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PRINCIPAL CONTENTS

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Announcements

Price

IT is with great regret that the publishers of this Journal have been compelled to increase its price from 1s. 6d. to 2s. in order to meet the rising costs of printing and production.

Naturally, we do not dwell on the difficulties under which we are working at the present time, but we would remind readers that the printing and allied trades have lost a large number of skilled operators and this combined with the difficulties of supply and distribution adds considerably to the troubles of producing a technical journal in war-time.

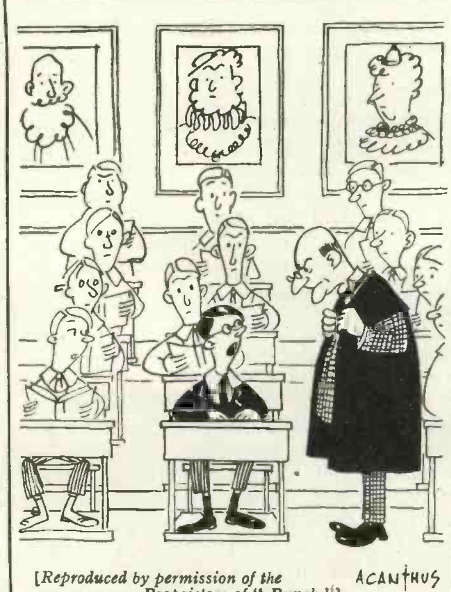
Commencing with the February issue, the number of pages allocated to editorial matter will be increased, which will, we think, more than compensate our readers for the slight additional cost of the journal.

Binding

Again, the supply of binding material and the shortage of labour present problems in the binding of the 1941 issues. It is possible that some readers would prefer to bind the journal under its new title in a separate volume from *Electronics and Television*, but the number of parts issued makes this an uneconomical proposition if the volume is terminated at December, 1941.

Accordingly, it has been decided to continue the run of the present volume into 1942 to enable a uniform size cover to be used. For those who wish to bind the whole of the 1941 issues in one volume, irrespective of the change in make-up, a limited number of brown binding cases, uniform with those of previous years, are available at a cost of 2s. post free.

Readers are asked to make their own arrangements for the actual binding of the volumes, as the publishers regret that they are unable to undertake the binding at the present time. Inquiries for binding cases should be addressed to the Circulation Dept.



[*Reproduced by permission of the Proprietors of "Punch"*] ACANTHUS

"*I've given all my books and papers to Lord Beaverbrook.*"

HAVE YOU ?

There is no need to be so drastic in clearing out books and papers, but the fact remains that there are literally tons of waste paper in old books and catalogues which could be used as a direct contribution to the war effort.

Please have a clear-out before the end of the year and ask your local council to collect the paper salvage—they will be quite willing to do so.

Data Sheets

As mentioned in an earlier issue, binding covers for Data Sheets are being prepared and will be ready shortly.

The cover is in stiff blue paper with a back carrying a number of strings through which the individual sheets can be slipped. The case opens flat and allows the data sheet to be easily read. The capacity is 24 sheets, and the cost is 2s. 6d. post free from this office.

For those readers who have not been able to obtain a complete set of the issues containing the Data Sheets a small number of binders can be supplied complete with sheets for 4s. 6d. post free.

Owing to the paper shortage and the impossibility of reprinting sheets at the present time, back copies of the Data Sheets can only be supplied to readers taking out a regular subscription to the Journal.

New Year Greetings

The Editor takes this opportunity of wishing all his readers a better New Year with the prospect of even brighter ones to come.

Television Picture Storage

A New Method of Electronic Storage Excitation of Television Receiving Screens

by A. H. ROSENTHAL, Ph.D., F.R.A.S.

At the beginning of 1940 a description of the Skiatron—a new development in television picture projection—was given in "Electronics and Television." Dr. Rosenthal, the inventor of the method, who was formerly associated with the Scophony Laboratories in this country, is now a consulting physicist in New York.

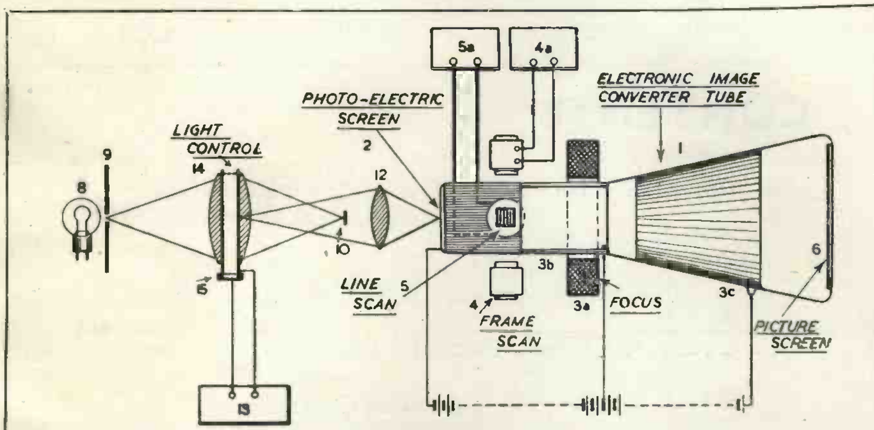


Fig. 3. A diagram of the proposed new method of picture storage.

IN conventional cathode-ray tubes for television reception the picture screen* is scanned by a cathode-ray spot of an area equal to the approximate size of a picture element.

Each elemental picture area of the screen is excited for only the duration of the elemental scan, or, in other words, at any moment, only one elemental area is excited.

In optical-mechanical television systems the corresponding fact of only one elemental area being illuminated at any moment, combined with the optical law that any such illuminated area cannot have a brilliancy higher than that of the light source, actually limited these systems to low definitions. Only through the use of the supersonic light modulator which permitted a simultaneous illumination of several hundred picture elements, *i.e.*, by applying the principal of *optical storage* could optical-mechanical television overcome this serious limitation, and, as developed in the Scophony system[2], achieve surprising successes for brighter and larger pictures than had been possible before with any other system.

The general importance of optical storage for television reception, in analogy to electrical storage for television transmission has been universally recognised, and its application in electronic reception system described.[1, 2]

In systems of this kind the information is carried to a storage screen by scanning this screen by a single modulated cathode-ray beam, and by some time-lag properties of this screen the elemental light values are retained there for a certain time, or a certain number of picture elements are simultaneously optically active.

Fluorescent screens show a certain amount of storage caused by phosphorescent afterglow and by using screens with a suitable afterglow characteristic improved television standards have been aimed at.[3]

The object of this paper is to investigate whether electronic picture screens (and particularly fluorescent screens) if excited as usual by a single cathode-ray beam of elemental spot size show limitations similar to those found in older optical mechanical systems, and to discuss ways to overcome these limitations by the use of *electronic storage excitation*.

In such single beam cathode-ray receiving tubes the area of the scanning spot will determine the definition of the picture. If this definition is to remain unaltered the only way to increase the intensity of the picture, for a given accelerating voltage, is to increase the current density of the beam. There exist, however, severe limits to the possible amount of this increase. At high beam densities the active centres of a fluorescent screen are rapidly destroyed and thus the life of the screen is seriously shortened.

