smaller ones.

Sensational figures have been published to the effect that the intensity of the light emitted from lasers is millions of times greater than that from the surface of the sun, which we imagine to be itself fantastic. And so it is, if radiation at all frequencies is counted, or even visible light frequencies only. But if we select the minute bandwidth covered by a laser, the sun's radiation amounts to only microwatts per sq. cm., or less! Obviously the total output of a laser cannot be large, because the input (on a continuous rating) is very modest, and the efficiency of gas lasers at least is very low because of their high selectivity. The intensity of light from lasers is remarkable only for its concentration within a phenomenally narrow band, which makes it quite different from ordinary light.

There is its directivity, for example. The beam from a gas laser diverges at an angle of about 0.5 minute, which means that at a distance of a mile it has spread out to the extent of only about 9 inches. In terms of the so much lower frequencies we use even in microwave radio, a laser is a beam aerial of enormous gain, able to project to distant points an intensity comparable with that at the generator. The interference effects which are relied on for beam aerial arrays are effective in lasers without the severe limitations that apply to incoherent light.

Incidentally, lest anyone continue in the popular error that the reputation of Einstein is built on relativity alone, it is worth mentioning that among his many prophetic achievements was the working out of the theory of stimulated light emission more than 40 years before the laser was invented!

CONGRESS ON MICROWAVE TUBES

PARIS, 14th-18th SEPTEMBER, 1964

THE 5th International Congress on Microwave Tubes was well attended, some 135 papers were presented. The papers were divided into three groups and continuous sessions were held in all three halls throughout the week. Clashes of interest were inevitable and frequent hurried walks between the quite widely separated lecture halls were an essential part of the conscientious delegate's day. In the circumstances it was rather unfortunate that the organizers decided not to do a pre-print of the papers.

However, the Congress as a whole was certainly worth while attending, for a large number of quite important papers was presented. As seems to be usual at conferences in Paris continuous translation facilities were available at the Ecole Polytechnique, and French, English and German speakers could present their papers in their own language knowing that their words and, what is perhaps even more important, the full contents of questions and answers at the end of each session would be immediately translated into the two other official languages of the conference.

The three groups into which the papers were divided were entitled "Low Power," "High Power" and "Advanced Studies." Each group was then further divided into a number of sub-groups.

Low Power: Forty-three papers divided into eight groups were devoted to low power microwave tubes. Eleven papers were given over to the subject of low noise tubes. The characteristics of electron beams and electron guns weer discussed in four papers. Eight papers on the subject of travelling wave tubes were presented and cyclotrons were discussed in another five papers. Klystron oscillators formed the subject of four papers and the remaining eleven papers were divided between magnetrons, carcinotrons and new structures.

K. Milne of the G.E.C. Hirst Research Centre described the problems met in the design of a wide-band travelling-wave tube for operation between 200 and 1000 Mc/s. The t.w.t. is attractive as a wide-band amplifier at these frequencies and if the helix voltage is sufficiently low a reasonable tube length can be obtained. Measurements of the dispersion of the interaction helix have shown that this and, not what might have been expected, the coupled helix couplers will be the limiting factor in obtaining wide-band performance. The tube designed by K. Milne has a bandwidth in excess of two octaves (from 250 Mc/s to 1000 Mc/s) and a noise figure of better than 7 dB. Typical values of gain are in excess of 20 dB and the saturated output powers are about 1 mW.

A transverse travelling-wave tube which holds promise of high efficiency has been operated at S-band at the Services Electronic Research Laboratories at Harlow. J. Carroll of that establishment said that the interaction is obtained between a slow cyclotron wave on a 7 kV beam and a transverse field slow wave on a meander line circuit. The basic theoretical efficiency of the amplifying mechanism is 75% in the tube that has been built. This is one reason for the high efficiency but, in addition, the electron beam carrying the cyclotron wave has no velocity modulation and can be expected to be used successfully with a depressed collector. Suitable collector depression can also recover the rotational power left on the cyclotron wave and increase further the theoretical limiting efficiency.

An interesting oscillator which utilizes a superconducting cavity, was described by Nguyen Tuong Viet of the Institut d'Electronique, Orsay. Such a cavity has a Q factor which exceeds 7 millions. When it is placed in the feedback path of a travelling-wave tube amplifier which functions as an oscillator the microwave signal frequency obtained is very stable and can easily be set to the resonant frequency of the cavity. The author suggested that this could be used as a high Q factor measuring instrument.

Experimental techniques which have been devised at a wavelength of 1.8 mm for comparison of the relative advantages of confocal Fabry-Perot resonators to plane-parallel ones were described by A. P. Sheppard and J. W. Dees of the U.S. Army Research Office, Durham. They said that cavity resonators operating above 100 Gc/s pose severe problems due to the small physical size, and the Fabry-Perot interferometer offers an attractive possibility for overcoming these difficulties. With direct waveguide coupling into a semi-confocal Fabry-Perot interferometer of silvered plates a resonator Q of 90,000 has been observed.

Quite a number of interesting papers were devoted to carcinotrons, and this is a field in which the French have done a considerable amount of work, and in which they appear to be the leaders—at least in Europe.

High Power: Fifty papers divided into five groups were devoted to high power microwave sources. Seven French authors, all of whom were with C.S.F. presented papers on parasitic oscillations in high-power microwave tubes. Five papers were devoted to high-power travelling-wave tubes and electron guns and beams for high-power tubes were described in eight papers. No fewer than fifteen papers were devoted to high-power klystrons and an equal number of papers were given over to cross-field devices.

C. Enderby of the General Electric Co., Palo Alto, California described a new family of travelling-wave amplifiers which use ring-plane circuits which make possible high-power amplification at millimetre wavelengths. They are a variation of planar ladders in which the ladders are bent and placed together to form the ring-plane geometry. They have the high impedance and high heat-handling capability of ladder circuits and, in addition, they have a geometry which allows a solid cylindrical beam to utilize these advantages efficiently. An experimental tube constructed by Enderby produced 43 kW of power at 33 Gc/s. The low level gain of the tube was 20 dB and the tube operated at a voltage from 15 to 100 kV. The power output and efficiency ranged from 100 W at 4% to 43 kW at 12%. Instantaneous bandwidth was 300 Mc/s (1%).

A review of some recent advantages in high power klystron amplifiers was given by A. J. Prommer of Litton Electron Tube Corporation, San Carlos, California. He described a tube which, he said, was representative of the trend to higher average powers. This tube provided 5 MW peak power output and 300 kW average power output at L band. He said that the major problem areas on this programme were the generation, focusing and collection of an electron beam with an average power in excess of 1 MW, transmission of the 300 kW average power through the output window and beam control by a non-intercepting modulating anode at voltages up to 110 kV. In the field of high peak power development, he discussed the design of two 30 MW klystrons. Both klystrons are rated at 30 MW peak power and 30 kW average power with a 10-microsecond r.f. pulse length. One klystron works in the u.h.f. band and the other in L band. As an example of a high-mu modulating anode design, he described development work on a 10-MW hollow beam klystron. This klystron uses a magnetron injection gun, and utilization of the high gun perveance of this design has led to a low enough modulating anode voltage to allow the generation of beam pulses in the nanosecond region.

Advanced Studies: Forty-six papers divided into six sections were devoted to advanced work. Six papers were given over to electron optics. The generation, amplification and propagation of waves in plasmas were discussed in eight papers. Seven papers covered the field of electronic emission and particular emphasis was laid on laser applications. Thirteen papers were presented on the subject of interactions between electron beams and plasmas. Six papers were devoted to such subjects as electron cyclotron masers, tunnel lasers and optical masers using the negative glow of cold cathode discharges. The final six papers were devoted to new systems of generation.

A way round the difficult problem of designing coupling systems for millimetre beam plasma amplifiers was described by J. Allison and G. S. Kino of Stanford University. They used two counter-streaming electron beams in a caesium plasma. A system such as this has been suggested as the mechanism for certain types of radiation from the sun's corona. Allison and Kino constructed their tube to test this hypothesis and found that it also provided a useful amount of radiation at the second harmonic of the plasma resonant frequency. The plasma radiates directly and there is no need for any coupling device.

D. J. Blattner and his co-workers at R.C.A., Princeton, have developed an improved version of the travelling-wave photo-tube detector. They have inserted a transmission type photo-multiplier between the photocathode and the travelling-wave tube part of the standard detector. Their experimental tube was designed for a centre frequency of 1.5 Gc/s and has a bandwidth of 1.0 Gc/s. This new type of laser light detector combines the high noise-free gain of a secondary-emission multiplier with the very large bandwidth of a microwave helix.

M. A. Allen and C. S. Biechler of Microwave Associates Inc., Burlington, Mass., described experiments on beam-plasma amplifiers which gave output powers several orders of magnitude higher than any reported elsewhere. To date, output powers over 10 kW have been achieved at 3 Gc/s with efficiencies greater than 25%. Interaction bandwidths of 20% have been observed at gains greater than 30 dB. The tube uses a 20 kV, 2.8 A pulsed electron beam in conjunction with a hot cathode plasma injection gun discharge. An axial magnetic field is used to confine the plasma and focus the beam. The possibilities of increasing the power of this type of device to the megawatt average power range were discussed by the authors.

A very interesting way of obtaining the output from a semiconductor acoustic* amplifier was described by Dr. E. A. Ash of University College, London. He pointed out that one of the main obstacles that hinders the practical exploitation of the semiconductor acoustic amplifier is the difficulty of coupling to the drift-electron, acoustic-wave system. At microwave frequencies, particularly, coupling losses tend to be very large, even when using narrow-band resonant cavities. In his paper he described an output system in which the mechanical displacement in the semiconductors is used directly to modulate a beam of light. The end of the crystal forms one mirror of a two-mirror optical cavity. An unmodulated laser beam excites this cavity and the movement of the mirror results in amplitude modulation of the optical signal. This amplitude modulation can, if desired, be detected; but as Dr. Ash pointed out, there are many systems where an output in the form of an amplitude modulated optical signal would be a great advantage.

Conclusion: Details of only a few of the many interesting papers at this Conference are given here, and it would seem very well worth while obtaining the full volume of papers when they become available. (Dunod Editeur, 92, rue Bonaparte, Paris 6e.) At 230 francs—nearly £17—the price may seem a little high, but, for this, one would receive a very good review of current developments in the whole of the microwave field, including plasmas and quite a few papers on lasers.

* Involving vibrational energy, not necessarily audio—Ed.

Elements of Transistor Pulse Circuits

10.—COUNTER/TIMERS (FREQUENCY METERS)

By T. D. TOWERS,* M.B.E.

NOWADAYS, if you work in an electronics laboratory, you are expected to have some knowledge of counter-type frequency measuring instruments. These are steadily moving into the status of essential laboratory equipment—alongside the multimeter, the signal generator and the oscilloscope. Many basic pulse circuits discussed previously find a place in such counters, and looking at the working of a counter conveniently shows them in action.

Until the invention of the electronic counter, precise frequency measurements were carried out by comparing the unknown frequency with some accurately known variable standard, zero-beating the standard against the unknown. Heterodyne frequency measurements of this type tended to be slow, difficult, ambiguous and expensive. The electronic counter has changed this. It has placed within the reach of the average small laboratory equipment which is capable of measuring frequencies accurately to say seven or eight significant figures (e.g. to 1 cycle in 10 Mc/s), and yet of giving these precise results almost instantaneously, without ambiguity and with almost no manipulation of apparatus.

The basic principle of the electronic counter-timer is simply to count the number of cycles of an unknown frequency that occur in unit time. Fig. 91 illustrates this in block diagram form. To measure a frequency, the signal is fed into a gate as shown in Fig. 91(a). This gate is controlled by a timebase in the instrument, which arranges to open the gate for a precisely determined length of time, say 1 second. During the second that the gate is open, the input signal passes through and records on the counter the number of cycles passed in that second. The gate then closes and the number of cycles is left recorded and visually displayed on the counter readout. Thus we can read directly on the counter the frequency of the input signal with the potential accuracy of the number of digits displayed. For example, with six-digit displays it is possible to count up to 999,999.

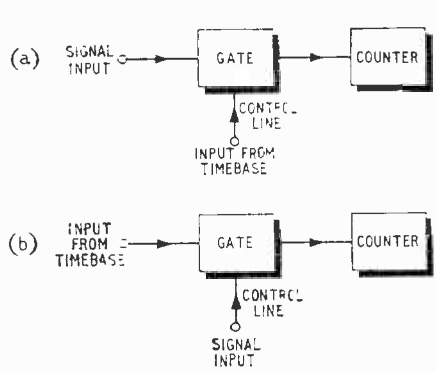


Fig. 91. Basic principle of counter/timer. (a) Measuring number or frequency (b) Measuring time or period.

The same equipment can be used also to measure time or period, rather than number or frequency. To do this, it is arranged as in Fig. 91(b). Here the signal input is used to control the opening and shutting of the gate, and the input from the timebase is fed through to the counter. The signal input opens the gate for the duration of one period of the signal frequency. If the timebase is operating at say 1 Mc/s frequency, it will be supplying one pulse every microsecond into the gate. Thus there will be recorded on the counter the number of microsecond pulses passing through in one cycle of the signal input; the counter will then read in microseconds the period of the signal frequency.

We have illustrated the principle of the counter/timer in terms of a periodic signal input. However, it can be used equally well to measure the total number of impulses coming in on the signal input line (Fig. 91(a), even though they do not recur at a regular periodic rate. Similarly in the case of Fig. 91(b) the equipment can be used to measure the total time between two isolated events which are fed into the signal line, one opening the gate and the other closing it.

Typical Electronic Counter/Timer

Counter-timers (or digital frequency meters as they are sometimes called) in their commercial versions show many detailed differences of circuitry, but stripped of unessentials they all reduce in essence to the arrangement shown in block diagram form in Fig. 92. After the frequency to be measured has been fed into the input, the *input pulse shaper* amplifies the signal and converts the waveshape into the standard form of a square wave with very fast rise and fall times, suitable for driving an electronic counter. The shaped pulses are permitted to pass through the *signal gate* for such time as it is opened by the *gate control unit*. The whole system has been reset to zero before the beginning of the gate-open period so that the number of pulses passing through the signal gate are stored in total in the *counter*. The counter contents are recorded in visual numerical form in the *readout display*.

Thus far the gate control unit has performed two functions. It has reset the counter to zero, and opened the gate for the required gate period. The control of the gate period is effected from a timebase. This timebase comprises (a) *a master oscillator* which

*Newmarket Transistors Ltd.

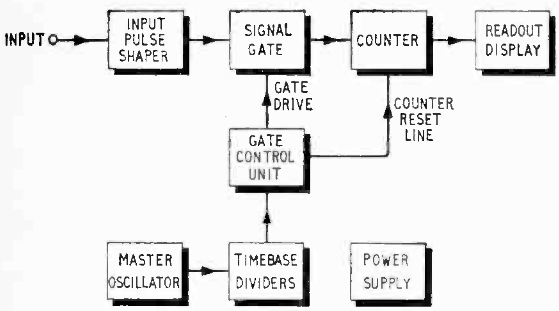


Fig. 92. Simplified diagram of typical electronic counter/timer.

is a stable crystal-controlled circuit and (b) a set of decade counter "times-ten" *dividers*, usually referred to as DCUs. The master oscillator itself is sinusoidal, but it includes a buffer squaring circuit which provides a timebase square wave of frequency f. The timebase DCUs provide derived square-wave outputs at frequencies $f/10$, $f/100$, $f/1{,}000$, etc. The operator selects the desired timebase frequency from those available and switches this manually into the gate control unit.

The *power supply* shown separately in Fig. 92 is the unit which supplies all the d.c. rail voltages to the other sections of the counter-timer. As it has no particular features of interest in the pulse circuitry aspect, we will not consider its detailed design. It is worth noting, however, that with the modern trend to transportable instruments for use independently of mains supplies, commercial units nowadays show a tendency to end up with a 12V supply which can be interchangeably provided by a car battery (for mobile work) or derived from the mains (for fixed station operation).

In the discussions of the previous paragraphs, we have considered that the gate has opened for only a single period of the timebase and closed again. However, if the unknown frequency varies with time, a single sample of this sort will not be satisfactory. The gate control unit has therefore to provide for successive repetitive sampling of the input signal so as to correct continuously the readout display of the frequency. This will be discussed in greater detail later.

Counter

The most important and probably the least understood section of a frequency meter is the *counter* itself. This usually comprises a counting chain of decade counter units arranged in cascade so that each one divides the output of the previous one by ten. Each DCU is a cascade of four binary counters. You can follow the build-up from the basic binaries to the complete counter in Fig. 93.

A typical basic *binary* element is illustrated in Fig. 93(a). This type of circuit has been discussed in some detail in previous articles in this series. It is a bistable multivibrator of the Eccles-Jordan type. Pulses fed into the input line are steered alternately to the two transistor bases in turn and cause the circuit to switch from the condition where the right-hand transistor is conducting to cut-off and back again. We take the condition with the right-hand transistor conducting as the "set" state of the binary and where it is cut off as the "unset" state. Every time the binary is driven to the set state a positive-going pulse appears at the output, and to the unset state a negative-going one. The reset line shown is normally shorted to the earth line, but, if it is disconnected from earth, the right-hand transistor is driven hard on through the 470 ohms and 6.8 kΩ connected from the negative rail to its base. Thus to reset the binary, all that is necessary is to open circuit the connection between the reset line and earth. The actual example of a binary given in Fig. 93(a) is capable of switching reliably at not less than 1 Mc/s.

The *decade counter unit* shown in diagrammatic form in Fig. 93(b) is an arrangement of four binaries with two feedback loops. The output of each binary feeds into the input of the next one. Each binary divides by two, so that the cascade of four would without feedback divide by sixteen, i.e., would give one output pulse for each sixteen that are fed into the input. To divide by ten, two feedback loops are inserted which cause the decade counter unit to skip altogether six counts. For every ten input pulses then, the DCU gives one output pulse. There are many ways of arranging feedback loops to convert a divide-by-sixteen into a divide-by-ten. The method of Fig. 93(b) is only one of the many possibilities. The output terminals marked "$\bar{1}$," "1," "$\bar{2}$," "2," etc., in Fig. 96(b) are taken from the collectors of the respective binaries. They are used to drive a separate *readout* display which shows numerically by some form of lamp or illuminated figure the actual number of counts stored in the DCU. These may vary from 0 to 9. When the contents of the DCU reach 10, it gives out a pulse and reads 0 again to carry on counting the next ten pulses.

The complete *counter* unit is made up of a cascaded series of DCUs as shown in Fig. 93(c), each one feeding its output into the input of the next one. The same number of DCUs are required as the number of digits required in the readout.

Readout Display

The *readout display* unit shown in the layout of Fig. 92 as fed from the counter can take many forms in practice, varying from a cheap meter for each digit place, up to very complex electroluminescent elements. Much ingenuity has been consumed in trying to produce a cheap reliable readout device, but with no great success so far. Most of the commercial readout units are expensive in themselves and also require an expensive decoding network to transform the binary information from the counters into a decimal display. Because there is little standardization in this field yet, we will not consider further here the decoder circuits, etc., required in a practical instrument.

Input Pulse Shaper

Logically the next element to be considered in the layout of Fig. 92 is the *input pulse shaper*. Fig. 94 gives a detailed circuit of a typical pulse shaper capable of operation up to 1 Mc/s and over. Here the input signal passes through a 4.7 kΩ resistor attenuator with a 10pF compensating shunt capacitor

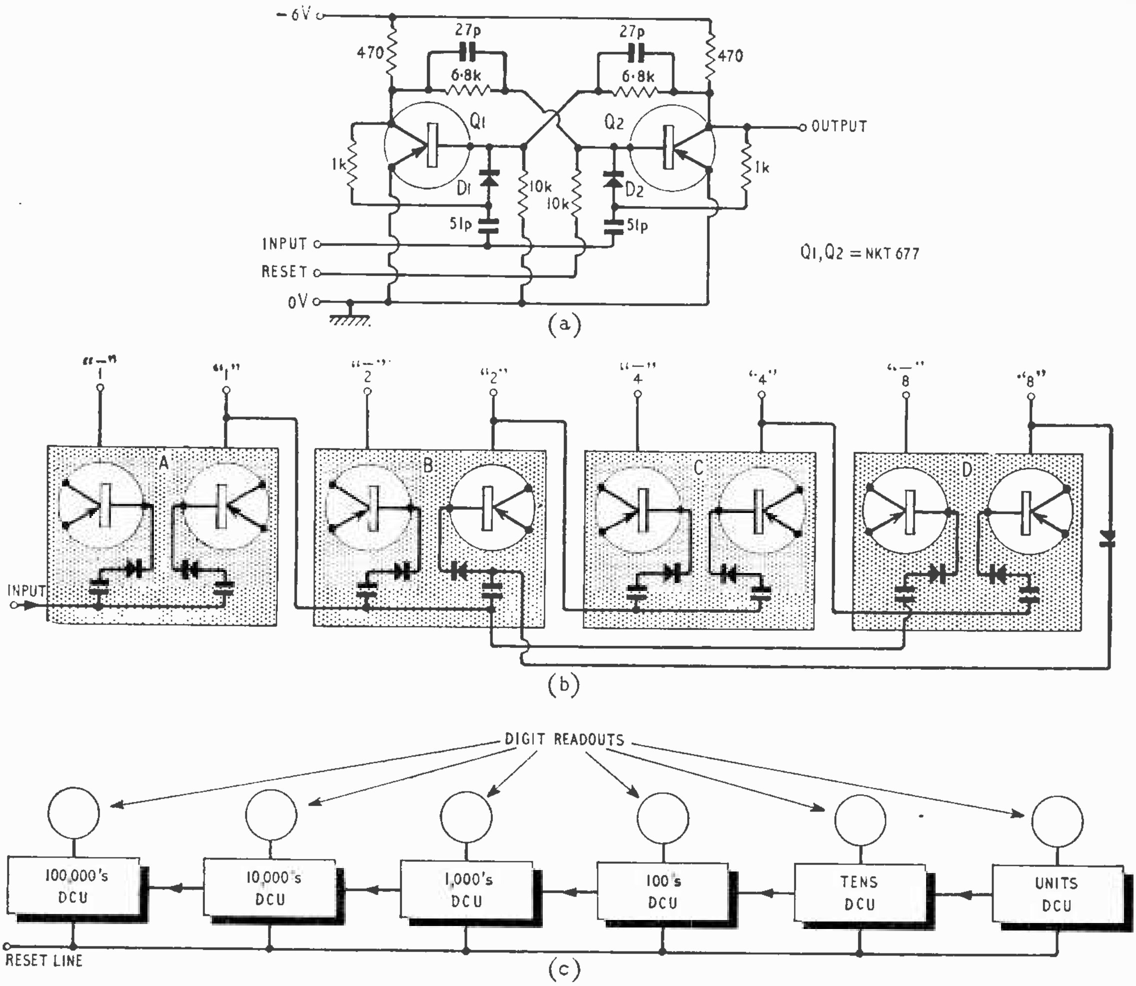


Fig. 93. Counter unit build-up. (a) Typical 1Mc/s binary (b) Four binaries connected as decade counter unit DCU (c) Six DCU's cascaded as six-digit counter.

for operation up to the highest frequency handled. At point A the signal is rectified by the diode D1 and the base-emitter diode of transistor Q1. After amplification through Q1, the signal reappears rectified at the collector as shown, i.e., in the form of a semi-sinusoidal pulse for each cycle of the input. Applied to the input of the Schmitt trigger Q2, Q3, the signal finally appears at the output of Q3 as a steep-sided square-wave pulse suitable for driving the counter circuits.

Bias to the base of the first transistor Q1 is provided from the 50 kΩ potentiometer across the −6 V, +3 V lines. By varying the setting of the slider on this potentiometer, it is possible to set the bias point of Q1 from full cut-off to full conduction. By this means we can adjust the setting of Q1 for optimum trigger sensitivity for any amplitude of input signal. In most commercial equipments the pulse shaper circuit is like the typical example given, i.e., a rectifying amplifier followed by a squaring Schmitt trigger.

Signal Gate

The *signal gate* shown in block form in Fig. 92 as controlling the transfer of pulses from the input pulse shaper to the counter can take many forms, depend-

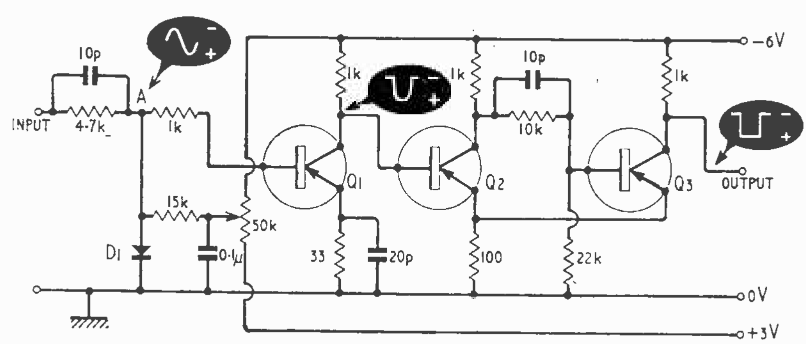
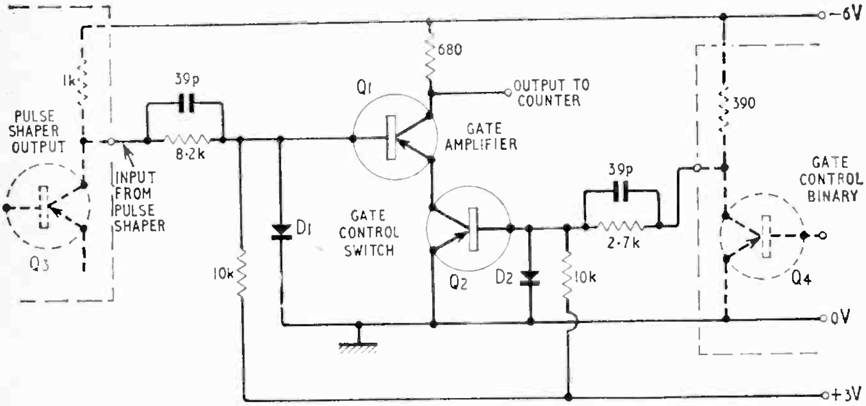
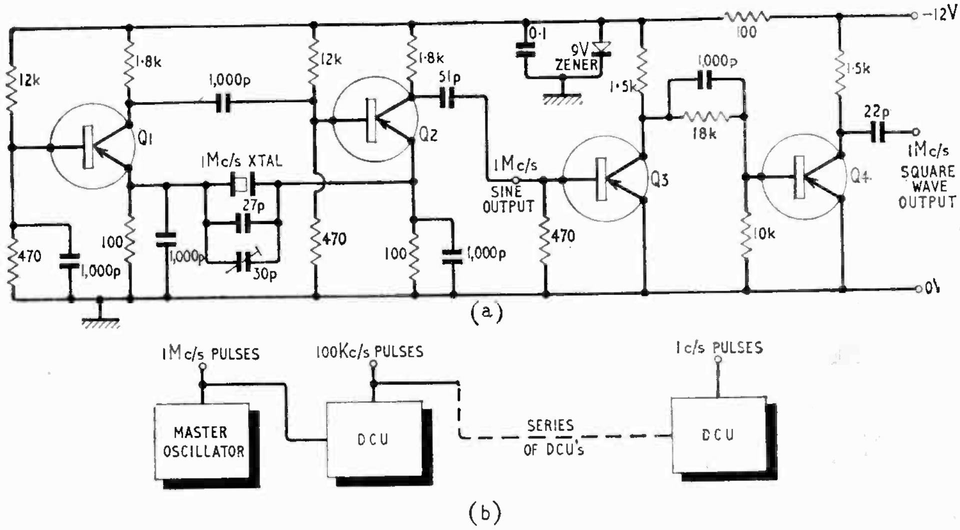


Fig. 94. Input pulse shaper (typical 1Mc/s circuit).



Left:—Fig. 95. Signal gate (typical 1Mc/s circuit).

Below:—Fig. 96. Timebase master oscillator and dividers (a) Typical 1Mc/s high stability master oscillator (b) DCU's used to provide range of timebase frequencies by repeated tens-division from 1Mc/s.



ing on the type of gate circuit favoured by the designer. Fig. 95 shows a typical circuit for this position. The output from the pulse shaper is fed via an 8.2 kΩ isolating resistor into the base of the transistor Q1. As before, the input signal is rectified by the diode D1 and the base-emitter diode of Q1. This rectified input waveform provides an amplified output unidirectional pulse at the collector of Q1. This gate amplifier can operate as a common emitter amplifier and thus pass a pulse through only if its emitter is effectively short-circuited to deck by the control gate switch transistor Q2 being bottomed (full on with a relatively low resistance from its collector to emitter). When the base of Q2 is negative the transistor is driven hard on and the gate amplifier can operate, i.e., the gate is open. Conversely, when the base of Q2 is positive, the transistor is cut off, and its collector presents a high impedance to the collector of Q1. This means that the gate is then closed.

The control voltage for the base of Q2 can be provided from the collector of the gate-control binary Q4, through a dropping resistor of 2.7kΩ and a 10kΩ resistor to the positive rail. When the gate control binary Q4 is cut off its collector rises close to −6V and the base of Q2 connected to the centre point of 2.7kΩ, 10kΩ attenuator from −6V to +3V is at a negative potential with respect to deck. Q2 is then switched hard on and the gate is open. Conversely it can be shown that, when the gate control binary Q4 is switched hard on, the collector is virtually at deck potential, and the base of Q2 is therefore positive. This cuts off Q2 and closes the gate.

Master Control Oscillator

Considering the elements in Fig. 92 further, the next one to look at is the *master oscillator* unit. As mentioned earlier this is usually a crystal-controlled oscillator feeding into a squaring circuit unit. A typical 1Mc/s master oscillator circuit is given in Fig. 96(a). Q1 and Q2 form a Butler crystal oscillator circuit with positive feedback from the emitter of Q2 to the emitter of Q1 via the tuned circuit comprising the 1Mc/s crystal shunted by the fixed capacitance of 27pF and the variable 30pF. The variable capacitor is used for fine adjustment of the oscillator frequency. The 1000pF capacitors connected from the emitters of Q1 and Q2 to deck may look like decoupling capacitors across the 100Ω emitter resistors which would prevent the circuit oscillating. However, the decoupling is not complete because at 1Mc/s 1000pF has 150Ω impedance and the resultant impedances in the emitter circuits are not less than 60Ω each. The sinusoidal output of Q2 is taken via a 51pF into the base of the over-

(*Continued on page* 573)

driven amplifier Q3 followed by a further d.c. coupled amplifier Q4, which between them square off the sine wave input and give a 1Mc/s square wave output from the collector of Q4. To achieve high oscillator stability, the crystal is normally mounted in a constant-temperature oven. For stabilization against supply voltage changes, the d.c. supply to the two oscillator transistors is provided from a 9V Zener-stabilized rail as shown in the diagram.

Timebase Dividers

The master oscillator by itself can provide only a 1μsec gate time, and to deal with lower frequency input signals outputs at 100kc/s, 10kc/s, etc., down to 1c/s are normally required to provide gate times of 10μsec, 100μsec, etc., up to 1 second. The *timebase dividers* section of Fig. 92 is usually arranged as shown in Fig. 96(b). Here the master oscillator square-wave output at 1Mc/s is fed into the first DCU, which divides by ten and gives a 100kc/s square-wave output. This again is fed into a further string of DCUs, each of which in turn divides by ten. The outputs from these dividers provide trains of square waves at decade intervals down to 1c/s. Any of these can be selected for controlling the open time of the signal gate.

Control Unit

We now come in Fig. 92 to the logical heart of the frequency meter, the gate *control unit*. This is the most complex part of the equipment aid in Fig. 97 is set out in diagrammatic form the control unit of a typical counter to use as an illustration of how it works. Designs may differ in detail from Fig. 97 but all follow pretty much this pattern. The control unit is designed primarily to control the signal gate shown dotted at the top of the diagram, and its operations are controlled partly by the output from the timebase shown coming in on the left of the diagram. To illustrate the operation, let us assume to start with that the signal gate is closed. Let us also assume that the various binaries have all been reset in the control unit and the counter.

The first pulse from the timebase arriving at the open timebase gate passes through to the gate control binary and flips it over. The flipped binary feeds a d.c. signal through the signal buffer amplifier to open the signal gate. This lets the stream of signal pulses start feeding into the counter. The next timebase pulse arriving still finds the timebase gate open and flips the gate control binary back to its original state. This cuts off the d.c. control voltage to the signal gate and closes it. The stream of signal pulses to the counter is cut off, and the counter holds and displays the number of pulses that it has received while the gate was open.

In flipping back to its original "set" state, the gate control binary sends a positive pulse to the latch binary which causes it in turn to flip over. In flipping over, the latch binary sends a negative d.c. voltage to the latch buffer amplifier inverter. This voltage is inverted in the latch buffer and appears as a positive voltage to close the timebase gate so that further timebase pulses are inhibited (prevented passing through into the system) until the gate can be reopened.

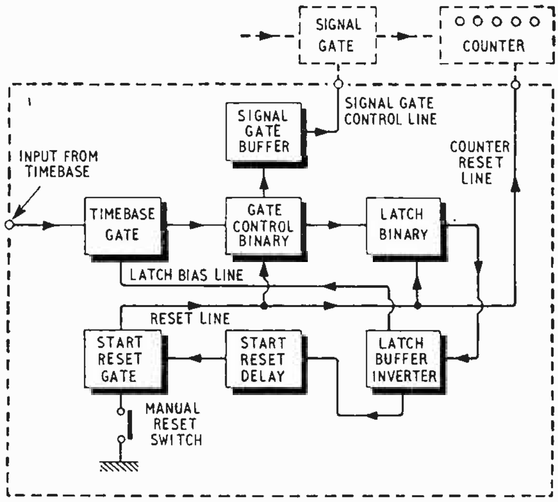


Fig. 97. Control unit of counter timer.