But also with undamaged screens the fluorescent efficiency is lowered with higher current densities as a consequence of the saturation of the phosphorescent materials at high current densities so that no increase of the screen brightness can be obtained except by increasing the spot size, *i.e.*, allowing a certain amount of overlapping of adjacent picture elements, and thus impairing the definition. These difficulties become most acute in the case of cathode-ray tubes for optical projection. Owing to the relatively small size of the very bright fluorescent picture the current density of the cathode-ray spot must be very great, and usually a certain overlap, *i.e.*, a compromise with definition, must be tolerated in order not to impair too seriously the optical efficiency by surpassing the saturation limits of the screen.

These considerations have been illustrated by experiments in which a screen is scanned by an unmodulated cathode-ray beam. The spot size is varied by defocusing, keeping the total beam current constant, and thus varying the current density; the light output and thus the efficiency fall considerably when the spot is sharply focused, *i.e.*, when the current density is high (2 mA per square millimetre) and the screen material is saturated.[4]

Reducing the current density by distributing the total beam current over a larger area by defocusing the spot, and thus reducing the definition is a most unsatisfactory compromise. The only reasonable way to effect such a distribution of the beam current over a larger screen area without impairing definition would be the use of many sharply focused and independently modulated cathode-ray beams, following each other in the scanning direction, their modulations having the proper phase differences.

Such a multiple cathode-ray beam comprising n elemental beams might for instance simultaneously scan many n elements of a picture line, and would effect for any picture element a "storage excitation" of a duration n times the elemental scanning period.

Before discussing a simple promising way how to realise this object a simplified calculation may show what can be expected.

If excited by a cathode-ray beam of constant intensity the luminosity L of a phosphorescent material increases close to an exponential function and approaches with increasing time of irradiation a saturation value L_0. When the excitation ceases, the luminosity decreases, also in many cases approximately in an exponential function. Experience shows that materials with a

* "Picture Screen" in the following denotes various kinds of screens used in cathode-ray tubes, especially fluorescent screens, but also screens the transparency of which can be varied in accordance with the intensity of the electron beam, as for instance used in the Skiatron.[1]

short afterglow also show a fast increase of their luminosity towards saturation whereas materials with a long afterglow show a slow increase of their luminosity during the excitation. For a simple calculation we may consider an exponential increase and decrease of luminosity and equal time constants for the increase and the decrease. The luminosity increases then according to $L = L_0 (1 - e^{-\alpha t})$, and decreases according to $L = L_m e^{-\alpha t}$, where L_0 is the saturation luminosity and L_m the luminosity when the excitation ceases.

(a) An elemental scanning cathode-ray beam may excite a small screen area during the elemental period t_e up to a luminosity L_m equal to a certain fraction r of its saturation luminosity L_0 (Fig. 1):

$$r = L_m / L_0 = 1 - e^{-\alpha t_e}$$

After the time t_e the luminosity L_m decreases. The total effective light output of this area is determined by the integral over the luminosity curves from the time zero to infinite:

$$I_a = \int_0^{t_e} L_0 \cdot (1 - e^{-\alpha t}) \cdot dt + \int_{t_e}^{\infty} L_m \cdot e^{-\alpha t} \cdot dt$$

the calculation results in:

$$I_a = L_0 \cdot t_e \left(1 - \frac{r^2}{\log e \frac{1}{1-r}} \right)$$

(b) Now we regard the storage excitation of a screen area by a multiple cathode-ray beam comprising n elements and thus lasting for a time $n.t_e$ (Fig. 2). Here the effective light output is:

$$I_b = \int_0^{n.te} L_0 (1 - e^{-\alpha t})\, dt + \int_{n.te}^{\infty} L_n \cdot e^{-\alpha t}\, dt$$

where L_n is the luminosity reached after excitation time $n.t_e$ the calculation resulting in:

$$I_b = L_0 \cdot n \cdot t_e \left(1 - \frac{[(1-r)^n - 1]^2}{n \cdot \log e \frac{1}{1-r}} \right);$$

Of particular interest is the gain in light output obtained by this *storage excitation* as compared to elemental excitation:

$$g = \frac{Ib}{Ia} = n \cdot \frac{\log e \frac{1}{1-r} - \frac{1}{n}[(1-r)^n - 1]^2}{\log e \frac{1}{1-r} - r^2}$$

Since $r < 1$ a sufficient approximation is:

$$g = n \cdot \frac{\log e \frac{1}{1-r} - \frac{1}{n}}{\log e \frac{1}{1-r} - r^2}$$

An evaluation of these equations shows that for excitations leading after the elemental period t_e either to very low or to very high fractions r of the saturation g approaches n, or the light output is multiplied with the same factor as the excitation time of the screen area For all intermediate values of r, the gain factor g is larger than n and a maximum of $g = 1.55.n$ exists for an r of approximately 70 per cent., *i.e.*, for such screen material and current density that a small screen area has after the element period t_e reached about 70 per cent. of its saturation luminosity.

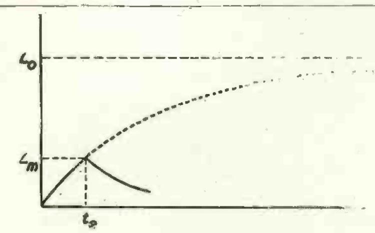


Fig. 1.

The following table shows some values of r and g:

r	0	1	10	50	90	99	99.9	100%
g	1	1.01	1.1	1.55	1.54	1.27	1.17	1

These considerations may be interpreted in two ways:

(1) If the excitation time of a screen area and the total beam current is increased by a factor n (over the element period t_e) the light output of that area is increased by a factor at least equal, generally larger than n. But the current density has remained unchanged.

(2) If the excitation time of a screen area is increased by a factor n, but the total beam current kept unchanged, or, in other words, the elemental cathode-ray beam is replaced by a multiple cathode-ray beam of equal total beam current, the light output remains at least the same, and is generally even increased. But in this case the current density has decreased in proportion to n.

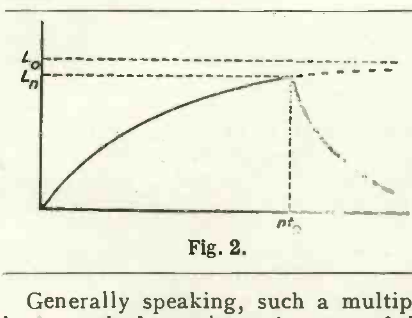


Fig. 2.

Generally speaking, such a multiple beam method permits an increase of the beam current by a very high factor without an increase in the current density beyond the saturation limit, and leads to an increase in light output by an even higher factor without the compromise of overlap impairing definition.

The advantages of such a multibeam cathode-ray system appear obvious after this discussion, but the realisation of a cathode-ray tube with a great many scanning cathode-ray beams modulated independently in proper phase relations to each other seems hopeless with conventional design. However, available electronic technique provides means with which to obtain automatically such a multi-beam device.[5]

Use is made of the multi-beam imaging properties of an electronic image converter tube 1 (Fig. 3) containing a photo-electrically active screen 2 and a fluorescent—or other image screen 6 and in which the electron beams liberated by an optical image projected on to the photo-electric screen can be imaged point by point on to the fluorescent screen by electron-optical focusing devices of an electrostatic and/or electromagnetic type, 3a, 3b, 3c.

This device is associated with a supersonic light modulator 14, as used in the Scophony optical-mechanical television system (2). This light modulator, which has been described in detail previously, contains a piezo-electric crystal 15 which is energised by high frequency voltages modulated in amplitude in accordance with the received picture signals from the oscillator-modulator 13.

The oscillating crystal sets up in the liquid 11 with which the cell is filled a travelling train of supersonic waves carrying these modulations.

A lens 12 collects the light from a light source 8 and diaphragm 9 which is diffracted by the waves and thus passes the stop 10, and forms an image of the moving train of waves on the photo-electric surface 2. This image will be that of a portion of a picture line, representing several hundred picture elements, their number depending upon the length of the wave train in the liquid. But the picture points reproduced on the photo-electric screen 2 and their electron-optical image on the fluorescent screen 6 are moving at speeds depending upon the velocity of propagation of the waves in the liquid.

Deflecting coils 5 (or electrostatic deflection means) move this picture of part of a line over screen 6 in the line scanning direction in such a sense and at such speed that the moving wave images represented in this line picture are immobilised with respect to the screen 6. Deflecting coils 4 move these line images over the screen in the frame-scanning direction so that a complete picture is reconstructed on the picture screen 6. The deflecting means are energised by time bases 4a and 5a synchronised by the television synchronising signals.

Thus the total number of elemental areas of the picture screen which are simultaneously excited by the electron beams is equal to the number of picture points represented in the wave train of the supersonic light modulator, and the optical storage inherent in this light modulator effects a "storage excitation" of each screen area for a time in excess of the usual elemental scanning period by a factor equal to the number of elements optically active in the supersonic cell which is usually several hundred.

The supersonic light modulator may be imaged on to the photo-electric screen by a cylindrical lens system in such a way that the magnification perpendicular to the line scanning direction is larger than in this direction, so that the line width on the screen 2 is considerable. The photo-electric screen is then imaged on to the picture screen 6 by an electron-optical imaging system, which in the line direction has a higher magnification than perpendicular to it (an anamorphotic* system 6) producing on the picture screen a line image of the proper ratio of length to width. This procedure makes it possible to distribute the total light flux from the light modulator over a larger area of the photo-electric surface, thus reducing the light flux per surface unit, and permitting a higher total beam current.

In the arrangement as shown in Fig. 3 both the photo-electric screen 2 and the fluorescent screen 6 have to be semi-transparent. An arrangement illustrated in Fig. 4 in which the electron beams are bent about 90° by suitable magnetic or electric deflection fields 21, permits the use of opaque (metal) screens which may be cooled, and which can be of considerably higher photo-electric efficiencies than semi-transparent screens.

In a normal image converter arrangement where the image on the photo-electric screen is stationary this screen must have a very uniform sensitivity over its surface, a property which is difficult to combine with highest photo-electric efficiency. In the present arrangement, however, where the picture points of the optical wave image, are in motion with respect to the photo-electric surface, any non-uniformity in the sensitivity of this surface will equally affect all picture points and will therefore not appear in the final picture.

The picture produced on the picture screen 6 may be utilised for direct viewing, or it may be projected optically on an enlarged scale on to a viewing screen.

The electron image converter effecting an electronic immobilisation of the moving wave images offers further the possibility of an image amplification, *i.e.*, the brilliancy of the fluorescent image produced on the picture screen may be many times greater than that of the image of the wave trains projected on to the photo-electric screen 2. This increase of brilliancy, or in the light flux depends upon the efficiencies of the photo-electric screen p and of the fluorescent screen f, and upon the accelerating voltage V applied in the tube.

* Anamorphosis : Distorted drawing appearing regular from one point.—O.E.D. (*Ed*).

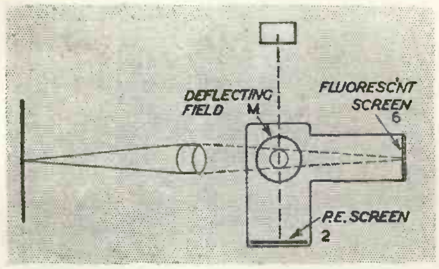


Fig. 4. Modification to the system permitting the use of opaque screens

The total light flux radiated by the fluorescent screen is :

$$L' = 4\pi . p . f . V . L,$$

where L is the light flux falling on to the photo-electric screen.

Thus the gain factor for the light flux is $4\pi . p . f . V$ with a p of 30 microamperes per lumen and an f of 4 candles per watt, and an accelerating voltage V of say 30.000 volts the gain factor would be approximately 45. However, since the light flux from the fluorescent screen is radiated in all directions, only a fraction of it is utilised. For instance, when projecting the fluorescent image on to a viewing screen by an optical system of an aperture f:1, about 5 per cent. of the light is collected by the optical system.

In comparing the system with a conventional cathode-ray tube for projection we can obtain an estimate of the advantages resulting from the above considerations relating to the screen saturation.

Suppose a cathode-ray tube with a 3 in. by 4in. fluorescent picture obtained with a beam of 3 milliamperes. With a 525 line-standard, in order to avoid overlapping, the spot area has to be of the order of 1/50th square millimetre, thus resulting in a current density of 150 mA per square millimetre, far exceeding the saturation limit of average fluorescent screens (about 2 mA/sq. mm). A considerable overlap with loss of definition will be required in order to prevent serious damage to and quick destruction of the screen. (*See* also Reference 4).

In the present quasi-multi-beam system the same current will scan with full definition simultaneously, say 250 picture-elements — depending upon the length of the supersonic cell utilised—and thus be distributed over an oblong line area of 0.14 × 36 mm., or more than 5 square millimetres, resulting in a current density of only 0.6 mA per square millimetre which is far below the saturation limit.

By using a larger fluorescent screen, even with a single-beam tube the current density could be considerably reduced, however, the cost of the optical projection equipment rises with a high power of the screen size, and screen pictures much larger than 3 in. by 4 in. would lead to prohibitive optical costs.

With the above-mentioned photo-electric efficiency a beam current of 3 mA will require on the photo-electric screen a light flux of 100 lumens.

Using as mentioned above, optical and electron-optical imaging systems of different magnifications in and perpendicular to the line direction, in such a way that the image of the aperture of the supersonic light control is spread over a considerable area of the photo-electric screen 2, of say 4 in. by 4 in., the illumination on this screen is approximately 1,000 foot-candles or of the same order of magnitude as the illumination caused on an outdoor object by diffuse daylight. A highly sensitive photo-electric layer, even if of the semi-transparent type on a glass carrier, can stand considerably higher illuminations without harmful effects.