If we want the counter/timer to sample the input signal repetitively we must arrange for the whole system to be automatically reset to the initial state to let the cycle repeat again and again. This is achieved by using the same positive d.c. voltage shift from the latch buffer inverter that was used to close the timebase gate also to trigger off the start reset delay unit in Fig. 97. This circuit is a pulse delay unit which emits an output pulse at a predetermined interval after it receives the input pulse. The delayed output pulse from the start reset delay is used to open-circuit the start reset gate in the reset line. The opening of the start reset gate removes the earth connection of the reset line and resets the latch binary and the DCUs in the counter. In being reset, the latch binary emits a negative d.c. voltage level change which (inverted through the latch buffer inverter to a positive d.c. voltage level change) opens the timebase gate once again and lets the counting cycle repeat "from scratch." As the start reset delay circuit is arranged to respond only to a positive going pulse, it is not affected by this output from the latch buffer inverter.

Besides the automatic start reset provided by the start reset delay circuit, equipments usually include some form of manual reset switch as shown in Fig. 97. This is normally closed, but can be open-circuited manually to reset the whole counter ready to start counting.

The complete operation of the control unit may seem rather complex at first sight, but careful following through of the logic will soon show how the various circuit blocks work together.

Buffer Amplifier and Inverter

Before we go on to consider the start reset delay circuit in detail, we will first take a look at how the two *buffer amplifiers* in Fig. 97 work.

The *signal gate buffer* amplifier is an emitter follower of the type illustrated in basic form in Fig. 98(a). The output is in phase with the input, so that the negative output voltage from the gate control binary is transmitted to the signal gate also as a negative voltage, suitable for opening the gate.

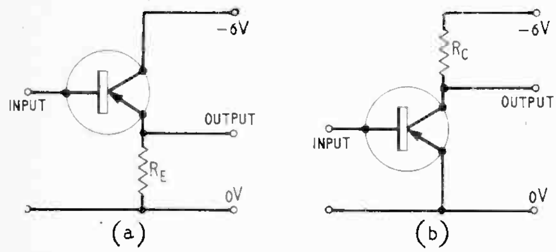


Fig. 98. Buffer amplifiers. (a) Emitter follower. (b) Inverter.

The *latch buffer inverter* in Fig. 97, on the other hand, is a common emitter amplifier of the type shown in Fig. 98(b) which inverts the phase of the input signal and provides from the negative input voltage fed to it from the latch binary the necessary output positive voltage to trigger on the start reset delay circuit.

Start Reset Delay

This brings us finally to detailed consideration of the last remaining circuit in Fig. 97, i.e., the *start reset delay* circuit. This is usually some form of monostable, and generally it is a multivibrator. Fig. 99 gives the circuit of a typical example to illustrate the design points. Q1 and Q4 together form a monostable multivibrator, with resistive cross-coupling from collector to base in one direction and capacitor cross-coupling in the other. The collector of Q4 is cross-coupled to the base of Q1 via the 3.3kΩ resistor. The 1,000pF capacitor across the 3.3kΩ is merely a speed-up capacitor, negligible compared with the resistor. The capacitor cross-coupling from the collector of Q1 to the opposite base is via the emitter follower isolating transistor Q2 through the 50μF coupling capacitor to the base of Q3. Q3 and Q4 form a Darlington pair equivalent to a single very high gain transistor.

The circuit operates roughly as follows. In the quiescent condition, the transistor Q4 is switched hard on by the 6.8Ω, 470kΩ and variable 250kΩ network to the negative supply rail. When a positive pulse comes in from the latch binary, it passes through the diode D1 and the 100pF capacitor to the base of Q4. This switches transistor Q4 off and Q1 on. A negative-going pulse appears at the collector of Q4 and is transmitted through the output coupling capacitor to the base of the start reset gate switch transistor Q5. Now Q5 is normally held switched hard on by the bias network of resistors to its base, and the additional negative pulse merely drives it harder on. Q5 thus remains fully bottomed, keeping the reset line shorted to deck.

During the quasi-stable state of the monostable, the 50μF capacitor discharges steadily with a time-constant set by the variable network of resistors connected to its right-hand end. Eventually, at a time determined by the setting of the 250kΩ variable resistor, the monostable flips back to its stable condition. When this happens, a positive pulse appears at the collector of Q4, is transmitted to the base of Q5 and cuts that transistor off for a short time. This open-circuits the reset line from deck and causes all the binaries in the control unit and the DCUs in the counter to be reset.

When the latch binary is reset in Fig. 97 the negative voltage on its output disappears and similarly the positive inverted voltage at the output of the latch buffer inverter which has been holding the timebase gate closed also disappears. Thus the timebase gate is reopened.

The whole system is thus reset ready to start as before, and the next input pulse from the timebase triggers off the counting cycle once again. The repetition time between sample counts is set by the 250kΩ variable resistor control on the start reset delay unit. In a practical instrument the recycling time is normally variable between a small fraction of a second and some tens of seconds.

Summary

In this article we have attempted to set out the main operational features of a typical counter-type frequency meter as a demonstration of uses of some of the basic elementary pulse circuits dealt with in previous articles in the series. In the space available it has not been possible to go into all the possible circuit variations likely to be met with in a commercial counter/timer, but most counter/timers work very much along the main lines indicated above.

[The final article in this series will appear in our next issue. We have received many applications for back issues containing earlier articles in this series, but many of these are already out of print. However, arrangements are being made to re-issue the whole series in book form in the New Year and the publication date will be announced as soon as it is known.]

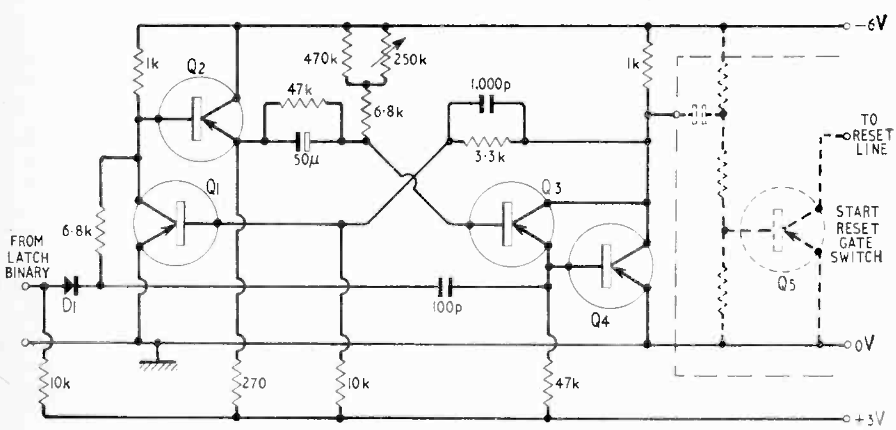


Fig. 99. Start reset delay monostable circuit. All transistors are silicon v.h.f. type.

WORLD OF WIRELESS

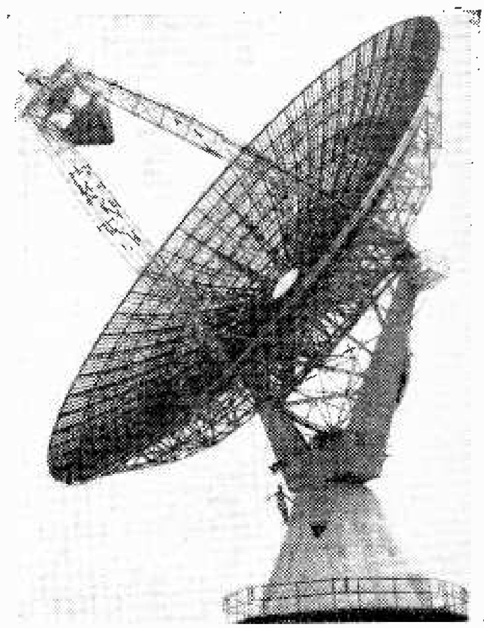
International Colour TV

ANOTHER series of demonstrations of colour television is being staged in London during the last week of October for some 50 European delegates. The members of the colour television *ad hoc* group of the European Broadcasting Union have invited representatives of the Eastern European O.I.R.T. (International Radio and Television Organization) to attend and it is understood that five have accepted. In addition two members of the secretariat of the C.C.I.R. will also be present.

The delegates are meeting primarily to study results of additional tests made since the last meeting of the *ad hoc* group some months ago when it was felt there was insufficient information available on certain aspects of the various systems. One of the things that has come out of earlier discussions is the necessity for making tests over long circuits. It is therefore planned, during this meeting, to transmit each of the three systems, PAL, SECAM and N.T.S.C., over a circuit from London via Germany to Rome and back to London via France—the longest distance so far attempted. It has also been agreed in principle to arrange a London-Moscow-Paris link later in the year.

The London demonstrations are being staged by the B.B.C., the Independent Television Companies' Association, and some manufacturers.

A meeting of the *ad hoc* group will be held in Hilversum in December to study the results of these demonstrations and to produce a report which will be submitted to the meeting of the C.C.I.R. in Vienna next March at which it is hoped an international decision will be reached on a colour system.



This 85-foot dish at the U.S. Navy's Point Mugu tracking station in California was modified for the reception of the Olympic Games television transmissions via Syncom III by the Hughes Aircraft Company who designed and built Syncom.

Televising the Olympic Games

A SEVEN-STOREY building recently completed in Tokyo was the broadcasting centre for this year's Olympic Games. It contains seven television studios and forty-five sound studios. Some 80 television cameras, 32 video tape recorders (including two with slow-motion facilities), 30 microwave terminals, 700 audio tape recorders and 600 microphones are in use. Incidentally, after the Games the central broadcasting building will be modified to form part of a permanent broadcasting

Two of the television studios in the central broadcasting building were used by the E.B.U. member countries, of which Britain is one, to prepare material for transmission. To speed up the transmission of television pictures to Europe, use was made of Syncom III which is stationed above the Pacific at a height of approximately 22,300 miles mid-way between Japan and America. Highlights of the Games, recorded on video tape, were transmitted by microwave link from Tokyo to the new satellite ground station at Kashima and thence via Syncom III to Point Mugu in California. From there microwave links carried the recorded video signals across the U.S.A. to Montreal. The received signals were then re-recorded and flown—by chartered jet—to Hamburg, Germany, for feeding into the Eurovision network. The video recordings were converted from the Japanese 525-line, 60-field standard to the 625-line 50-field standard for feeding to those countries using it, but we in this country and France received the 525-line signal for "home conversion". The commentaries to accompany the pictures were sent via the normal cable and wireless links.

The cost of the use of the satellite and charter jet for the period of the Games is said to be in the region of £350,000. This will be met by the Eurovision countries.

This year's I.E.E. **Faraday Lecture** is being given by F. C. McLean, C.B.E., the B.B.C.'s director of engineering. His subject is "Colour Television." These annual lectures are designed to appeal to the layman and are given in a number of provincial centres as well as in London. Mr. McLean will deliver his lecture first in Swansea on November 24th and in Bristol on the 26th. The London lecture will be on February 17th at 6.0 in the Central Hall, Westminster. The lecture tour includes Birmingham (Jan. 19); Leicester (Jan. 21); Manchester (Jan. 26); Stoke-on-Trent (Jan. 28); Portsmouth (Feb. 9); Bradford (Feb. 24); Sheffield (Mar. 2); Belfast (Mar. 18); Edinburgh (April 1); Newcastle (April 6). Admission is by ticket obtainable from the local centre of the I.E.E. In most centres Mr. McLean will repeat his lecture on the following day to an audience of students.

Trafalgar Square Television.—A Marconi large-screen colour television projector, modified for back projection of black and white pictures, was installed in Trafalgar Square to show the waiting thousands the B.B.C. television coverage of the General Electeion as the results became known. The projector, installed and operated for the B.B.C. by Colour Television Engineering Services Ltd., used three E.E.V. high-brightness projection c.r. tubes.

"The Age of Automation." is the title of this year's B.B.C. Reith Lectures which are to be given by Sir Leon Bagrit, the chairman of Elliott-Automation. Sir Leon will develop his ideas on automation in a course of six weekly lectures beginning on Sunday, 8th November, in the Home Service. Each lecture will be repeated in the Third Programme later in the week.

Amateur Television DX.—It is learned from the *R.S.G.B. Bulletin* that a new world record for amateur television was set up in this country on September 3rd. A 200-mile two-way vision and sound link was established between J. R. T. Royle (G3NOX/T), of Saffron Walden, Essex, and L. Hunton (G3ILD/T), of Darlington, Co. Durham. The vision frequencies used were 428 and 436 Mc/s. The Essex station used a power input at peak white of 150 watts and the Durham station 100 watts.

An additional aerial has been ordered from E.M.I. Electronics by I.T.A. for installation on the 750ft mast at St. Hilary, nr. Cardiff, to radiate a Welsh programme. It will transmit in Channel 7 (the existing aerial radiates in Channel 10) and will have a maximum e.r.p. of 100kW with the main lobes of its kidney-shaped radiation pattern directed towards the N.E. and N.W. The new aerial will be vertically polarized and is due to be completed by December.

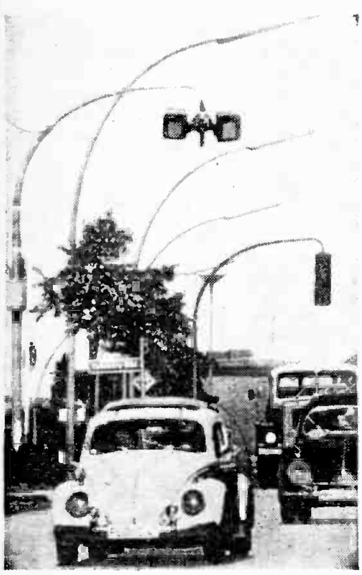
BBC-2 test transmissions for the Birmingham area are expected to start in Channel 40 from Sutton Coldfield on 16th November. The actual service is scheduled to begin on 6th December.

Holland's second television service was introduced on 1st October. Initially it is being radiated by only one station—Lopik-IJsselstein. Both services are conducted by the Netherlands Television Foundation to which each of the five Netherlands broadcasting authorities contribute.

PAL, Telefunken's colour television system, has recently been demonstrated to Eastern European broadcasting authorities in Moscow and Sofia.

The next **German Radio and Television Show** will to be held in Stuttgart from 27th August to 5th September, 1965.

Asian Broadcasting Union.—The first general assembly of this Union, which formally came into being in July, will be held in Sydney, Australia, in November. Eleven broadcasting organizations have joined the A.B.U. which has its headquarters in Tokyo. The countries represented are Australia, India, Japan, Korea, Laos, Malaysia, New Zealand, Pakistan, Philippines, Taiwan and United Arab Republic.



Six radar detectors, are used with an electronic traffic analyser to control traffic lights at one of the major road junctions in Hamburg. The number, length and speed of all vehicles travelling along one lane are detected and this information is fed to the analyser in pulse form. Using this system, which is still undergoing tests in Hamburg, the traffic lights of a whole district can be regulated. Telefunken supplied the detectors and Standard Elektrik Lorenz the analyser.

B.V.A. Officers.—F. V. Green, Brimar representative on the Council of the British Radio Valve Manufacturers' Association, has been re-elected chairman. The new vice-chairman is A. Deutsch (Thorn-A.E.I. Radio Valves & Tubes).

E. E. A. Joins EUROSPACE.—The Electronic Engineering Association, which represents the capital goods section of the British electronics industry, has joined EUROSPACE as a collective member. This European Industrial Space Study Group was formed in 1961 to promote space activities in Western Europe.

The I.I.T. Research Institute, formerly the Armour Research Foundation of Illinois Institute of Technology, has opened a European office at 1 Av. du Général Leclerc, Fontenay-aux-Roses (Seine), France. It will provide training in and information about the A.P.T. (Automatically Programmed Tools) system for programming numerically controlled machine tools.

Hong Kong-U.K. Trade Over £110M.—According to figures in the 1964 edition of the Hong Kong "Commerce, Industry and Finance Directory" the Colony's export of transistor portables for 1963 amounted to H.K.\$68.3M (£4.2M) and shows an increase of almost 100% over the previous year's total. Copies of this 180-page review of Hong Kong's industrial growth are available, on receipt of a self-addressed label bearing 1s 2d in stamps, from the Hong Kong Government Office, 54 Pall Mall, London, S.W.1.

The **National** range of radio receivers, recorders and domestic electrical appliances made by the Matsushita Electrical Industrial Company, of Japan, are on shew at the Japan Trade Centre, 535 Oxford Street, London, W.1, until 12th November. The exhibition is open to the public from 9.30 to 5.30 on Mondays to Fridays.

Semiconductor Diodes.—Two new colour filmstrips have been introduced by the Mullard Educational Service. The first, entitled "Basic principles of semiconductor diodes" (30 frames) gives a detailed description of intrinsic, p-type and n-type semiconductors.. The second strip, entitled "Practical semiconductor diodes" deals with the processing of materials, the construction of devices and characteristics. Other items of interest in this 25-frame strip include avalanche Zener and tunnel effects, and the high frequency and thermal limitations of diodes. Both strips are available from the distributors, Unicorn Head, Visual Aids, Ltd., 42 Westminster Palace Gardens, London, S.W.1. Comprehensive teaching notes are included in the price of 25s per strip.