It remains to estimate the light flux which will pass through, and is controlled by a supersonic cell light modulator of convenient design.

In the plane of propagation of the supersonic waves the linear aperture, a of the cell is determined by the number n of elements optically stored by the cell, and the velocity of the supersonic waves v. If $n = 250$ and $v = 10^5$, a is 7.5 cm. The angular aperture α in this plane is determined by twice the ratio of the wavelengths of the light to supersonic waves. With an average light of 5,000 ÅU and a carrier frequency of 20 Mc/s, resulting in a supersonic wavelength of 5.10^{-3} cm, we obtain $\alpha = 1/50$.

In the plane perpendicular to the propagation of the supersonic waves the apertures are only limited by optical considerations. Convenient values are, for the linear aperture, $b = 1.5$ cm, for the angular aperture $\beta = \frac{1}{4}$.

The light flux L which can be passed through such an oblong aperture by suitable optical imaging processes, and originating from a light source of intrinsic brilliance B, is in good approximation.

$$L = a.b.\alpha\beta.B,$$

with the above values :

$$L = 7.5 \times 10^{-2}.B$$

Using as light source an incandescent projection lamp of a usual brilliance of 1,600 stilb (cdls per sq.cm.) we obtain a maximum light flux passing through the supersonic cell of $L = 120$ lumens.

These rough calculations show that with available means, using a projection incandescent lamp, one can easily obtain sufficient light, and a suitable illumination on the photo-electric surface of an ordinary image converter tube, to result on the fluorescent screen of this tube in a small picture of high intensity and full definition, and remain safely below the saturation limit of the fluorescent screen.

The conditions underlying these estimates can be considerably improved by using in the image tube intermediate secondary-electron emissive screens, on to which the moving, or the im-

(*Continued on p.* 600)

Plastics in the Radio Industry

III.—Manufacture

E. G. COUZENS, A.R.C.S., B.Sc., and W. G. WEARMOUTH, Ph.D
(Messrs. B. X. Plastics Ltd.) (Messrs. Halex Ltd.)

THE manufacture of a plastic can be divided into three phases, the preparation of the basic plastic substances, its manipulation into forms suitable for the manufacture of articles and thirdly, the actual manufacture of the article itself.

The first phase consists of chemical reactions which fall into the following classes: (1) The preparation of the cellulose derivatives, *e.g.*, such compounds as cellulose nitrate, acetate or ether: (2) the formation of "thermo-setting" compounds by the reaction of formaldehyde with phenolic substances or urea; (3) the preparation of thermoplastic synthetic resins by a process of polymerisation of relatively simple substances containing a "double bond" (*see* part 1, p. 482) into "long-chain" substances of great molecular weight. Examples of these are polystyrene and polymethyl methacrylate, or "Perspex."

The second phase corresponds to the differing chemical structures of the three classes of the first phase. The cellulosic derivatives by themselves are only thermoplastic to a limited degree and have to be "plasticised" or softened by the addition of a high boiling solvent. They are then manufactured into sheets, rods and tubes, the raw materials of the final phase of article-manufacture, by a process known as the "celluloid technique," which involves the use of a low-boiling solvent to assist in mixing.

The second phase of the manufacture of the formaldehyde compounds is entirely different. Here the object is the production of a moulding powder and this is done by processing the raw material with fillers, colours and accelerators on hot rolls, followed by grinding. The casting process is an exception.

The thermoplastic synthetic resins are manipulated by processes derived in part from the celluloid technique or by casting or by the moulding or extrusion of powders to form sheets, rods and tubes, the process depending upon the nature of the resin, but unlike the cellulose thermoplastics, low-boiling solvent is not used.

The third phase, the manufacture of articles, is less diverse, the thermoplastic materials whether cellulosic or synthetic resinoid, being shaped by heat or machined from some stock form or moulded from powder, while the thermo-setting materials are used solely in a compression moulding process with the exception of cast resin already mentioned and the special case of the extrusion of tubes.

It is proposed to deal first with two outstanding examples, Xylonite (cellu-

Fig. 2 (right). Removing polished celluloid sheet from polishing "nip."
Courtesy of BX Plastics, Ltd.

Fig. 1 (left). Slicing celluloid block.
Courtesy of BX Plastics, Ltd.

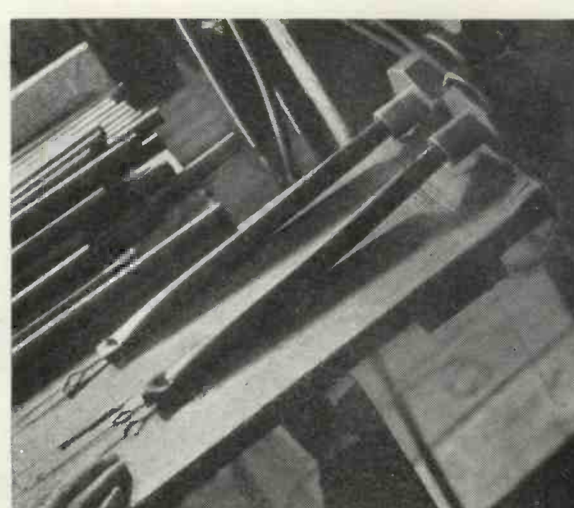


Fig. 3 (right). Extruding celluloid tubes.
Courtesy of BX Plastics, Ltd.

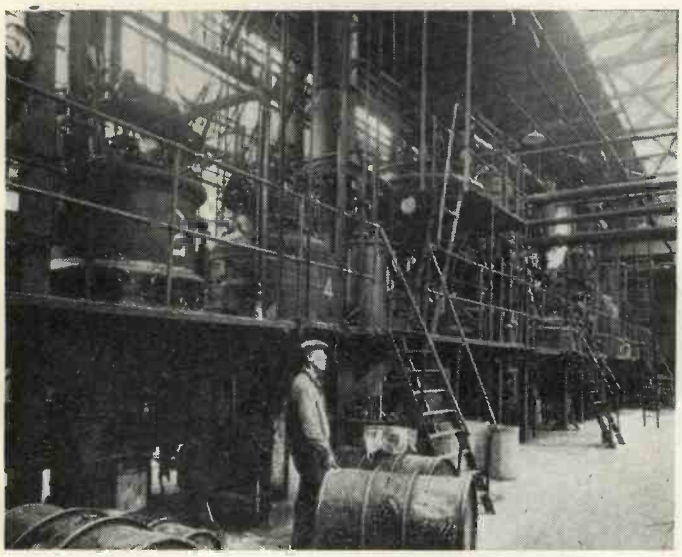


Fig. 4. Battery of resinoid digesters.
Courtesy of "Plastics."

loid) and phenol-formadehyde (Bakelite).

Xylonite

The starting material is the short fibre of the cotton-seed, known as "linters." After purification and drying it is treated with a controlled mixture of nitric and sulphuric acids and water in stainless-steel vessels provided with stirrers which are mounted above a stainless-steel centrifuge. During the process of nitration, cellulose nitrate is formed. During the process the cotton appears to be unchanged and after being discharged along with the acid into the centrifuge it is spun off and dropped through an opening in the bottom of the centrifuge into a large quantity of water, forming a slurry which is then pumped into vats, where it is washed until the acid content of the water is very small. It is then boiled with steam to destroy a sulphuric acid compound which renders it unstable, after which it is bleached with hypochlorite and spun off.