Reporting at the annual general meeting of the **Electrical Industries Benevolent Association** the retiring president, F. E. C. Miller, drew attention to the fact that the Association suffered a deficit of £5,660 last year compared with a surplus of £6,500 the year before. The Association, which last year helped 2,000 people, is supported entirely by voluntary contributions.

The "**Handlist of Basic Reference Material**" for librarians and information officers in electrical and electronic engineering has been revised by the ASLIB Electronics Group. It is broken down into the following sections; enclyopædias and dictionaries, handbooks and yearbooks, trade directories, guides to literature (sub-divided into book lists, periodicals, lists and subject bibliographies), standards, tables, valves, tubes and transistors handbooks, technical writing and terminology and guides to sources of information. Copies of this 48-page publication are available, price 12s, from Miss B. Newman, Ericsson Telephones, Beeston, Notts.

A symposium on the theory and application of the **silicon gate-controlled switch** will be held at Enfield College of Technology, Queensway, Enfield, Middx., on 2nd December.

PERSONALITIES

Sir Robert Cockburn, K.B.E., C.B., M.Sc., Ph.D., chief scientist of the Ministry of Aviation, has been appointed director of the Royal Aircraft Establishment, Farnborough, in succession to **Dr. M. J. Lighthill,** F.R.S., who is taking up a Royal Society Research Professorship. Sir Robert, who is 55 and was knighted in 1960, taught physics at Portsmouth and West Ham Municipal Colleges before joining the Radio Department of the Royal Aircraft Establishment in 1937. Two years later he became head of the Radio Countermeasures Division of T.R.E. (now Royal Radar Establishment), Malvern. In 1945 he transferred to atomic energy research and three years later was appointed scientific adviser to the Air Ministry. He became principal director of scientific research, guided weapons and electronics, in the Ministry of Supply in 1954. In the following year he was appointed deputy controller of electronics, a year later controller of guided weapons and electronics, and since 1959 has been chief scientist.

Sir Robert Cockburn

M. Lionel Jofeh, O.B.E., M.I.E.E., managing director of Sperry Gyroscope Company for the past two years has additionally been elected chairman of the board. He joined Sperry in 1947 to undertake research and development work on radar, fire control and missile systems and 10 years later became chief engineer. In 1959 he was appointed manager of the newly formed industrial division and in the following year joined the board. Prior to joining Sperry Mr. Jofeh, who is 51, was for 11 years with Cossor, first as a television research engineer and later as technical assistant to the director of research. He was appointed an O.B.E. in 1956 for his work on defence projects.

Sir Ian Orr-Ewing, Bt., O.B.E., M.P., M.A.. M.I.E.E., has accepted the invitation of the recently formed Society of Electronic and Radio Technicians to become its first president. Sir Ian, who resigned from the Government as Civil Lord of the Admiralty in 1963, started his career in the radio and electronics industry as a graduate apprentice with E.M.I. in 1934. After war service, during which he was at one time chief radar officer, air staff, S.H.A.E.F., he was manager of the B.B.C's television outside broadcasts, and from 1949-57 was a director of A.C. Cossor. He is now chairman of United Dot Products and Carr Fastener Co. Ltd.

J. T. Wiltshire, M.A., has been appointed controller of G.E.C. (Electronics) Ltd. with responsibility for finance and production. He has been the company's director for commercial and production matters for two years. Prior to joining G.E.C. Electronics Mr. Wiltshire, who is 44, was divisional manager and co-ordinator of a complete guided missile system with E.M.I. Electronics. He joined R.E.M.E. as a regular officer from Cambridge (where he obtained double First Class Honours in Mathematics and a Research Scholarship) and was an instructor at the Military College of Science.

Three appointments are announced by A.E.I. following the reorganization of the company's valve and semiconductor business at Lincoln. **Eric Willis-Jones** becomes general manager, semiconductors, and will be responsible for the design, production and marketing of semiconductor devices. He was previously commercial manager of the Lincoln valve & semiconductor department. **H. W. Cumming** has become project manager, semiconductors, responsible for the development and production of all devices at A.E.I. Lincoln. He joined A.E.I. (Siemens Bros.) in 1956 after 16 years with Siemens Electric Lamps and Supplies. In 1958 he was appointed chief engineer, semiconductors, with Siemens Edison Swan at Woolwich and two years ago transferred to Lincoln as assistant manager, engineering. **Michael W. Blades,** who has been assistant manager of the Lincoln semiconductor sales department, becomes commercial manager. Mr. Blades, who is 35, joined the Brimsdown factory of A.E.I. (Edison Swan) in 1953. In 1961 he transferred to the radio and electronic components department as deputy chief engineer later becoming head of semiconructor product research at Lincoln.

A. Geoffrey Slemeck, B.A., marketing director of Standard Telephones & Cables, has been made president and managing director of I.T.T. Standard, the European marketing organization of International Telephone & Telegraph Corp., and will be based in Brussels. He joined S.T.C. in 1946 after war service in the Royal Navy, in which he reached the rank of Lieutenant Commander, and has been a director since 1962. The new marketing director of S.T.C. is **K. P. Wood,** B.Sc.(Eng.), M.I.E.E., who joined the company in 1962 from British Communications Corporation Ltd. where he was director and general manager. He has been manager of S.T.C's radio division and is also a director of the two subsidiaries coming under the management of the division—International Marine Radio Company and Hudson Electronics Ltd. Mr. Wood was at one time head of the Electrical Engineering Department at Medway College of Technology, Rochester, and from 1956-1961 was with Cossor.

H. E. F. Taylor, secretary of the Electronic Engineering Association since 1956, has retired. Mr. Taylor was a Lieutenant Colonel in the Royal Signals during the war and was concerned mostly with long-distance telecommunications projects. He has been a radio amateur for over 40 years, and whilst in India operated with the call VU2AT. His British call is G6HT. The new secretary of the Association is **A. S. Marshall,** who joined as deputy secretary in 1958 on retirement from the Royal Navy with the rank of Lieutenant Commander in the Fleet Air Arm. He was at one time in the Admiralty's Radio Equipment Department and later at the Admiralty Research Laboratory at Teddington.

A. S. Marshall

INDUSTRY NEWS

M. Esterson, B.Sc., A.M.I.E.E., manager of the high power klystron department of English Electric Valve Company for the past two years, has been appointed personal representative of the managing director to undertake special duties concerning the integration of the recently acquired Lincoln factory into the Chelmsford organization. The high power klystron and microwave research sections will now be combined to form the microwave beam tube department of which **G. O. Chalk,** M.Sc., A.M.I.E.E., will be in charge. Mr. Chalk graduated at University College, Exeter, in 1950, and after service in the R.A.F. joined the Mullard Research Laboratories in 1953 where he was concerned with research and development of klystrons. He has been with E.E.V. since 1957 latterly as assistant manager of microwave research.

G. O. Chalk

Dr. L. Rohde, of Rohde & Schwarz, of Munich, for whom Aveley Electric Limited are U.K. distributors, has joined the board of Aveley. The Company also announces the appointment to the board of **A. C. Green,** who has been works manager since he joined the company on its formation in 1954.

John R. Erskine, who was made chief engineer of Aveley Electric's Avel-Toroid Division on its formation in 1960, has been appointed chief engineer of Aveley.

Eric Lawson has joined Cosmocord, of Waltham Cross, to be manager of the new instrument division. He was previously in charge of test and instrumentation with Wilkinson Sword (Colnbrook) Ltd.

B. R. Pittaway, chief engineer of Electro Acoustic Industries Ltd. since 1959, has been appointed a director. Since joining the company in 1951 he has been concerned with the design and development of the Elac range of loudspeakers.

Thorn-Ultra.—It having proved impossible to reach agreement, the negotiations between Ultra Electric (Holdings) Ltd. and Thorn Electrical Industries Ltd. with a view to Thorn acquiring Ultra have been terminated. The negotiations were also concerned with Thorn's acquisition of the 40% interest in Ultra Electronics at present held by Electronics International Capital Ltd., of Bermuda. Ultra's domestic radio and television interests (Ultra & Pilot) were acquired by Thorn in 1961.

Australian Computers Pty. Ltd. is the name of a new company jointly formed by English Electric-Leo Computers Ltd. and Amalgamated Wireless (Australasia) Ltd. It has an authorized capital of £A200,000 and is operating from ADC House, 77 Pacific Highway, North Sydney, N.S.W. English Electric-Leo Computers have a 60% interest.

English Electric Leo Marconi Computers Ltd.—The English Electric Company has agreed to purchase J. Lyons & Company's shareholding in English Electric-Leo Computers Ltd., which was formed as a result of a merger of the two companies' computer interests in April of last year. The effect of this change in share ownership is that the computer company is to be directly connected with English Electric's principal electronics subsidiary, the Marconi Company, and will in future be known as English Electric Leo Marconi Computers Ltd. Sir Gordon Radley remains chairman and Mr. W. E. Scott continues as managing director; the Marconi Company will be represented on the board.

Aero Electronics Ltd. have acquired the production facilities of G3SJ Quartz Crystals Resonators Ltd. at Stonehouse Street, Plymouth, and Mr. M. H. Nicholas a director of Aero Electronics has joined the board of the company which will in future be known as Aero Electronics (Crystals) Ltd. Sales will be handled through the parent company's head offices at Gatwick House, Horley, Surrey.

Compagnie générale de télégraphie Sans Fil (C.S.F.) and the General Dynamics Corporation, of America, have joined forces to form a new company to manufacture satellite tracking equipment. The company, of which C.S.F. hold a controlling interest, is to be known as **Société d'Equipements Spatiaux et Astronautiques** (SESTRO) and will have headquarters in Corbeville.

Norwegian Training Vessel.—Associated Electrical Industries and Solartron have provided electronic equipment for the Norwegian Ministry of Shipping and Trade's new training vessel *M.S. Trondelag.* The A.E.I. Electronics Group has provided an "Escort" 13-in radar for the bridge and an "Escort 601" radar for the blind navigation training room. In addition, six "Escort" displays will work in conjunction with a Solartron Type SY 1194 simulator. One is a true motion display, and the other five are relative motion displays.

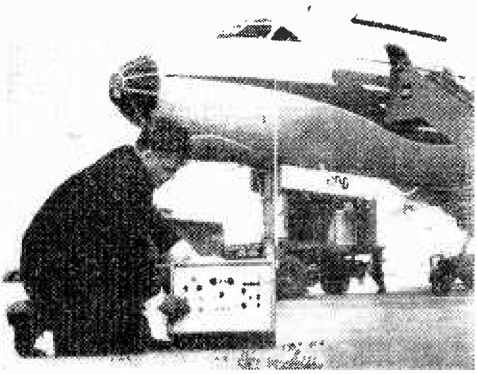
H.F. Distribution System for B.B.C. TV Centre.—A system of wired television using h.f. multi-pair techniques has been chosen to provide simultaneous distribution of up to 11 different television programmes plus the three B.B.C. sound programmes at the B.B.C. Television Centre at Wood Lane, London. Some 100 offices are covered by the Rediffusion system.

Choiceview, the registered name of the subscription television system developed by the joint Rank-Rediffusion research organization, is incorporated in the titles of two recently formed companies. They are Rank (Choiceview) Ltd., and Rediffusion (Choiceview) Ltd.

Beme Telecommunications Ltd., marine communications equipment manufacturers of 24 Upper Brook Street, London, W.1, who became part of the Derritron group of companies a year or two ago, is now to be known as Derritron Communications Ltd.

Crawford, Hansford & Kimber Ltd., of Farnham, Surrey, who make specialized electronic equipment for the industrial field, have recently entered the domestic market with a portable stereo record player called "The New Saturn."

E.E.V. Correspondence—Subsequent to their acquisition of the A.E.I. factory at Lincoln, all correspondence relating to export orders, Government contracts and microwave tubes should be sent to Chelmsford. Correspondence, however, concerning ignitrons, thermionic rectifiers, glass-to-metal seals, and all other products (except semiconductors) previously available from A.E.I. Lincoln, should now be addressed to English Electric Valve Co. Ltd., Carholme Road, Lincoln. (Tel.: Lincoln 26435.)



Complete checks of the aircraft's v.h.f. communications equipment, instrument landing system and V.O.R. omnirange beacon receivers in situ are effected by the QA8A test set in the foreground. The instrument is made by Standard Telephones and Cables Ltd. and is shown testing one of the TU-104 jets in the Czechoslovakian Airlines fleet.

Radar Display Systems.—The radar division of Cossor Electronics Ltd. has received an order, valued in the region of £200,000, for several Type CRD 100 transistor displays and associated equipment for the Ministry of Aviation's air traffic control training school and evaluation unit at Hurn Airport. This installation, which is to have 22 operational positions and two monitoring positions, will provide a system capable of accepting unprocessed radar data—from radar simulators, film recorders, video maps and external radar heads—and distributing, after processing, the information to any or all of the display positions.

The **S.T.C. entertainment range of transistors** marketed through Thorn-A.E.I. Radio Valves & Tubes Ltd. (Brimar) are in future to be distributed by S.T.C. Electronic Services, Edinburgh Way, Harlow, Essex.

Vactric Control Equipment Ltd. has acquired the whole of the issued share capital of A. P. Besson & Partner Ltd., whose products include miniature microphones, earpieces, transformers, and electromechanical and industrial equipment. Vactric Control also announces that it has exclusive manufacturing and sales rights for the Norden range of encoders in the United Kingdom and non-exclusive sales rights for the rest of the world with the exception of France, Canada and the U.S.A. A service and repair organization has been set up to cater for the encoders already in use in the U.K. and manufacturing plant is being installed at the company's premises in Garth Road, Morden. Norden is a division of the United Aircraft Corporation, of Connecticut, U.S.A.

U.K. manufacturing rights for the production of u.h.f. television transmitting aerial panels have been granted by **Rohde & Schwarz**, of Munich, to **Racal** Electronics Ltd., of Bracknell, Berks.

The American organization **Datapulse Incorporated** have signed an agreement with Digital Measurements Ltd., of Salisbury Grove, Mytchett, Aldershot, Hants., for the marketing of their pulse generators in the United Kingdom.

The Dutch range of **Jobo turntables**, which are made by Acoustical Handel Maatschappij N.V., of Holland, are available in this country from Colton & Co. (Lapidaries) Ltd., The Crescent, Wimbledon, London, S.W.19, who are the sole U.K. agents.

Dynaco Incorporated, of Philadelphia, have appointed Howland-West & Co., of 11 Howland Mews E., Howland Street, London, W.1, sole U.K. distributors of their range of products which includes several stereo amplifiers.

R. H. Cole Electronics Ltd. have moved from Caxton Street, Westminster, to 7-15 Lansdown Road, Croydon, Surrey. (Tel.: MUNicipal 4411.)

The **International Marine Radio Company**, formerly of Progress Way, Croydon, have moved to a larger factory in Peall Road, Croydon. (Tel.: THOrnton Heath 9771.)

The sales department of **Sydney S. Bird & Sons,** manufacturers of Cyldon components and accessories, has been transferred from Enfield, Middx, to the works at Fleets Lane, Poole, Dorset. (Tel.: Poole 1640.)

Jason Electronic Designs Ltd., makers of the **Jason** range of equipment and home-construction kits, and their associate companies Radio Traders Ltd. and the Lorlin Electronics Co., have moved from Wardour Street to Tudor Place, London, W.1. (Tel.: MUSeum 4666.)

ANNUAL REPORTS

The Plessey Company and its subsidiaries announce a trading profit for the year ended 30th June of £13,220,488, compared with £11,408,026 for 1962/63. After providing for taxation (£5,957,635, which is slightly up on the previous year) and setting aside £3,300,877 (£2,584,530) for depreciation and £860,000 to meet other deductions, the net profit for the year totalled £7,043,414 and shows an increase of £1,263,760 on the previous year's result. An indication of the company's activities may be obtained from the following percentage breakdown: telecommunications 53%; electronic and electrical products 16%; products for the consumer durable industries 14%; aircraft systems and equipment 8%; hydraulic and specialized engineering 6%; and telecommunications rental income 3%. Plessey-UK Ltd. increased their research and development effort and spent more than £5M in the year under review.

E.M.I. Group Profit Over £9M.—Pre-tax profits of Electric and Musical Industries Ltd. for the year ended 30th June, 1964, amounted to £9,104,000 and shows a considerable increase on the previous year's result of £5,058,000. Total group sales for the year were up nearly £10M at £94,675,000 and group trading profits rose by £4M to £11,679,000. Allowances for U.K. and overseas taxation almost doubled at £4,759,000 (£2,463,000) and so did the company's net profit of £4,161,000 (£2,405,000). All sections of the company, both at home and overseas, contributed to this improvement in the accounts.

Decca Profits Down.—Although the net profit of Decca Ltd. for the year ended 31st March, 1964, amounted to £1,339,000 and showed a slight drop on the previous year's result of £1,444,000, the consolidated turnover reached a new high level at £30,200,000 (£28,000,000). The record side of the business, which accounts for more than half of the consolidated turnover, showed increased profits, while on the electronics side there was a less favourable result.