The resulting damp cake of white nitrocellulose fibre is now washed with alcohol in a hydraulic press or in a centrifuge. The cake still unchanged in appearance, is now placed in a special type of mixer rather like a baker's dough mixer and to it is added from 25 to 40 per cent. of camphor, together with the necessary dyestuffs and a filler if an opaque sheet is required. The alcohol in the nitrocellulose and the added camphor form a solvent for the nitrocellulose and in an hour or so, with gentle heating, the whole mass becomes a soft dough, containing about 40 per cent. of alcohol and some water. Since the cotton linters always contain some impurities it is now necessary to filter this dough and this is done in vertical hydraulic presses with jacketed cylinders through which hot water is passed. The great pressure used, above two tons per sq. inch, forces the dough through a cambric filter cloth mounted on gauzes supported on a bronze perforated grid on the underside of which the plastic emerges in long thin soft rods which are collected in a stainless steel pot. It is now necessary to reduce the solvent content of the dough to from 10 to 15 per cent. for the subsequent processes and this is done by milling on heated rolls, from which it is eventually transferred to calendering rolls which are slightly warmed and which convert it into firm sheets of plastic which are trimmed to a standard size and assembled in rectangular piles, which are then placed on cast iron bases with a grooved upper surface and pressed into solid blocks in heated hydraulic presses known as "block presses." The block of plastic is then mounted on the reciprocating travelling bed of a slicing machine which, like a plane, slices off sheets of any given thickness (Fig. 1) from 5/1,000 in. upwards, the solvent content of the dough being such that while firm and strong a knife will cut it readily. As the sheets still contain solvent, it is necessary to remove this by stoving for periods which vary from

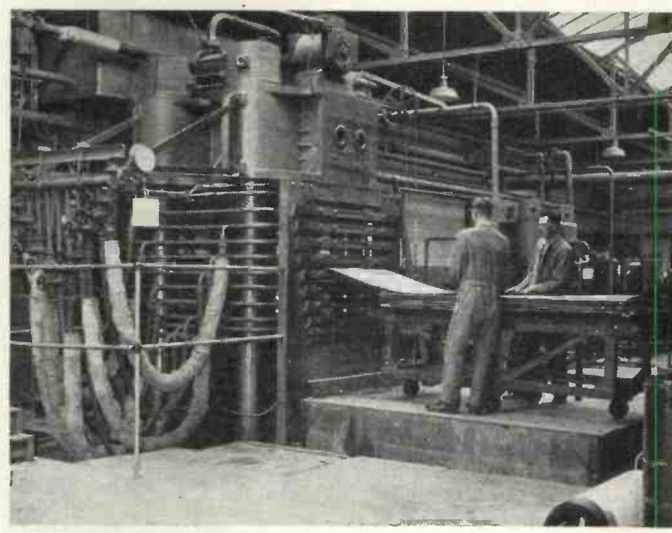


Fig. 6. Laminated sheet press.
Courtesy of Bakelite, Ltd

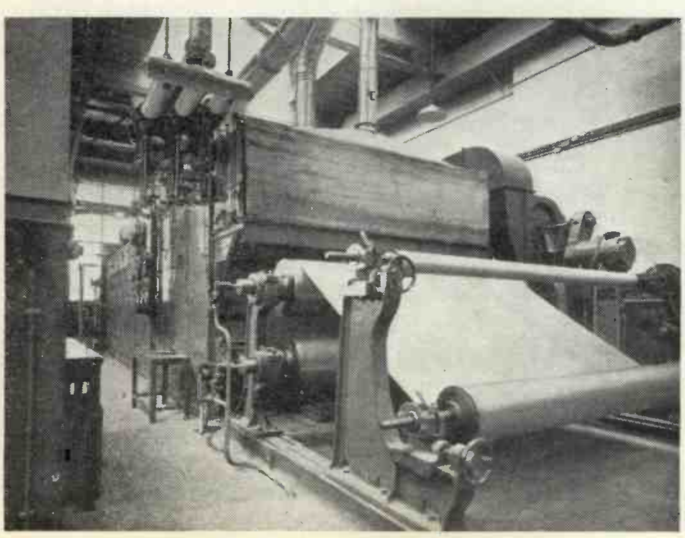


Fig. 5. Impregnating machine used in the production of Bakelite laminated sheet.
Courtesy of Bakelite, Ltd.

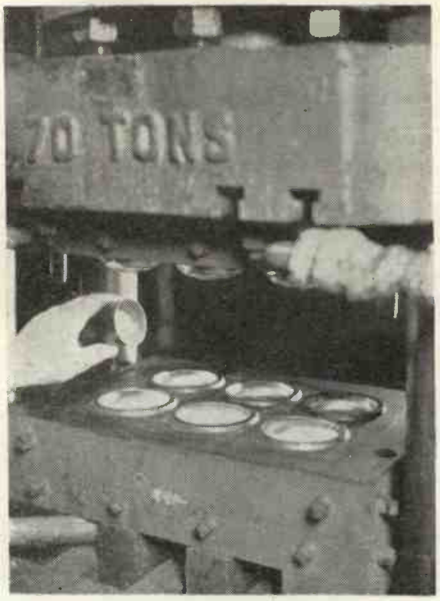


Fig. 7. Moulding of Bakelite (Loading mould).
Courtesy of Bakelite, Ltd.

a few days for thin sheets to many weeks for thick sheets.

The sheets, which are usually cut from blocks of such a size as to be about 4 ft. 6 in. by 2 ft. after the shrinkage caused by stoving, are now matt in appearance with fine lines, caused by the slicing-knife, running the length of the sheet and to remove these they are placed in contact with highly-polished metal plates in steam-heated hydraulic presses, the plastic being surfaced to a mirror finish by the highly-polished metal plate. A matt surface can be produced in the same way by contact with sand-blasted plates.

This complicated process is now highly standardised and can be very economically operated. One of its outstanding advantages is that not only by working in pieces of harder material of a different colour in the rolls can mottles be produced, but by packing freshly sliced sheets in various ways, repressing into a block and re-slicing, controlled patterns of great variety and beauty can be obtained. Ivory used for knife-handles is made in this way. The sheet from the calenders can also be rolled into a cylindrical block for insertion into a hydraulic extrusion press or cut into cubes for a "screw" extrusion machine, from both of which tubes or rods can be extruded (Fig. 3).

The thermoplastic nature of Xylonite permits of its being blown, moulded or shaped when hot into any form from a table tennis ball to a battery box, and separate parts can be readily cemented together by suitable solvents. With suitable precautions against firing by friction, it can also be machined, stamped or drilled in exactly the same way as wood or metal.

Phenol-formaldehyde Resins

The starting material for the phenolic resins is phenol or cresol and formaldehyde, the latter being in the form of a 40 per cent. aqueous solution, and the resin is formed by the reaction of these materials which unlike cellulose have no long chain characteristics to begin with.

The reaction is carried out by heating in digesters, *i.e.,* jacketed pots made of stainless-steel, nickel or copper, fitted with agitators and provided with reflux condensers, which prevent the escape of any vapours during the reaction. A battery of such digesters is shown in Fig. 4.

In order to promote the reaction between the phenolic substance and the formaldehyde a cataylst is used, which may be either an alkali or an acid, and the reactants are heated together in the digester at boiling temperature, for a short time if an alkali is used, for a substantially longer time and sometimes under pressure if an acid catalyst is employed. On cooling the water separates from the resinous layer and

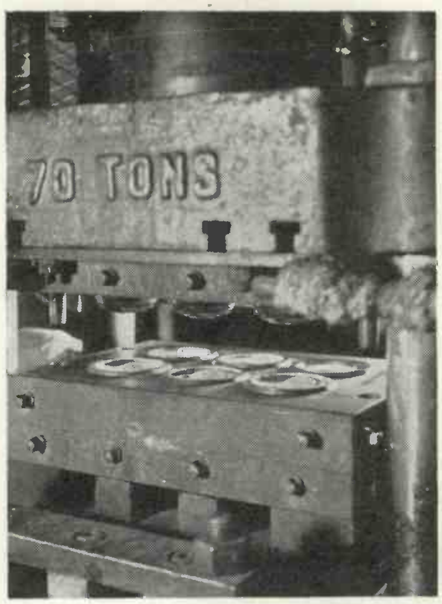


Fig. 8. Moulding of Bakelite (Press closing).
Courtesy of Bakelite, Ltd.

may be run off and any remaining water dried off by distillation in vacuo. The resin is either dissolved directly in spirit for impregnation purposes or run off, cooled and ground.

Two different types of resin called resoles and novolaks are produced by alkaline and acid catalysts respectively. Resoles cure, *i.e.,* pass to the infusible insoluble or "resite" stage very quickly by heating alone, but novolaks or acid-catalysed resins are permanently soluble and fusible resins unless treated with further quantities of formaldehyde. Resoles are used in the production of laminated sheets, a very important process not mentioned in the general introduction. Dissolved in alcohol or emulsified in water they can be used to impregnate thin wood or paper (Fig. 5) and on pressing between heated platens they are consolidated into boards of any required thickness, the resole changing by heat into hard infusible bakelite (Fig. 6). The novolak type of resin is used for moulding powders because to form them the incorporation of fillers and colours is necessary and if resoles were used, they would harden too quickly in the process of hot rolling required for incorporation. The easily fusible novolaks are therefore mixed with filler, which may be as much as 50 per cent. of the weight of the finished powder and may consist of wood flour for cheapness, cotton flock for toughness and asbestos for heat resistance. Colours are also added and the whole finally incorporated on hot rolls with hexamethylene tetramine, a compound of formaldehyde and ammonia, which provides the extra formaldehyde necessary to convert the novolak into the hard infusible form when it is finally moulded. The sheet material from the roll when cooled and ground to a suitable mesh is the raw material for the moulder, *i.e.,* for the third phase, the production of the finished article. Three self-explanatory illustrations are given (Figs. 7, 8 and 9) which are typical on a small and simple scale of all thermosetting mouldings. The process is rapid because the press can be opened hot, and metal inserts of all kinds can be incorporated in mouldings.

The world production of moulding powder of the phenolformaldehyde type is the largest of any one kind of plastic and amounted in 1938 to 60,000 tons.

The powder can, although thermosetting, also be extruded in a special machine, with a heating cycle adjusted so as to give just a long enough heating period to effect the final change to the fully heat-hardened resin. (Fig. 10).

The production of cast resin will be dealt with along with other special processes in a later article.

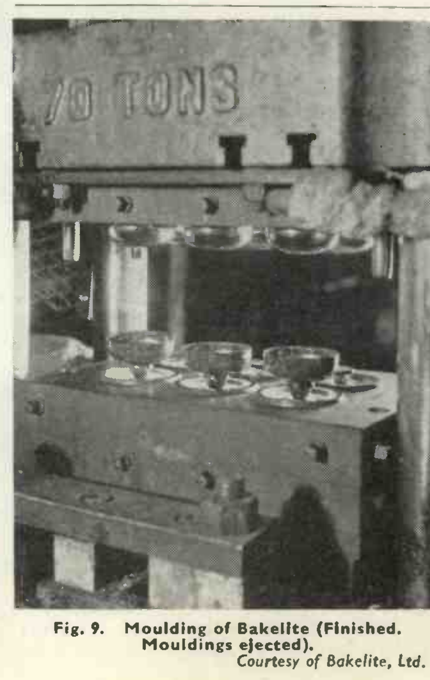


Fig. 9. Moulding of Bakelite (Finished. Mouldings ejected).
Courtesy of Bakelite, Ltd.

Frequency Modulation

Part III. — Methods of Modulating the Frequency or Phase of a Carrier

By K. R. Sturley, Ph.D., A.M.I.E.E.

PREVIOUS articles have already discussed the particular features and advantages of frequency and phase modulation and we will now deal with methods of producing these types of modulation. The close relationship between frequency and phase modulation has already been shown and only a very slight circuit modification is necessary to convert frequency into phase modulation and vice versa.[1] The modification consists of inserting a frequency discriminating network in the modulator stage. A resistance R and inductance L connected between the audio frequency source and the frequency modulator (Fig. 1a) results in a phase modulated output because the RL circuit makes the A.F. voltage amplitude directly proportional to its frequency. This means that a constant amplitude variable frequency voltage at AB gives an amplitude proportional to its frequency (f_m) across CD and the frequency deviation of the carrier is directly proportional to f_m, the condition for phase modulation. Similarly the RC circuit shown in Fig. 1b gives an amplitude across CD inversely proportional to f_m for a constant amplitude at AB and this in conjunction with the phase modulator produces a phase change of carrier inversely proportional to f_m, which is the characteristic of frequency modulation.