Telefusion Ltd. announce that its trading profit for the 53 weeks to 29th April, 1964, was £2,010,554. This shows an increase of almost £300,000 on the previous year's result. Depreciation this year rose from £1,003,333 in 1962/63 to £1,235,943, whilst taxation charges dropped slightly to £79,516 (plus £8,686 taxation adjustments relating to the previous year). The net profit amounted to £669,135 compared with the previous year's £438,247.

Fourth International Battery Symposium

BRIGHTON, 29th SEPTEMBER — 1st OCTOBER, 1964

ORGANIZED by the Inter-departmental Committee on Batteries the Fourth Symposium with nearly 400 representatives from 22 countries was held at Brighton and was successful from many viewpoints not least the surroundings (in this connection it should be noted that the fifth Symposium will be held at the Hotel Metropole, Brighton, September 20th-22nd, 1966).

Lead-Acid Batteries: A paper on the old problem of initial voltage dip and peak in the discharging and charging curves of batteries by Dr. D. Berndt (Varta) gives an explanation for positive plates, hitherto not quite so detailed, on the basis of crystallization over-voltage (for the voltage dip case). In the case when the active material is pure αPbO_2 no dip exists and as yet there is still no satisfactory explanation for this.

Design and development of lead-acid batteries was discussed in a number of papers; the use of dispersion hardened lead (lead strengthened with a disperse phase of oxide) was one of the more interesting topics. This is found to increase the life of the positive plates but, of course, there are problems in the fabrication and joining, mainly because the lead cannot be melted or cast and consequently the processing is expensive or wasteful, involving either machining or stamping.

Sealed Cells: One of the most recent innovations in sealed cell technology is the use of an auxiliary electrode for improved overcharge capability and end-of-charge control. The Adhydrode control electrode has been developed, based on work by R. C. Shair and H. N. Seiger (Gulton Industries Inc.) on the theory of the mechanism of the oxygen reaction at the negative cadmium electrode in nickel and silver-cadmium cells. Electrons from a corrosion reaction due to a $Cd/Cd(OH)_2$ couple at the negative electrode, form adsorbed hydrogen at what has been termed the "Adhydrode" electrode and current flows through an external circuit to the negative electrode. This is greatest when oxygen arrives at the Adhydrode electrode (to react with the hydrogen) as the cell becomes fully charged. The current in the auxiliary electrode circuit is proportional to the oxygen pressure and thus a predetermined pressure can be set to control the end-of-charge. The Adhydrode provides a control method for recharging cells rapidly at high currents.

Solar Cells: Development of silicon photovoltaic cells continues to be dominated by space vehicle requirements. The major limiting factor for the extensive application of silicon cells on earth has been the high cost involved (in space high cost is not such a limiting factor). A paper by V. Magee and A. Bardsley (Ferranti Ltd.) described recent work aimed at reducing costs and mention is made of power generation costs of 0.1d/watt hour. Storage batteries (nickel-cadmium) obviously must be provided in a satellite since a third of the orbit time is in darkness, and J. L. Blondstein (B.A.C.), D. F. C. Poole (B.A.C.) and D. E. Mullinger (R.A.E.) discussed power regulation systems in satellites. A basic system would involve use of two batteries, a solar array and logic-controlled switching. The circuits must be arranged to trickle-charge the stand-by battery, isolate the load while discharged batteries recover, sense battery charge current, temperature, terminal voltage and cell pressure. In addition timing circuits may be needed to terminate charging or to reconnect the load after isolation and count orbits. In the system proposed for the UK3 satellite some 8,000 silicon cells are arranged in two groups, one for charging the batteries (twenty-two sealed nickel-cadmium cells of 3A-hr. capacity) and the other for supplying the load. The large number of cells used allows for a 25% reduction in all output due to proton bombardment.

Radio-Isotope Cell: Energy from radioactive emanations has been transformed into useful electrical energy as long ago as 1903, by collecting the beta particles from a radium source. The isotope used in this type of cell should be a beta emitter within an acceptable energy range, with minimal gamma radiations and a half-life compatible with the required operating life. The isotope must be available in suitable form and Krypton-85 was chosen for the cell discussed, in view of its low chemical activity. The choice of dielectric (separating source and collector) is governed in the first instance by its resistivity and resistance to radiation damage and these factors lead to the choice of polystyrene. Krypton, of course, is an inert gas and does not form any solid compounds, but it will form inclusion compounds or clathrates in which the krypton atoms are trapped in the host crystal lattice (hydroquinone) and cannot escape unless the lattice is broken. A cell measuring nearly an inch in diameter and $\frac{3}{16}$in thick can produce 500 volts at around 280$\mu\mu$A. The cell can develop 220 V across a 0.1μF capacitor (insulation resistance $10^{13}\,\Omega$) in a day and limits at 2.5 kV. With very low current gas diodes timing circuits and low-frequency sawtooth waveforms with periods of hours or even of days can be produced.

Fuel Cells: The choice of electrodes for acid electrolytes, preferred by many because of their ability to reject CO_2 thus enabling use of hydrocarbons directly, is an interesting question. Carbon and graphite are relatively inert but have a short life. Electrodes metallized with platinum, palladium or silver and using platinum-black pastes are rather weak mechanically, while those consisting entirely of platinum or palladium are expensive. P. Ruetschi and J. Sklarchuk (Electric Storage Battery Co., U.S.A.) discusses electrodes which only contain a very small percentage of the costly precious metals in a mechanically strong sintered-metal matrix. Ta, Nb, Zr, Ti and Pb have been investigated and titanium has been singled out for oxygen electrodes because of its low sintering temperature, ease of handling and relatively low price. But it was found that a long term disintegration probably takes place and tantalum and niobium may have to be used. For use in the hydrogen electrode tantalum and niobium appear to provide the best corrosion resistance. The amounts of platinum required for activation was of the order of 100 mg/cm^3.

A paper read in the absence of its authors (J. A. Le Duc and C. Lurie (M. W. Kellog Co., U.S.A.) was concerned with barium fuel cells which offer high current drain capability and one such cell can offer the same power delivery as seven or eight hydrogen-oxygen cells. These cells are, of course, continuous feed primary systems and solid barium fuel electrodes were coupled with oxygen and chlorine. The cells can handle large overloads and performed well at current densities of 1,000 A/ft^2. The results indicate that the barium systems have higher net energy densities per unit volume than any known electrochemical power source.

Introduction to

Practical Transistor Circuit Design

3.—TOLERANCES AND BIASING

By G. P. HOBBS,* B.A., Grad. I.E.E.

SOME radical properties of the two forms of transistor are depicted in Fig. 1. Normal operation is with one junction (base-emitter) forward conducting and the other (collector-base) biased in the reverse direction. At Fig. 1(a) a positive pulse applied to the base initiates collector current and the collector terminal then assumes some potential within the bounds set by the supply rails. By contrast the base potential will always remain linked, within a matter of a few hundred millivolts, to the emitter potential.

Two further circumstances may befall the transistor:

(1) *Where the base is driven to cut off the flow of collector current.* The drive will then be in the opposite direction to that shown in Fig. 1 and the potentials of all three electrodes become " free."

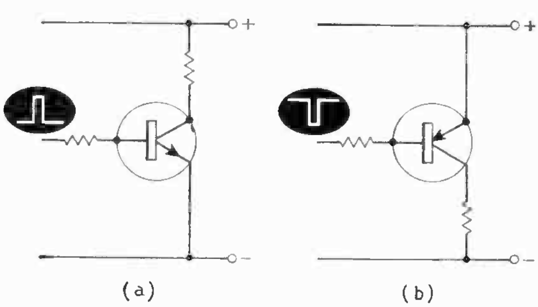


Fig. 1. Collector current flows (a) in an n-p-n transistor for a positive base drive and (b) in a p-n-p transistor for a negative-going drive.

(2) *Where sufficient drive is applied to bottom the stage as indicated by a change in the polarity of V_{CB}.* In this condition all three electrodes become tied together, linked within orders of a few hundred millivolts.

These conceptions of transistor behaviour are essential to the problems of biasing whether the stage is acting as an amplifier or a switch. This article will consider the issues of leakage current and of tolerance in current gain for the transistor acting as an amplifier, while the final article will deal with the subject of base-emitter voltage variation both for amplifiers and switching stages.

In an amplifying stage the working condition is somewhere between the cut off and bottomed conditions. The biasing requirements are therefore more exacting than for either of the two latter limiting conditions.

Let us consider how we might forward-bias the transistor to set up a standing collector current. The two extreme possibilities would be either to apply a small bias voltage to the base terminal with respect to the emitter, or to supply the base with a bias current from (ideally) an infinite source resistance. Under the first procedure the standing current would be influenced by any change in the base-emitter voltage, V_{BE}. By the second method, a change in the current gain, β, would influence the standing current. For both examples the transistor leakage current will displace the working point (but to different extents) and upon all these factors temperature wields a considerable influence.

The problem of biasing is indeed more complex than many would like to admit. Leaving aside V_{BE} variation to the final article we must now consider the subject in detail.

Dependence of the Current Gain on I_C and Temperature:—For a start the typical variation of current gain with collector current is shown in Fig. 2. Strictly it is the d.c. current gain h_{FE}, that is relevant to the biasing question, but here we shall choose to ignore the difference between h_{FE} and h_{fe} (or β) as we shall need to make use of some of the information on parameters discussed in the first article in this series. Generally it will be found worth while only to draw the distinction in high current stages where the incremental current gain has begun to fall away from its maximum value.

A wide range in current is covered by the horizontal logarithmic scale of Fig. 2 and it is only in high-power stages where the amplitude of signal swing could cover such a wide range that the linearity might suffer from change in β. It will be remem-

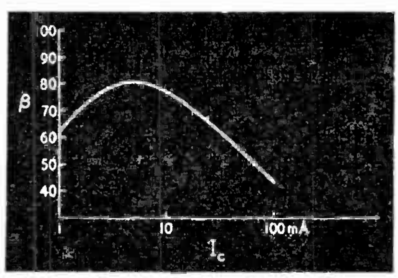


Fig. 2. Variation of current gain with collector current.

*The Marconi Company Ltd., Chelmsford, Essex

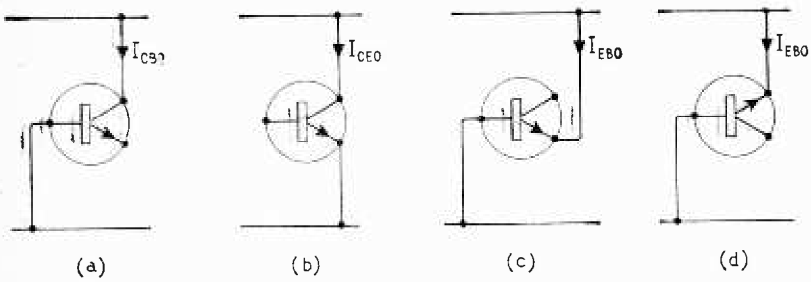


Fig. 3. Designations for leakage currents.

bered that changes in input impedance or g_m are the principal causes behind non-linearity in small signal stages. The main purpose of Fig. 2 is not to show non-linearity, but to demonstrate how the value of β at the chosen operating current can differ from the value quoted in the maker's data. The maximum and minimum values quoted for, say, $I_C = 10$mA would need to be modified in the light of Fig. 2 if the chosen standing current were to be 100mA.

A second, probably more important but often neglected, influence on the limits of β is temperature. β is found in practice to increase with temperature and by quite appreciable amounts: typically between +0.3% per °C and +0.8% per °C. Over the temperature range 0°C to +50°C β could, for instance, drift from a figure of 75 up to 90. So again if we are to design for a wide range of temperature, the extreme maximum and minimum values of β may not directly correspond to the figures quoted in the manufacturer's data; the tolerances may extend further.

All this has really been to show how necessary stabilization can be. Now as a preamble to discussing stabilization against tolerance in β, the problem of leakage current will be examined.

Leakage Current:—In germanium transistors, the generation of leakage comes principally from within the body of the semiconductor; it is a bulk phenomenon. Thermal agitation disrupts electrons from the lattice network of the atoms that form the semiconductor and the so-called electron-hole pairs are produced. The number of electron-hole pairs generated depends on the temperature and on the potential energy associated with their breaking away from the lattice. Normally, electron-hole pairs are recombining as fast as they are formed and the total number remains constant at a given temperature. In the presence of a steep potential gradient such as occurs in the depletion region of the collector junction, the charges will, on the other hand, be swept away before recombination can take place. This charge flow constitutes the leakage current; in diodes it is known as a reverse current. It is characterized by being solely dependent on temperature and originates at a reverse biased junction. Even at very low collector-to-base voltages, the majority of current carriers are swept away before recombination has a chance to take place and no further increase in the applied voltage can appreciably increase the current flow; the leakage is a saturation current.

At least it is true to say that the leakage current is constant up to the point where the electrons gain sufficient energy, whilst accelerating through the potential gradient, to ionize further atoms by collision. In producing additional electron-hole pairs, the process multiplies. It is an avalanche phenomenon and is one mechanism for the voltage breakdown in transistors[1].

There is a fundamental diode equation for semiconductors which shows that the dependence of leakage current on temperature is exponential. Leakage increases rapidly with temperature and in the absence of data, a handy rule of thumb is to double the figure at 25°C for each further 10°C rise of junction temperature. From time to time one comes across quotations for the temperature rise corresponding to a doubling of the leakage current which are either greater or smaller than our choice of 10°C. Really this should be no matter for concern; the rule is only intended as a general guide and the true source of accurate data must be the responsibility of the manufacturer of the device.

Besides the bulk phenomenon, another leakage component is that across the surface of the semiconductor element. In germanium transistors the *proportion* of surface to bulk leakage is small. In silicon, the relative proportion can be significantly higher, because the semiconductor potential energy associated with electron-hole formation is greater and consequently the leakage contribution by the bulk component is several orders of magnitude down. Considered quantatively, however (i.e., in terms of actual current flow), both the surface and the bulk leakage components are small enough to be neglected. (Nobody quite seems to have unravelled the full story for the mechanism of leakage in silicon types and it would therefore be unwise to generalize over a figure for the temperature law.) In this article, when matters of leakage arise, the discussion is intended to be relevant to germanium transistors only.

It is the temperature in the region of the semiconductor junctions which is relevant where leakage current and β variations are concerned, for this is the operative region of the transistor. Because heat is carried away from this region, the area of the junctions must prove to be the hottest part of the transistor, The junction temperature is always higher than the case temperature and in turn the case is always above ambient. (But the case temperature can be brought closer to ambient by the means of a heat sink.) Where data on thermal resistance is available, the excess of junction temperature over the case or ambient temperature can be estimated from a knowledge of the power input to the transistor. This and a method for measuring the junction temperature of a transistor in circuit will be described in the final article in the next issue.

The basic leakage current arises, as we have said,

[1] "Voltage Breakdown of Transistors," by S. C. Ryder-Smith, *Electronic Technology*, October, 1961

in the collector-base junction; it is given the symbol I_{CBO}. The transistor being a three-terminal device has at least three important leakage currents. Each is measured between a pair of terminals. The basic one, I_{CBO}, is measured between collector and base with the emitter left open-circuit. The terminology should be apparent, but Fig. 3 is included to illustrate it.

The measurement is made at the first subscript with reference to the second and the plight of the third terminal is shown by the third subscript. In most instances the leakage will be measured with the third terminal open-circuit as designated by an 'O.' 'S' is used to indicate short circuit, and 'R' to indicate some specified resistance, in each case between the third terminal and the reference terminal (second subscript). The older designation for I_{CBO} (Fig. 3(a)) was I_{CO}, generally read as "I cut-off"; I_{CEO} (Fig. 3(b)) was sometimes written as I'_{CO}. The new descriptions reveal the definitions of the respective leakage currents quite unambiguously.

The identical subscript terminology has been made to serve transistor voltage ratings as well. The ratings may depend on the choice of reference terminal and the state of the third terminal, and it is for instance necessary to distinguish between V_{CBO}, V_{CEO} and V_{CER}[1]. Very often the voltage ratings are quoted in terms of minimum breakdown voltages and the appropriate terms would be BV_{CBO}, BV_{CEO} and BV_{CER}. It is a little difficult to define the exact value of a breakdown voltage; it would be said to occur in the region where the collector characteristics begin to turn round and run away. The standing collector current ought to be stated when quoting breakdown voltages, for higher operating voltages are possible at lower currents.

Fig. 3(c) shows the leakage current associated with the emitter junction, I_{EBO}. Although the base-emitter junction in its operative state is forward-biased, reverse-biasing will occur in switching stages and it is conceivable that I_{EBO} could be a relevant design factor. Naturally, when the emitter-base junction is reverse-biased it behaves in the same fashion as a collector-base junction. I_{EBO} is redrawn in Fig. 3(d) with the transistor "the wrong way up" to emphasize the similarity to Fig. 3(a). The order of magnitude of I_{EBO} will be the same as I_{CBO}, particularly so for a so-called symmetrical transistor. A symmetrical transistor will perform almost as efficiently (that is with a reasonable value of β) when collector and emitter are interchanged. They are of value in certain types of switching circuits.

I_{CEO}, the leakage across both the collector-to-base and base-to-emitter junctions remains to be discussed. The only reverse-biased junction is the collector-to-base one and leakage I_{CBO} therefore originates at this junction. However, in flowing to earth it must need to pass through the base-emitter junction where, like all base currents, it initiates a collector current β times as great by transistor action. If this argument proves unacceptable then the reader may consider it helpful first to imagine I_{CBO} drawn off at the base to complete the journey of I_{CBO}, and then to follow this move by reinserting a current equal in magnitude to I_{CBO} back into the base terminal to satisfy the boundary condition that the base lead is open-circuit. This reinserted current then gives rise to an additional collector current of βI_{CBO}, giving a total of $(\beta + 1)I_{CBO}$. To sufficient accuracy we may write $I_{CEO} = \beta I_{CBO}$[1]. It must be emphasized again that the quantity I_{CBO} is the fundamental leakage component, and that I_{CEO} is the *measured* "leakage current" when the base terminal is made open circuit.

Collector Leakage Current for the General Case:—So much for the specialized leakage currents; our ideas must now be extended to cover the general case in which there is some specific resistance connecting the base and emitter terminals.

Consider once more the current I_{CBO} proceeding from the collector junction. As it has been said the impedance from base to ground is low, but now, Fig. 4(a), there are two contending paths for the current to take. The impedances they present are R_B and $\beta(r_e + R_E)$; the proportion of the current I_{CBO} passing through the base-emitter junction will be a fraction $\dfrac{R_B}{R_B + \beta(r_e + R_E)}$ of the whole and through R_B a fraction $\dfrac{\beta(r_e + R_E)}{R_B + \beta(r_e + R_E)}$ of the whole.

The former gives rise to a collector current β times greater, by transistor action again, and the total leakage becomes

$$\frac{\beta(r_e + R_E) + \beta R_B}{R_B + \beta(r_e + R_E)} \cdot I_{CBO} \qquad (\beta >> 1)$$

that is

$$\frac{r_e + R_E + R_B}{r_e + R_E + \dfrac{R_B}{\beta}} \cdot I_{CBO} \text{ or } S.I_{CBO} \quad .. \qquad (2)$$

It is usual to employ an emitter resistor very much greater than the resistance r_e, so that the factor S we have introduced above becomes

$$S = \frac{R_E + R_B}{R_E + \dfrac{R_B}{\beta}} \quad .. \quad .. \qquad (3)$$

This, the stability factor, relates the leakage in the general case to the basic leakage current I_{CBO}. Expression (3) is perhaps the most practical way of writing the stability factor although some people may be more familiar with the form $S = \dfrac{1}{1 - \dfrac{\alpha R_B}{R_E + R_B}}$ or other related expressions.

Examination of equation (3) shows that S must lie in the range from unity to β. The best stability is realized when R_B is small compared to βR_E, when a good approximation for S is $\dfrac{R_B}{R_E}$. In the limit, of

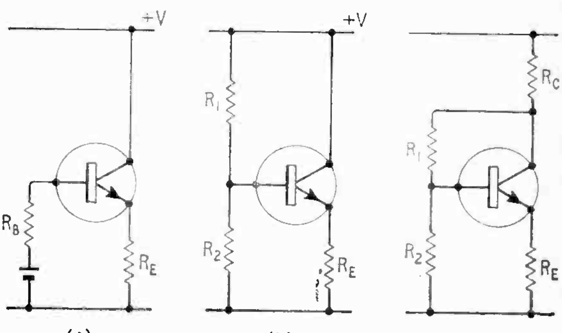


Fig. 4. Leakage current when there is some specific resistance between base and emitter.

course, the smallest leakage that one can aim for is I_{CBO} as S tends to unity. The largest is I_{CEO} which, as we found at (1) is βI_{CBO}. The value of β one employs in equation (1) is influenced naturally by the standing current. This is mentioned because the value of β at low currents may fall off appreciably and so the value of I_{CEO} in the absence of any external forward bias will be smaller than might at first have been imagined. Furthermore, as I_{CEO} is dependent on the current gain its maximum value is rarely quoted by manufacturers except in the case of silicon transistors. Its maximum value for germanium transistors could be alarmingly high, but in a well stabilized circuit where the leakage is closely related to I_{CBO} and not to I_{CEO} this would not matter.

The practical arrangements for the potentiometer method of biasing are depicted in Figure 4(b). Drawing the comparison with Figure 4(a), we can see that R_B should be given the value $R_B = \frac{R_1R_2}{R_1 + R_2}$ and to set up a similar standing current the far end of R_B should be taken to a potential $\frac{R_1}{R_1 + R_2} \cdot V \cdot$

In many instances, the biasing arrangement, whilst not being as simple as shown in Fig. 4 (b), will reduce to the equivalent circuit of Fig. 4(a). A further improvement in the stability is effected where R_1 is returned to the lower end of a collector load, R_C, rather than to the rail—Fig. 4(c). Provided that R_1 remains several times greater than the value of R_C, the stability factor may be found from: $S = \frac{R_E + R_B}{R_E + R_B\left(\frac{1}{\beta} + \frac{R_C}{R_1}\right)}$. Unless R_1 is split into two and the centre tap decoupled, the input impedance at signal frequency will be considerably lowered by the feedback that this method of biasing introduces.

The Stability Factor Applied to β Tolerance:— The biasing current is supplied through a resistor network which in our generalized diagram of Fig. 4(a) is a single resistor fed from a voltage source $\frac{R_1}{R_1 + R_2} \cdot V \cdot$ Suppose this resistor R_B takes a very large value such that the biasing current is held at a fixed amount. Any change in the value of β will then produce a proportionate change in the standing collector current. Where I_C and β are treated as mean values a change in current gain $\Delta\beta$ causes a shift in the collector current of $\Delta I_C = \frac{\Delta\beta}{\beta} \cdot I_C \cdot$ Naturally when R_B has a finite value there is some compensating action in the circuit; a change in biasing current alters the voltage drop across R_B and this to some extent offsets the chance in collector current. The contribution to ΔI_C will be less than $\frac{\Delta\beta}{\beta} \cdot I_C$ and it is no surprise to find that the actual contribution can be expressed in terms of the stability factor as

$$\Delta I_C = S \frac{\Delta\beta}{\beta^2} \cdot I_C \; \dots\dots\dots \quad (4)$$

When S is taken equal to β in the expression, which means infinite R_B as in the case that we have just considered, the correct answer $\Delta I_C = S \cdot \frac{\Delta\beta}{\beta} \cdot I_C$ emerges.

Both leakage and β variation are concerned about changes of base current and it is easy to visualize that as such they are parallel problems. It is a little difficult, however, to give a simple justification for equation (4); for one thing β, I_C and I_B are all changing at once and for another S itself involves β. A mathematical derivation for equation (4) is therefore given in the appendix. Certain assumptions have to be made, of course, in deriving such a simple expression and only first-order accuracy is to be expected from it; leakage current is ignored for example. It may have slipped by unnoticed, but, in deriving equation (2), we correspondingly tacitly assumed that β was constant.

To give some confidence in the use of equation (4) we will consider just one other extreme case. We will make R_E large and drive the stage from a voltage source so that the stability factor becomes unity. At a first glance at the problem, the stability would be thought to be so good that there could be no change in collector current with variation of β. In fact there is some because it is the emitter current which is held constant and the collector current must differ from the emitter current by I_B. ($I_C = I_E - I_B$). An increase in β causes a decrease in I_B and an increase again in I_C. The proportionate change in base current will be $\frac{\Delta\beta}{\beta} \cdot I_B$, that is to say $\frac{\Delta\beta}{\beta^2} \cdot I_C$ and it is to be expected that this should equal ΔI_C. This is indeed what equation (4) reduces to when S equals unity.

Establishing the Working Point:—In an amplifying stage, one of the basic decisions will be the choice of operating current around which the design is to be worked out. Just as in valve circuits the anode current will first be chosen, the requirements of the design will dictate an approximate value for the collector current. The chosen collector current will however be quite nominal because the tolerances, including the ones on β and the leakage current, will not allow the same current to result in every case (Circuits which employ preset components for setting up will still be liable to suffer from temperature drift and the inclusion of preset components anyway suggests an attitude of defeat).

It will be helpful if we can choose a mean value for I_C such that possible departures away from this nominal value range equally on either side of $\overline{I_C}$. Denoting this "design-centre current" by $\overline{I_C}$ and the maximum deviations by $\pm \Delta I_C$, the possible range of standing currents becomes $\overline{I_C} - \Delta I_C$ to $\overline{I_C} + \Delta I_C$. Whether such a range in collector current will be acceptable will be in a decision individual to each design. Here we need to discover how to design for $\overline{I_C}$ and how to calculate the contributions to ΔI_C.

The principle of working on a design centre is familiar enough, but with transistors its implementation proves to be a little tricky. Tolerances are not quoted plus and minus about mean values, but are given as extreme maximum and minimum limits. Although "typical values" often appear, these are not representative of any special mean of the limiting values. Reliance upon such "typical values" in design work can be a dangerous practice.

Let us consider how to choose a mean value of current gain, β, which results in equal shifts in I_C as β varies from β_{min} to $\bar{\beta}$ and from $\bar{\beta}$ to β_{max}. At (4) we have stated that $\Delta I_C = S.\frac{\Delta\beta}{\beta^2}.I_C$, so that for fixed values of S and I_C we may write $\Delta I_C \propto \frac{\Delta\beta}{\beta^2}$. It can then be deduced that the desired value of $\bar{\beta}$ is the reciprocal of the mean reciprocals of the extreme values β_{min} and β_{max}.

$\bar{\beta} = \frac{2\beta_{min}\,\beta_{max}}{\beta_{max}+\beta_{min}}$, and in terms of $\bar{\beta}$ and β_{min}

$$\Delta I_C = S\cdot\left(\frac{\bar{\beta}}{\beta_{min}} - 1\right)\cdot\frac{\bar{I}_C}{\bar{\beta}}$$

As an example suppose β_{min} (at the lowest working junction temperature) is 25 and β_{max} (at the highest temperature) is 100, then $\bar{\beta}$ works out to be 40. Had the extreme values been 25 and 125 respectively, then $\bar{\beta}$ would have worked out at 42. $\bar{\beta}$ lies closer to β_{min} than to β_{max} and is more critically dependent on the precise value of β_{min} than on β_{max}. That variations in β_{max} have the lesser effect on $\bar{\beta}$ is hardly surprising because at the higher current gains, the voltage dropped by the bias current across R_B is small and the changes are also correspondingly small. β_{min} is obviously the more valuable piece of information and this fact is often reflected in manufacturers' specifications where sometimes no data at all are given for β_{max}. A good transistor features a high β_{min}; and whilst a wide tolerance in β is undesirable, it is better for the tolerance to extend towards the higher values of β than towards the lower

To embrace leakage current in the design centre system is straightforward. Taking the minimum leakage to be zero, we can include half the maximum leakage as a contribution to ΔI_C and take the other half into account when setting up $\bar{I}_C$. The contribution to ΔI_C is $\frac{1}{2}.S.I_{CBO\ max}$ where $I_{CBO\ max}$ corresponds to the maximum operating junction temperature. From the two influences we have been considering in this article, the total ΔI_C is

$$\Delta I_C = S.\left(\frac{I_{CBO\ max}}{2} + \left(\frac{\bar{\beta}}{\beta_{min}} - 1\right).\frac{\bar{I}_C}{\bar{\beta}}\right) \quad .. \quad (5)$$

We can write down also a condition which will ensure that the appropriate value of $\bar{I}_C$ is obtained:

$$v = \bar{I}_C R_E + V_{BE} + \left(\frac{\bar{I}_C}{\bar{\beta}} - \frac{I_{CBO\ max}}{2}\right).R_B \quad .. \quad (6)$$

v is the potential that would appear at the junction of R_1 and R_2 if no current were drawn off at the base terminal; it is the unloaded voltage dropped across the lower resistor of the bias chain. Having said this, the derivation of equation (6) should be self-explanatory. It is assumed that the greater part of the leakage current, I_{CBO}, flows through R_B as it will do in a well stabilized circuit, and sufficient accuracy is obtained by writing the voltage drop from this cause as $\frac{I_{CBO\ max}}{2}.R_B$.

Equation (6) will not necessarily represent the normal biasing condition (say at room temperature with average transistors), it is merely some mean state. It is important, however, that the unloaded voltage, v, should be set up reasonably accurately so that the optimum use can be made of equation (5).

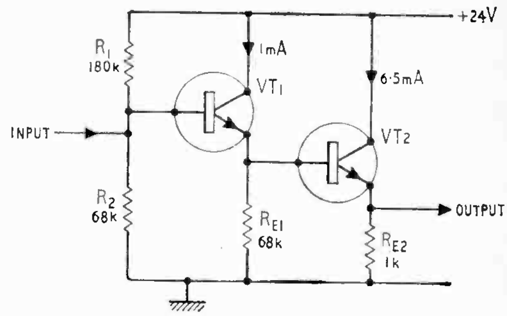


Fig. 5. Darlington pair with additional emitter resistor.

Design Procedure:—While it is not claimed that the design centre system will always be applicable, it is thought advisable to be aware of its implications. In many instances the designer may wish to discover, in an explanatory manner, where the limitations of his design lie. Equation (5) then displays at a glance the contributing factors to the d.c. stability, and shows what improvements might result when remedial measures are taken to improve the stability.

As an example of the method, Fig. 5 shows a Darlington pair.

The parallel combination of resistors R_1 and R_2 amounts to 49kΩ so that the ratio of R_B to R_E for the first transistor is 7 : 1. As previously stated we may take an approximate value for the stability factor equal to this ratio, 7. If β_{min} and β_{max} for VT1 are respectively 25 and 100, and if $I_{CBO\ max}$ is 70μA we can calculate that $S.\frac{I_{CBO\ max}}{2} = 225\mu A$ and that

$$S\left(\frac{\bar{\beta}}{\beta_{min}} - 1\right)\frac{\bar{I}_C}{\bar{\beta}} = 7.\left(\frac{40}{25} - 1\right).\frac{10^3}{40} = 105\ \mu A.$$

The sum of these gives $\Delta I_C = 330\ \mu A$. Although nominally the standing current is intended to be 1mA, the possible variations are from 0.67mA to 1.33mA. The unloaded voltage, v, at the top of R_1 and R_2 should be from equation (6), $6.8 + 0.25 + (0.025 - 0.035) \times 49$ which equals 6.56 volts. For a nominal 24 volts supply, values of 180kΩ and 68kΩ for R_1 and R_2 respectively will supply just this voltage.

It will be noticed that where we have calculated in equation (6) the quantity in brackets $\left(\frac{\bar{I}_C}{\bar{\beta}} - \frac{I_{CBO\ max}}{2}\right)$ representing the total bias current, this has taken a negative sign. The leakage current into the base region is more than enough to bias the stage and some of it has actually to be drawn out through the base lead. The base current is composed of the current in through the base terminal together with leakage from the collector junction which enters into the base-emitter junction internally. We must therefore allow for a current supply in either direction through R_B.

Now turning our attention to the second stage of Fig. 5 we can foresee a problem. At least there would be a problem in the absence of the resistor R_{E1} in the circuit. In the elementary form of the Darlington pair, there is no resistor R_{E1}, and the bias current for the second transistor is supplied directly by the emitter of the first. VT1 emitter will unfortunately only supply current in one direction so that the d.c. stability, whilst being good at low temperatures, can suddenly go completely when the leakage current of the second transistor exceeds a certain value. One has, naturally, the possibility of employing a silicon transistor with its inherent low leakage for VT2; but the other alternative must be to fit this resistor R_{E1}. Again where the amplifier of Fig. 5 is required to handle fast pulses, trouble can arise in connection with the input capacitance of VT2 and the rectifying junction at the emitter of VT1. An emitter resistor R_{E1} is therefore advisable in the Darlington pair when it is used in pulse work or in wideband applications.

We have dealt with the stability of the first transistor of the circuit of Fig. 5. Provided that VT1 always remains conducting, the second transistor sees an impedance at its base of $\frac{49k\Omega}{\beta_1}$ which is much lower than the value of R_{E1}. The S factor for VT2 will therefore be close to unity and the stability be very good. The change in the working point of VT1 will however be directly reflected into VT2 because the emitter potential of VT2 must closely follow the emitter potential of VT1. The proportionate change in standing current for VT2 will then be of the order of ± 2.5 mA in 6.5 mA nominal.

APPENDIX

Here is a mathematical derivation of equation (4) for finding the change in collector current resultant upon a change in the current gain.

Fundamentally the potentiometer method of bias stabilization relies on the balance of a voltage drop across $R_E + r_e$ to that across the base resistor R_B. For any change in the *status quo* the following condition must be satisfied: (Change in emitter current) × $(R_E + r_e)$ = − (Change in base current) × R_B .... (1)

Since the base current is by definition $\frac{I_C}{\beta}$, the change ΔI_B is given by the partial differential $\frac{\Delta I_C}{\beta} - \frac{\Delta\beta}{\beta^2} I_C$, for small changes. $\frac{\Delta I_C}{\beta}$ is the result of a change in I_C with fixed β, and the $\frac{\Delta\beta_C I}{\beta^2}$ term corresponds to fixed I_C and changing β. The approximation "for small changes" means that only first order accuracy will be obtained, but this has already been accepted in the text of the article.