Frequency modulation of an oscillator may be accomplished by varying the equivalent inductance or capacitance of its tuned circuit, the magnitude of the L or C change being controlled in such a way that the resultant frequency change of the carrier is directly proportional to the modulating voltage amplitude. The required relationship between the frequency deviation of carrier and the L or C change is dependent on the magnitude of the ratio of the former to the carrier frequency. Probable values of frequency deviation and carrier are ± 75 kc/sec. and 40 Mc. so that $\frac{f_{dev}}{f_{carrier}}$ = ± .001875; this low value makes simplifications in the analysis possible because it means that the ratio of change of inductance (ΔL) to the total inductance (L) of the tuned circuit is also small. For example, let

$$f_c = \frac{1}{2\pi\sqrt{LC}}$$

= initial carrier frequency (1)

If the frequency deviation of carrier is Δf, we have:

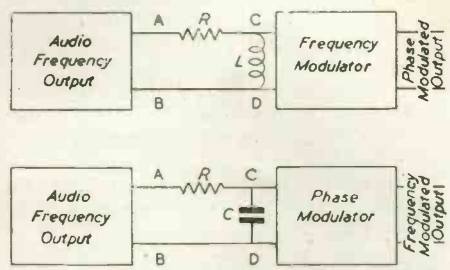


Fig. 1. (a) Phase modulated output from a frequency modulator. (b) Frequency modulated output from a phase modulator

$$f_c + \Delta f = \frac{1}{2\pi\sqrt{(L - \Delta L)C}} \qquad (2a)$$

and

$$f_c - \Delta f = \frac{1}{2\pi\sqrt{(L + \Delta L)C}} \qquad (2b)$$

Combining (1) and (2a)—the result is the same if we use (1) and (2b)

$$\frac{f_c + \Delta f}{f_c} = 1 + \frac{\Delta f}{f_c} = \frac{1}{\sqrt{1 - \frac{\Delta L}{L}}}$$

$$= \left[1 - \frac{\Delta L}{L}\right]^{-\frac{1}{2}} \quad \ldots \quad \ldots \quad (3)$$

Expanding by the Binomial theorem

$$1 + \frac{\Delta f}{f_c} = 1 + \frac{\Delta L}{2L} - \frac{3}{8}\left(\frac{\Delta L}{L}\right)^2 + \text{ etc.} \quad (4)$$

but since $\Delta f / f_c$ is very small it follows that $\Delta L/L$ is also small; hence $(\Delta L/L)^2$ is negligible and (4) becomes

$$\frac{\Delta f}{f_c} = \frac{\Delta L}{2L} \quad \ldots \quad (5a)$$

$$\text{or } \Delta f \propto \Delta L \quad \ldots \quad (5b)$$

Similarly for a small change of capacitance ΔC

$$\frac{\Delta f}{f_c} = \frac{\Delta C}{2C} \quad \ldots \quad (5c)$$

$$\text{or } \Delta f \propto \Delta C \quad \ldots \quad (5d)$$

Undistorted frequency modulation can therefore be realised by making the change in L or C directly proportional to the amplitude of the modulating frequency. It is not always convenient to make a direct change of L and it may be necessary to obtain it by varying an inductance, or its equivalent, placed in parallel with L. The following equations then result:—

$$f_c = \frac{1}{2\pi\sqrt{\frac{LL_1}{L + L_1}C}} \quad \ldots \quad (6)$$

$$f_c + \Delta f = \frac{1}{2\pi\sqrt{\frac{LL_2}{L + L_2}C}} \quad \ldots \quad (7)$$

where L_1 is greater than L_2

The lower deviation of carrier $f_c - \Delta f$ yields a similar equation, but need not be considered.

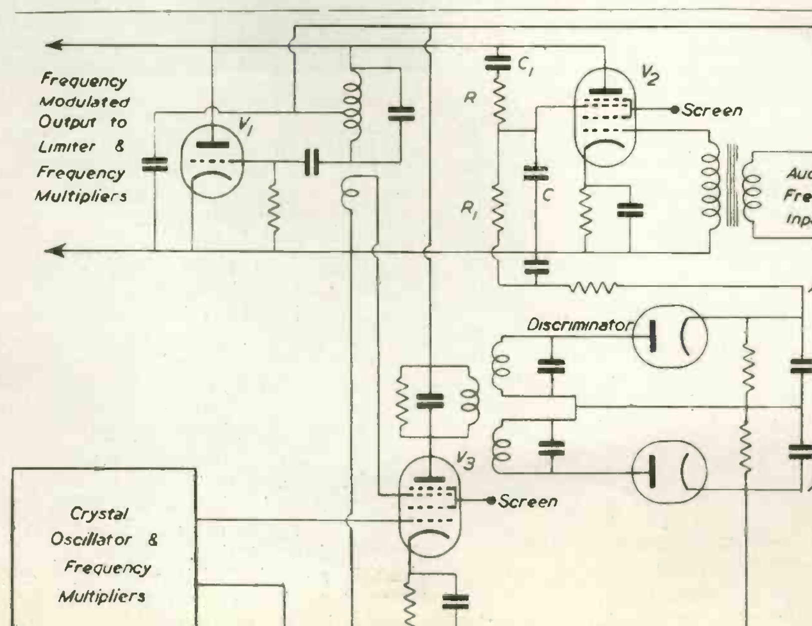


Fig. 3. Frequency modulation with stabilised variable reactance valve

$$1 + \frac{\Delta f}{f_c} = \sqrt{\frac{1+\frac{L}{L_2}}{1+\frac{L}{L_1}}} = \left[1+\frac{L}{L_2}\right]^{\frac{1}{2}}$$

$$\left[1+\frac{L}{L_1}\right]^{-\frac{1}{2}} = 1 + \frac{L}{2L_2} - \frac{L}{2L_1}$$

if L_1 and $L_2 \gg L$

$$\text{thus } \frac{\Delta f}{f_c} = \frac{L}{2}\left[\frac{1}{L_2} - \frac{1}{L_1}\right]$$

$$\text{or } \Delta f \propto \left[\frac{1}{L_2} - \frac{1}{L_1}\right] \quad \ldots \quad (8)$$

For Δf to be proportional to the modulating frequency amplitude, the latter must be proportional to $[1/L_2 - 1/L_1]$ and we shall see later that this result can be achieved.

An electrical method for varying the inductance or capacitance of the carrier oscillator tuned circuit is obviously preferable to a mechanical one. The variable reactance valve (as used in automatic frequency correcting circuits) is particularly suitable for this purpose since it acts as a reactance to any voltage source connected between its anode and cathode, the value of the reactance depending on the grid bias of the valve. To realise this property it is necessary to supply the grid with a proportion of the anode voltage which has been given a phase shift of 90°. The basic circuit is that of Fig. 2; impedances Z_1 and Z_2 act as a potentiometer to step down and phase shift the grid voltage. Thus if Z_1 is a resistance R, Z_2 a capacitance C, and the valve is a pentode or tetrode with a high internal resistance, the admittance across its anode and cathode (points A B) is given by

$$Y_{AB} = \frac{I_a}{E_a} = \frac{g_m E_g}{E_a}$$

where g_m = mutual conductance of the valve

$$Y_{AB} = \frac{\frac{g_m E_a Z_2}{Z_1 Z_2}}{E_a} = \frac{g_m Z_2}{Z_1 + Z_2}$$

$$= \frac{g_m}{1 + jR\omega C}$$

$$= \frac{g_m}{1+(R\omega C)^2} - j\frac{g_m R\omega C}{1+(R\omega C)} \quad \ldots \quad (9)$$

which is equivalent to a resistance

$R_{AB} = \dfrac{1+(R\omega C)^2}{g_m}$ in parallel with an inductance $L_{AB} = \dfrac{1+(R\omega C)^2}{g_m R\omega^2 C}$. Three other combinations of R and L or C are possible and the resultant parallel resistance and reactance components of Y_{AB} are tabulated below

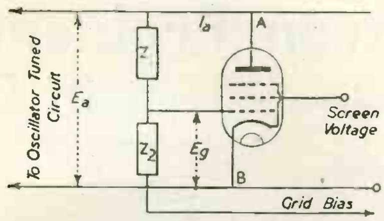


Fig. 2. Variable reactance valve circuit

From the first and fourth columns in the table we see that the equivalent inductance is inversely proportional to g_m so that expression (8) above becomes

$$\Delta f \propto g_{m_2} - g_{m_1}$$

and if the $g_m E_g$ curve of the variable reactance valve is a straight line

$$\Delta f \propto E_{g_2} - E_{g_1}$$

Hence by inserting a modulating voltage in series with the D.C. grid bias the resultant variation of inductance produces a deviation of carrier frequency directly proportional to the modulating voltage amplitude. Similarly if g_m is proportional to E_g, the change in capacitance in columns 2 and 3 of the table means that Δf is proportional to the modulating voltage amplitude (*see* expression 5d).