To find ΔI_E we must write $\Delta I_E = \Delta I_C + \Delta I_B$ and substitute $\Delta I_B = \frac{\Delta I_C}{\beta} - \frac{\Delta\beta}{\beta^2}.I_C$ into this identity.

Giving $\Delta I_E = \Delta I_C\,(1 + \frac{1}{\beta}) - \frac{\Delta\beta}{\beta^2}.I_C$.

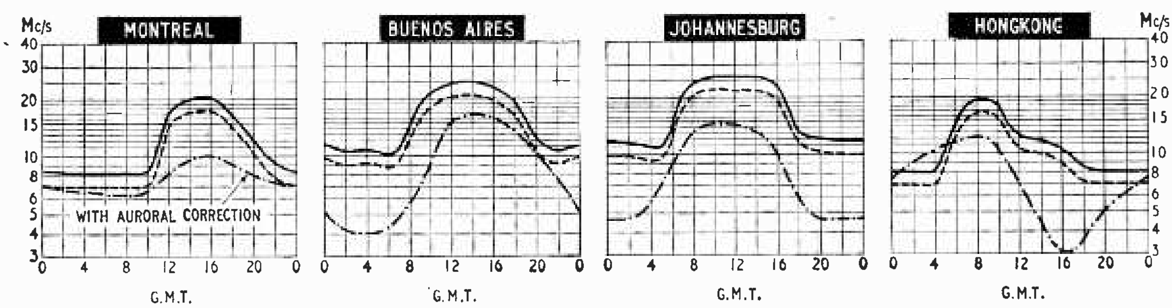
We now have expressions for ΔI_B and ΔI_E both in terms of ΔI_C and Δ_β, and can substitute both of these in our original condition at (1).

At the same time neglecting unity in comparison to β we obtain:

$$\Delta I_C\left\{r_e + R_E + \frac{R_B}{\beta}\right\} = \frac{\Delta\beta}{\beta^2}.I_C\,(r_e + R_E + R_B)$$

But as the stability factor S is used to denote $\frac{r_e + R_E + R_B}{r_e + R_E + \frac{R_B}{\beta}}$, we have in fact $\Delta I_C = S.\frac{\Delta\beta}{\beta^2}.I_C$.

H. F. PREDICTIONS—NOVEMBER



MEDIAN STANDARD MUF
OPTIMUM TRAFFIC FREQUENCY
LOWEST USABLE HF

The prediction curves show the median standard MUF, optimum traffic frequency and the lowest usable frequency (LUF) for reception in this country. Unlike the standard MUF, the LUF is closely dependent upon such factors as transmitter power, aerials, and the type of modulation. The LUF curves shown are those drawn by Cable and Wireless Ltd. for commercial telegraphy and assume the use of transmitter power of several kilowatts and aerials of the rhombic type.

The higher daytime MUFs, characteristic of the winter months, are now becoming apparent for circuits predominantly in the Northern hemisphere. The Northern Auroral Zone passes through Alaska, Hudson Bay, Iceland and Northern Norway and radio paths passing through this zone are subject to additional absorption, for which a correction is made in the calculation of the LUF.

MANUFACTURERS' PRODUCTS

NEW ELECTRONIC EQUIPMENT AND ACCESSORIES

Ultrasensitive Current Meters

A NEW series of ultrasensitive d.c. current meters has been announced by the Greibach Instrument Corporation, of New York. These meters which provide full-scale deflection sensitivities of the order of 200 to 300 nano-amps, without the use of amplifiers or other additional circuitry, incorporate the Greibach Bifilar suspension movement. This provides good stability, repeatability of ± 0.1% and ruggedness combined with low voltage drop. Indication is by means of a light-beam pointer operating over a six-inch scale. Single and multi-range instruments are among the models available, the latter having as many as 23 ranges providing complete coverage of the microampere, milliampere and low ampere current bands. All the models in the 700 range incorporate overload protecton circuits and are available for either for panel mounting or as complete portable instruments. They are distributed in the United Kingdom by the Instruments Division of Claude Lyons, of 76 Old Hall Street, Liverpool 3.

14WW 309 for further details

Instrument Control Knobs

CONTEMPORARY matt-black-finished control knobs in a matching range of three different sizes, are now available from A. F. Bulgin & Co. Ltd., of Bye-Pass Road, Barking, Essex. Moulded from Bakelite, these pointer-knobs are fitted with satin-aluminium flat disc inserts, or spin discs to order. They all feature the Bulgin locked insert, which is positively locked and keyed to the moulding and locked to component shafts by means of a 4 B.A. grub screw (or an Allen type on request). All three are suitable for use on quarter-inch shafts with or without flats.

14WW 310 for further details

Oscilloscope Calibrator

A PORTABLE oscilloscope calibrator, which provides in one unit all the waveforms necessary for the alignment of Telequipment and a number of other makes of scope, has been introduced by Telequipment Ltd., of 313 Chase Road, Southgate, London, N.14. It is a portable instrument with four separate outputs which provide a square wave switched at either 100 kc/s or 1 Mc/s; a square wave switched at either 10 kc/s or 1 kc/s; time marker pulses at switched rates of 1 Mc/s, 100 kc/s, 10 kc/s, 1 kc/s 50 c/s and a timing comb (negative-going with respect to earth); and a non-interlaced television waveform (switched ±ve, with approximately 200 lines, with a combined sync and video amplitude of one volt peak to peak). The calibrator known as the Type C1, incorporates a crystal-controlled source with an accuracy of 0.2% and has an overall power consumption of around 90 watts. It weighs 24 lb and measures $13 \times 6\frac{1}{2} \times 13$ in.

14WW 311 for further details

Electrolytic Range Extended

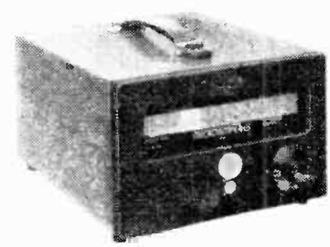
TWO new can sizes have been added to the Mullard Type C426 range of electrolytic capacitors which now runs to 48 types housed in six can sizes. Capacitor values range from 0.64 μF to 500 μF, with working voltages from 2.5 V to 6.4 V. At the same time as introducing the two new can sizes, the manufacturers announce that a higher voltage capacity product and volume efficiency have been obtained and also that these techniques will be used in a new range of electrolytics designated C437.

Thirty-two types comprise the new range, housed, according to their capacity and voltage rating, in cans of four different sizes. Capacitance values are from 64 μF to 4,000 μF with working voltages ranging from 2.5 to 64 volts. The revised C426 range and the new C437 range will be available from 1st January.

14WW 312 for further details

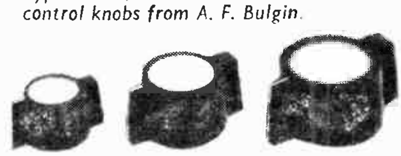
Tunable V.H.F. Amplifier

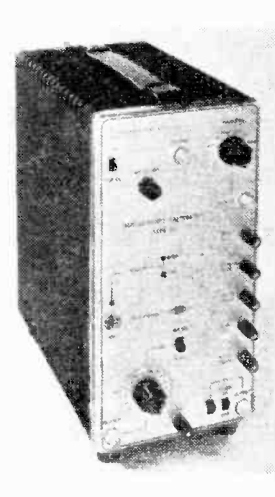
A MEASURING instrument which offers more possibilities than its name suggests is the Rohde & Schwarz Type ASV tunable v.h.f. amplifier. It covers the whole of the v.h.f. band, that is to say 30 to 300 Mc/s, and a few of the uses it can be put to are: v.h.f. signal generator, audio signal generator, frequency meter, frequency converter, a.m. and f.m. demodulator, a.m. modulator and of course a v.h.f. receiver; with the output on telephone jacks. One of the instrument's accessories is a mixer and harmonic generator, which can be used for frequency multiplication as well as to provide a wide range of harmonics below, in and above the v.h.f. band of frequencies. A wide-band three-stage tunable amplifier is used in the v.h.f. section and operates as a linear



Greibach Model 700 microammeter.

Type K 482, K 483 and K 484 instrument control knobs from A. F. Bulgin.





Telequipment oscilloscope calibrator.

amplifier on input voltages of up to 10 mV, whilst higher input voltages are progressively limited. In operation as a generator, the first amplifier stage functions as an oscillator while the other two stages provide for both amplification and buffering. This v.h.f. output can then be modulated internally or externally. A 1,000 c/s generator delivers the internal modulation voltage, which is also available at an output after amplification in the final modulation stage. The instrument will operate from any 115/ 125/ 220/ 235 volt 50 to 60 c/s supply, and is available in the United Kingdom from Aveley Electric Ltd., of South Ockendon, Essex.

14WW 313 for further details

Low-frequency Wave and Spectrum Analyser

COVERING a frequency range from 1 c/s to 5,000 c/s, the new Quan-Tech 304 spectrum analyser can be electronically swept through the frequency spectrum at three different speeds with bandwidths of 1 c/s, 10 c/s or 100 c/s. The speeds are 18 seconds, 3 minutes and 30 minutes per sweep. A d.c. output, proportional to the frequency at which the instrument is tuned, of 1 mV/cycle, is provided for plotting the spectrum of the input signal on an X-Y pen recorder or an oscilloscope. Sensitivity is quoted as 30 μV to 100 volts full scale with an accuracy of $\pm 5\%$.

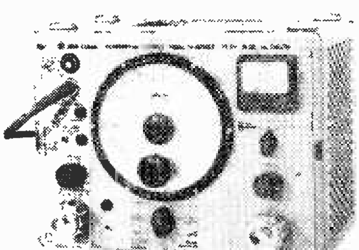
This instrument may also be used as a wave analyser by manually tuning the frequency range. It is available in the United Kingdom through Livingston Laboratories Ltd., of 31 Camden Road, London, N.W.1.

14WW 314 for further details

Metal-to-Ceramic Sealed Components

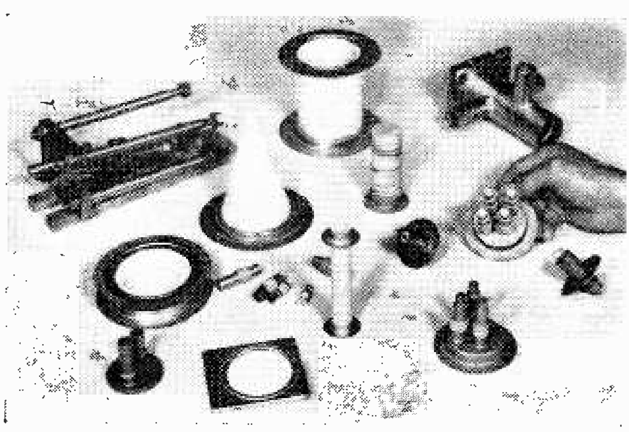
A WIDE range of metal-to-ceramic sealed components is now being made available by EMI Electronics Ltd., of Hayes, Middx. These cover a broad field and include insulators for various parts of microwave valves, high-voltage terminals for plasma research projects and waveguide windows of various shapes and sizes. The company also offers a service for the design and manufacture of special components and will undertake the sealing of customers' parts. A number of metallizing techniques have been developed and the company believes that it can produce a satisfactory bond to the required metal for almost any type of ceramic, bearing in mind that it is necessary to match the thermal expansion of the materials in order to avoid residual strains. In addition to being able to withstand very high temperatures, these components are claimed to possess excellent resistance to thermal shock.

14WW 315 for further details



Rohde & Schwarz 30 to 300 Mc/s tunable v.h.f. amplifier.

Quan-Tech Model 304 low-frequency wave and spectrum analyser.



A selection of metal-to-ceramic components manufactured by EMI Electronics Ltd.

Helical Potentiometer

THREE-QUARTER-INCH diameter helical potentiometers manufactured by Reliance Controls Ltd., of Relcon Works, Sutherland Road, Walthamstow, London, E.17, are now available in five- and ten-turn versions. Various improvements have been made to the series, known as the HEL 07, and the end stop torque has been increased to 5 lb/in. Both the five- and the ten-turn versions are in production with standard linearities of $\pm 1\%$ or $\pm 0.5\%$; or $\pm 0.25\%$ to special order. The resistance range covered is from 10 ohms to 50 kΩ, and a slide wire version is available for lower values up to 1.5 ohms. A three-turn version is to be added to the series later in the year and will incorporate the improvements of the other two, which include a new method of bush mounting and a re-designed wiper with a floating spring action.

14WW 316 for further details

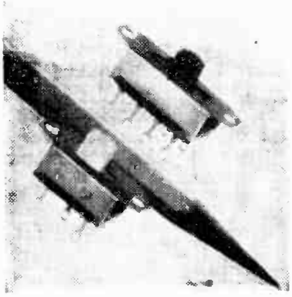
Pulse Generator

A GENERAL-PURPOSE pulse generator, known as the Model 6613, is the first of a new series of British-made test equipment being manufactured by Texas Instruments Ltd., of Bedford. This unit provides coincident positive and negative output pulses at a rate which can be determined either manually or by the internal clock generator or an external signal source. The repetition rate of the clock generator can be varied, in six decade stepped ranges, from 15 c/s to 15 Mc/s. Output amplitude can be varied from zero to 10 volts (into 50 ohms) and a protection circuit is provided against overload. The pulse width is variable up to 90% of the duty cycle. Printed circuit board techniques are used and the instrument measures only $8\frac{1}{2} \times 8\frac{1}{2} \times 12$ in and weighs 10 lb.

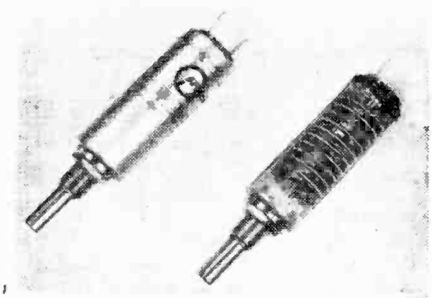
14WW 317 for further details

High-voltage Silicon N-P-N Transistors

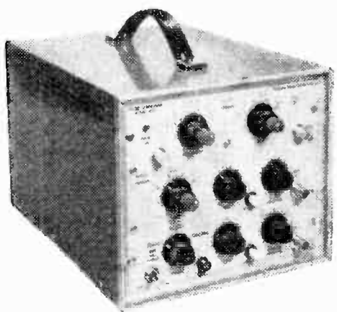
FOUR new triple-diffused, high-voltage silicon transistors for low-power amplifier, oscillator and low-current switching applications have been announced by R.C.A. Great Britain Ltd., of Lincoln Way, Wind-



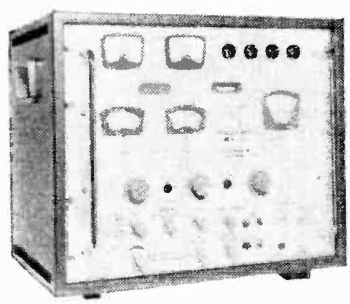
Miniature slide switches from Highland Electronics Ltd.



Helical potentiometers from Reliance Controls Ltd. The one on the right has been cut away to show its construction.



General purpose pulse generator from Texas Instruments Ltd.



Type 331/C frequency meter and generator from J.A.C. Electronics Ltd.

mill Road, Sunbury-on-Thames, Middx. They all have low saturation voltage and fast response characteristics and thus should find many uses in the high-voltage fields such as inverters, regulators, differential and linear amplifiers, etc. Two of the new units, the 2N3439 and the 40255, have a one amp rating with a maximum collector-to-base voltage of 450 V, whilst the other two units, the 2N3440 and the 40256, have a one amp rating with a collector-to-base maximum of 300 volts. The maximum collector-to-emitter voltage for the first two is quoted at 350 volts and a hundred volts lower for the latter two, with an emitter-to-base voltage of 7 V maximum for all four.

The Type 2N3439 and the 2N3340 are housed in standard JEDEC Type TO-5 cases, and the other two are housed in the new diamond flange version of the TO-5 case, which feature increased heat dissipation.

14WW 318 for further details

Miniature Slide Switches

APPROXIMATELY one third the volume of standard slide switches is claimed by Highland Electronics for their new range of single and double-pole miniature slide switches. Two and three position units with detent or spring return mechanisms are included in the new range which is obtainable from 26-28 Underwood Street, London, N.1. The contact rating for these units is quoted at 500 mA at 120 volts, and the minimum insulation resistance is given as 100 MΩ after 48 hours in an environment of 95% humidity at 50°C. The dielectric strength is tested at 900 volts a.c. for one minute between contacts and frame.

14WW 319 for further details

U.H.F. Transistor with A.G.C. Characteristics

GOOD forward a.g.c. characteristics are claimed for the Type AGC/1 silicon planar amplifier transistor from S. G. S. Fairchild, of 23 Stonefield Way, Ruislip, Middx. At current levels up to about 3 mA the characteristics of this unit are similar to the Type 2N918 u.h.f. amplifier transistor, but when the collector current is further increased linear reduction suddenly begins and a reduction in gain of 30 dB is obtained with 9 mA on the collector. This should, therefore, enable i.f. and r.f. amplifiers to be built without the previous limitations imposed such as current starvation or reverse a.g.c. The transistor can be operated with a constant collector voltage and this, of course, keeps the collector capacity constant and avoids detuning of the circuit at different signal levels.

14WW 320 for further details

Frequency Meter and Generator

THE frequency meter and generator Type T.D.1. from J.A.C. Electronics Ltd., (formerly Telemac-Southern Ltd.) of Station Estate, Blackwater, Camberley, Surrey, is now replaced by a new model, known as the 331/C, of improved layout and design. The general electrical characteristics of the instrument remain the same, hence the range as a frequency meter is 10 kc/s to 3,000 Mc/s and as a frequency source from zero to 3,000 Mc/s. Reference standard stability is quoted as ± 5 parts in 10^{10} at constant ambient temperature with better than ± 1 part in 10^{9} drift per day. Whereas the T.D.1. had a frequency counter as an optional extra, this is now an optional deletion and provides a reading accuracy of ± 1 cycle as against ± 10c/s. Built-in facilities include automatic harmonic identification, direct measurement of drift about any set frequency, provision of 37,000 standard frequencies (at 10 kc/s intervals) phase locked to crystal accuracy, and the facility of amplitude modulating the output at 1,000 kc/s. Another feature of this instrument is that it has a standard 19in front panel.

14WW 321 for further details

INFORMATION SERVICE FOR PROFESSIONAL READERS

To expedite requests for further information on products appearing in the editorial and advertisement pages of *Wireless World* each month, a sheet of reader service cards is included in this issue. The cards will be found between advertisement pages 16 and 19.

We invite readers to make use of these cards for all inquiries dealing with specific products. Many editorial items and all advertisements are coded with a number, prefixed by 14WW, and it is then necessary only to enter the number(s) on the card.

Readers will appreciate the advantage of being able to fold out the sheet of cards, enabling them to make entries while studying the editorial and advertisement pages.

Postage is free in the U.K., but cards must be stamped if posted overseas. This service will enable professional readers to obtain the additional information they require quickly and easily.

RECENT TECHNICAL DEVELOPMENTS

Laser Developments from Bell

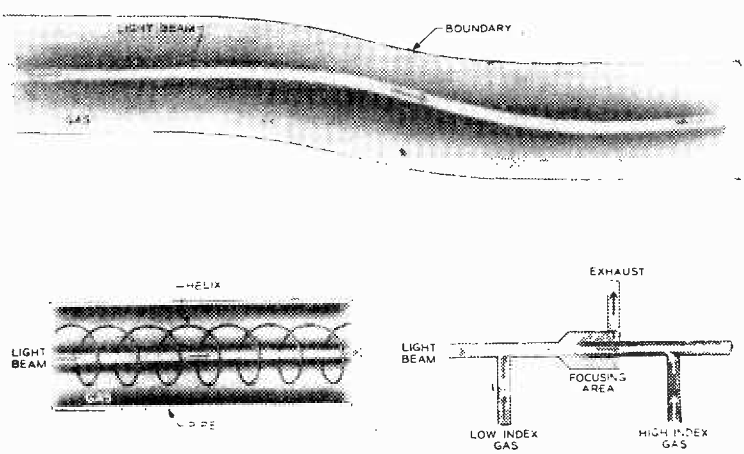
The highest continuous power so far obtained from gas lasers is claimed by Bell Telephone Laboratories. A power output of 1 watt, a tenfold increase over previous gas lasers, has been measured from a helium-neon laser at 6,328 angstroms.

Low-loss mirrors and a tube length of 5.5 metres contribute to the relatively high output of the laser, which requires an input power of 500 watts.

Experiments designed to investigate the problems of communicating by light beams have revealed that light power fluctuates randomly in all kinds of weather, due to minute inhomogeneities in the atmosphere. In rain, attenuation reaches 30 dB; in fog and snow it can be as high as 80 dB. As a result it may be necessary to transmit the light beams through pipes and recently gas lenses have been invented which will guide the light beams in pipes, having the advantages that losses are much less than conventional optical components and curved transmission lines are more feasible.

One lens structure consists of a gas-filled pipe inside which is a helix that heats the gas to a few degrees above the pipe. This causes the index of refraction to rise in the centre of the pipe, and focuses the beam. Focal length is adjusted by the helix temperature.

Our front cover shows two 10 mW lasers in action, one being at a distance of $1\frac{1}{2}$ miles. The 10-point star is only indirectly a quality of the laser light; it is caused by diffraction at the edges of the camera lens iris!



Two types of Bell Telephone Laboratories lenses that use differences in the refractive indices of gas to focus and guide light beams. The diagram on top shows how the gas focusing principle may be used to guide a laser beam through a pipe.

Electrothermal Recording on Plastic Films

It has long been known that images can be formed by the interaction of electrostatic charges with deformable insulating layers. The Eidophor projection television system is a good example. Permanent recording on solid dielectrics has also been reported.

Further developments in the art were described in *R.C.A. Review* (June 1964) where N. E. Wolff describes a method for recording on photoconductive thermoplastics layers. Corona charging with a line or dot screen is necessary to obtain grey shades between black and white and development is accomplished by heating the film to the softening temperature at which point the electrostatic and surface tension forces combine to form undulations which set on cooling. The image is then viewed through the medium of a schlieren projection system.

In the same issue, F. H. Nicholl describes a new surface phenomenon which has been applied to image recording. When a surface layer ranging in thickness from 10Å-1000Å depending on the material, is added to a thermoplastic (polystyrene for instance) corona charging and subsequent heating results in deformation.

Half-tone effects can be obtained without the use of a screen and it is possible to view images by reflected or transmitted light using either conventional or schlieren projection.

Infra-red Fault Detection

A fast-scanning infra-red microscope that can locate failures before they occur in electronic space devices without long, complex tests is being developed by the Raytheon Company under a NASA contract.

Because virtually all potential causes of failure in a semiconductor device or integrated circuit result in abnormal thermal behaviour, use of the so-called Performance Analyzer (which can detect minute amounts of heat) is believed to be one of the most practical ways of testing and predicting the reliability of micro-circuits.

The infra-red detector measures the heat dissipated when a component or circuit is operating. By comparing the heat pattern with a standard, the engineer can determine performance and discover the presence of hidden defects which might lead to failures. It is intended that later models will use a computer memory bank to store test results so that reliability of parts can then be checked automatically.

Transistor with Million Megohm Input Resistance

The thin-film transistor described by P. K. Weimer in *Proc. I.R.E. June, 1963* had an input resistance of 1 MΩ and an input capacitance of 50 pF and oscillation was obtained up to 17 Mc/s. It consisted of a glass substrate on which a gold "source" and "drain" were deposited and semiconducting cadmium sulphide deposited between and on the source and drain. A gold "gate" was insulated from the sulphide by a thin layer of silicon monoxide.

Mullard Ltd. announce progress in

this field with M.O.S.T.—metal oxide silicon transistor—a somewhat unfortunate title since it can be construed that the transistor contains a metal oxide. A p-type silicon substrate is used (this is connected to the TO-5 envelope and can be used as a modulation electrode) on which the closely spaced n^+ source and drain are diffused. A thin silicon oxide layer is then grown on the substrate and acts as a dielectric between this and the evaporated aluminium gate. The device, designated 95BFY, is claimed to have no storage effect and to have the following characteristics:—

Input resistance	$10^{12}\,\Omega$
Input capacitance	4 pF
Output resistance	40 kΩ
Output capacitance	3 pF
g_m 1mA/V at I_d=3mA	

This data may differ from later samples as development progresses, but fuller details and samples can be obtained from the manufacturers for evaluation purposes.

High Energy Discharges for Crushing

The discharging of energies of the order of a megajoule stored in capacitors charged to 40 kV for the production of plasma in thermonuclear reactions was reported in an article in *Wireless World, August 1964*. Inductances as low as 3 μH and a total lead resistance of 0.2 milliohm were achieved. The A.E.R.E. at Harwell have used similar methods to produce underwater plasma giving shock waves of an intensity approaching that of chemical explosives. The shock waves have been used for crushing materials such as gravel, glass, carbides and oxides which are difficult to crush by conventional methods. To obtain maximum effect a short discharge time and low inductance is required. With a capacitor of 0.05 μF charged to 50 kV through 50 MΩ at 1 mA, and an inductance of 0.1 μH (giving a ringing at 3 Mc/s) discharge times of 0.075 μsec to peak and 1.5 μsec total have been obtained. Tempered silver steel electrodes ground to a point have been found most satisfactory and wear is about 1 mm/1000 pulses. (The number of pulses required depends on the material and particle size required: for instance 100 pulses at 35 kV on pellets of UO_2 10 mm × 10 mm reduced size 50% below 0.4 mm). Crushing efficiencies are 30 times higher than those obtained by previous workers on comparable electrohydraulic systems with inductances of 30 μH.

NOVEMBER MEETINGS

Tickets are required for some meetings: readers are advised, therefore to communicate with the society concerned.

LONDON

2nd. I.E.E. & R.Ae.Soc.—"N.A.S.A. research and technology programmes in space navigation" by J. I. Kanter at 5.30 at Savoy Place, W.C.2.

3rd. I.E.E.—"The problem of vestigial-sideband transmission and reception" by W. Wharton and B. J. Rogers at 5.30 at Savoy Place, W.C.2.

4th. I.E.E.—"A push-pull parallel aerial system for navigation and application to communication problems with simple methods of side-lobe cancellation" by Prof. E. O. Willoughby at 5.30 at Savoy Place, W.C.2.

4th. I.E.R.E.—"U.H.F. television reception over long-distance paths" by B. W. Osborne at 6.0 at the London School of Hygiene, Keppel St., W.C.1.

4th. B.K.S.—Symposium on "Underwater cinematography and television" at 7.30 at Shell-Mex House, Strand.

6th. Television Soc.—"Factors affecting the acceptability of colour reproduction" by Dr. R. W. G. Hunt at 7.0 at I.T.A., 70 Brompton Rd., S.W.3.

10th. I.E.E.—"Computer-aided study of character recognition" by J. A. Weaver at 5.30 at Savoy Place, W.C.2.

11th. I.E.R.E.—Papers on "Electromechanical filters" at 6.0 at the London School of Hygiene, Keppel St., W.C.1.

12th. I.E.E.—"Pressure sensitive effects in semiconductors" by K. Preece, P. Lundberg, P. Selway and Dr. V. G. Tull at 5.30 at Savoy Place, W.C.2.

12th. Radar & Electronics Assoc.—"Televising the Tokyo Olympics" by L. F. Mathews at 7.0 at the R.S.A., John Adam Street, W.C.2.

17th Soc. of Relay Engrs.—Papers on transistor repeaters for h.f. and v.h.f. relay systems, by W. Dougharty and A. Isaacs at 2.30 at 21 Bloomsbury Street, W.C.1.

18th. I.E.R.E.—"An air surveillance radar system" by R. L. Burr and J. Flounders at 6.0 at the London School of Hygiene, Keppel Street, W.C.1.

18th. B.K.S.—"Radiophonics" by C. Booker at 7.30 at Central Office of Information, Hercules Road, S.E.1.

19th. Television Soc.—"A British video tape recorder" by J. L. E. Baldwin at 7.0 at I.T.A., 70 Brompton Road, S.W.3.

23rd. I.E.E.—"Loop gain and return difference of transistor amplifiers" by R. F. Hoskins at 5.30 at Savoy Place, W.C.2.

25th. I.E.R.E. & I.E.E.—Discussion on "Symbols for logic circuits" at 6.0 at the London School of Hygiene, Keppel Street, W.C.1.

25th. Inst. Eng'g. Inspection.—"Inspection in the radio and electronics industry" by J. C. McCullagh at 7.30 at South-West Essex Technical College, Forest Road, E.17.

25th. B.K.S.—"Deformation of gramophone record grooves" by D. A. Barlow at 7.30 at Shell-Mex House, Strand, W.C.2.

27th. R.S.G.B.—"Moonbounce" by P. K. Blair at 6.30 at I.E.E. Savoy Place, W.C.2.

30th. I.E.E. & I.E.R.E.—Discussion on "Medical applications of lasers" at 5.30 at Savoy Place, W.C.2.

ABERDEEN

11th. I.E.E.—"Loudspeakers" by K. F. Russell at 7.30 at Robert Gordon's Technical College.

BASINGSTOKE

26th. I.E.R.E.—"Colour television transmission systems" by W. Wharton at 7.30 at Technical College.

BIRMINGHAM

25th. I.E.E.—"Television in the service of technical education" by J. Scupham at 6.15 at the College of Advanced Technology, Gosta Green.

25th. Television Soc.—"Transcoding from SECAM to N.T.S.C." by H. Steele at 7.0 at the College of Advanced Technology.

BRISTOL

18th. I.E.R.E. & I. Prod. Engrs.—"Quality and reliability" by D. J. Hewitt at 7.0 at the University Engineering Laboratories.

26th. I.E.E.—Faraday Lecture on "Colour television" by F. C. McLean at 6.0 at Electricity House, Colston Avenue.

CAMBRIDGE

3rd. I.E.E. & I.E.R.E.—"Integrated circuits" by J. S. Walker at 8.0 at the Engineering Lab., Trumpington St.

26th. I.E.E.—"Through communications to electronics" by G. G. Gouriet at 8.0 at the Engineering Laboratory, Trumpington Street.

CARDIFF

11th. I.E.R.E.—"Automobile electronics" by R. A. Evans at 6.30 at the College of Advanced Technology.

CHELMSFORD

30th. I.E.E.—"The digital computer—how it works and its impact on the solution of engineering problems" by P. S. Brandon at 6.30 at the Lion and Lamb Hotel.

CHIPPENHAM

10th. I.E.E.—"Connections in electronic circuits" by G. W. A. Dummer at 6.0 at the Concert Hall, Westinghouse Brake & Signal Co.

DUNDEE

12th. I.E.E.—"Loudspeakers" by K. F. Russell at 7.0 at the Electrical Engineering Dept., Queen's College.

EDINBURGH

3rd. I.E.E.—"Lasers, what they are and what they do" by N. Forbes at 6.0 at the Carlton Hotel, North Bridge.

11th. I.E.R.E. & I.E.E.—"Signal processing filters and networks" by D. J. H. MacLean at 7.0 at the University, Drummond Street.

EVESHAM

26th. I.E.R.E.—"U.H.F. broadcasting and BBC-2" by D. B. Weigall at 7.0 at the B.B.C. Club.

GLASGOW

11th. I.E.E.—"Electronic instruments and medicine" by I. W. Stevenson at 6.0 at 39 Elmbank Crescent.

12th. I.E.R.E. & I.E.E.—"Signal processing filters and networks" by D. J. H. MacLean at 7.0 at 39 Elmbank Crescent.

LEEDS

24th. I.E.E.—"Newton and Heaviside in control" by G. G. Gouriet at 6.30 at the Electrical Engineering Dept., the University.

LEICESTER

10th. Television Soc.—"Outside broadcasting" by Barry Edgar at 7.15 at Vaughan College, St. Nicholas Street.

12th. I.E.R.E.—"Education and training for professional radio and electronic engineers" by A. J. Kenward at 6.30 at the University.

LETCHWORTH

4th. Inst. Eng'g. Inspection.—"Applications of electronics in metrology" by B. S. Pearn at 7.15 at the College of Technology.

LIVERPOOL

9th. I.E.E.—"Silicon diodes" by R. A. Reid at 6.30 at the Royal Institution, Colquitt Street.

MANCHESTER

4th. I.E.R.E. & I.E.E.—"Television receiving aerials at u.h.f." by C. F. Whitbread at 6.0 at Reynold Building, College of Science and Technology.

17th. I.E.E.—"Cybernetics" by Prof. J. C. West and Dr. J. L. Douce at 6.15 at the College of Science and Technology.

NEWPORT, I.o.W.

27th. I.E.E.—"The present state of colour television" by S. N. Watson at 6.30 at the Technical College.

PLYMOUTH

19th. I.E.R.E. & I.E.E.—"Ballistic missile early warning system" by Brian Batt at 7.0 at the College of Technology.

PORTSMOUTH

4th. I.E.R.E.—"The application of radar and other electronic techniques to meteorology" by W. A. Grinstead and A. P. Tuthill at 6.30 at Highbury Technical College, Cosham.

PRESTON

11th. I.E.E.—"The results of tests at Goonhilly with the experimental earth satellites Telstar and Relay" by F. J. D. Taylor, W. J. Bray and R. M. White at 7.30 at the Harris College of Further Education.

SHEFFIELD

4th. I.E.R.E. & I.E.E.—"Environmental testing of electronic equipment" by K. C. Davies at 6.30 at Dept. of Electrical Engineering, the University.

SOUTHAMPTON

10th. I.E.E.—"A microcircuit logic system" by Dr. D. J. Truslove at 6.30 at the University.

13th. Inst. Electronics.—"Applications of threshold logic" by Lt. W. Sudweeks, R.N., at 7.0 at Grosvenor House, Cumberland Place.

SWANSEA

24th. I.E.E.—Faraday Lecture on "Colour television" by F. C. McLean at 6.30 at Brangwyn Hall.

WHITBY

10th. I.E.E.—"Engineering and scientific aspects of the Canadian ionospheric satellite" by E. D. R. Shearman at 7.0 at Botham's Cafe, Skinner Street.