It is not usual to employ the reactance valve to modulate an ultra high frequency oscillator; a lower carrier frequency is used and frequency multiplier stages inserted between it and the aerial to step up the frequency to the required 40 Mc/s. This has the advantage of simplifying reactance valve circuit design and of reducing interaction between the oscillator and succeeding amplifier stages. Furthermore greater oscillator stability is possible at lower frequencies. The resistance component R_{AB} is undesirable because it causes amplitude modulation, but by a suitable choice of R and L or C and a reduction of the resonant impedance (L/CR) of the carrier tuned circuit this effect can be reduced to small proportions. A limiter may be inserted before the multiplier stages to reduce still further any amplitude modulation. For example if the oscillator frequency is 1,000 kc/sec. Z_1 a capacitance of 5 $\mu\mu$F and Z_2 a resistance of 5,000Ω, mean g_m = 1 mA/volt, and oscillator tuning capacitance 200 $\mu\mu$F.

$$C_{AB} = \frac{g_m CR}{1+(R\omega C)^2} = 24.4\ \mu\mu\text{F}.$$

total tuning capacitance = 224.4 $\mu\mu$F.

the required change of capacitance $\Delta C = \pm 224.4 \times .00375 = \pm 0.84\mu\mu$F. (note from the first part of the article and expression 5c that $\Delta f/f_c = 0.001875$ and $\Delta C/C = 2\Delta f/f_c = 0.00375$). This means a mutual conductance change of $\pm 0.84/24.4 = \pm 0.0344$ mA/volt. The value of R_{AB} for g_m = 1 mA/volt is 40,400Ω and it varies from 41,800Ω to 39,000Ω. If the oscillator coil has a Q of 50 giving a resonant impedance of 35,500Ω, the variation in amplitude due to the variation of R_{AB} corresponds to a modulation percentage of approximately 1.6 per cent. The variable reactance value can be placed across a part only of the oscillator coil if desired and this calls for a greater change of g_m to give a specified frequency modulation. Amplitude modulation by R_{AB} is slightly increased. The component R_{AB} can be made very high by the use of special circuits in the phase splitting network Z_1 Z_2 or by negative feedback in the oscillator circuit.

When a reactance valve is used as a modulator it is essential to prevent change of g_m by means other than the modulating voltage amplitude; variation of HT, hum voltages due to the L.T., ageing of the valve, all can contribute to change of g_m. The first two effects can be reduced to small proportions by using two reactance valves in push-pull, but a more satisfactory method[4, 5] of maintaining carrier stability is to employ an automatic frequency correction circuit to control the D.C. bias on the reactance valve. A basic circuit is shown in Fig. 3. Valve V_1 is the carrier oscillator (Hartley circuit) and V_2, a hexode valve, is the variable reactance device. The audio frequency voltage is applied at the first grid (the signal grid when V_2 is used as a frequency changer) to control the mutual conductance of the third grid to which the RC phase splitting network is connected. The third grid (normally the oscillator grid) is used because its I_aE_g curve is more linear for large applied voltages than that of the first grid, and to get the required reactance change the valve must be connected across the major portion of the oscillator coil. Condenser C_1 is the A.C. coupling between anode and grid and has a large value (0.1 μF) whilst R_1 has a resistance (about 0.5 MΩ) large compared with the reactance of condenser C. In the first article we showed that a frequency modulated wave consisted of a central carrier frequency (equal to the unmodulated value) and sidebands. This

(*Continued on page* 600)

Z_1	R	C	R	L
Z_2	C	R	L	R
R_{AB}	$\dfrac{1+(R\omega C)^2}{g_m}$	$\dfrac{1+(R\omega C)^2}{g_m\ (R\omega C)^2}$	$\dfrac{R^2+\omega^2L^2}{g_m\ \omega^2L^2}$	$\dfrac{R^2+\omega^2L^2}{g_m}$
X_{AB}	$L_{AB} = \dfrac{1+(R\omega C)^2}{g_m\ R\omega^2 C}$	$C_{AB} = \dfrac{g_m\ RC}{1+(R\omega C)^2}$	$C_{AB} = \dfrac{g_m\ RL}{R^2+\omega^2L^2}$	$L_{AB} = \dfrac{R^2+\omega^2L^2}{g_m\ R\omega^2 L}$

Secondary Electron Problems in Beam Tetrodes

By J. H. OWEN HARRIES, A.M.I.E.E.

THE critical distance tetrode consists, in essence, of a structure a section through which is sketched in Fig. 1. Beams of electrons travel from the cathode to the anode as indicated. The beams are formed by an electron lens system which exists due to the configuration and relative potentials of the cathode, control grid and screen grid. This valve is of a type which first reached the market in 1935.[1]* Some beam tetrodes are provided with "beam forming electrodes" of cathode potential such as those indicated by the dotted lines in Fig. 1. Critical distance valves were devised by the author in 1931,[2] and came into wide use under the name of the 6L6 beam tetrode when this valve was first marketed by the R.C.A. in 1936. Beam forming electrodes were used in these 6L6 valves.

The beam tetrode operates and has a "dynatron-free" characteristic because of the effect of a potential minimum due to space charge between the screen grid and the anode. In valves of the 6L6 and similar types, the beam forming plates exercise only a slight effect in producing the beam; but have an appreciable effect in producing the retarding potential;—indeed, tubes of this type first built up by the author in 1932 were looked upon by him as hybrids and as being somewhere between the pure beam tetrode and the pentode valve in which undesirable secondary electron current is got rid of by the action of a suppressor grid of zero potential positioned between the screen grid and the anode.[11]

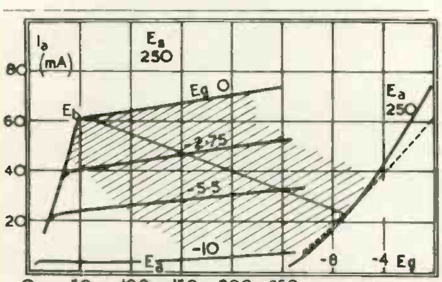


Fig. 2. Measured characteristics of critical distance tetrode.

An outline of the history of the beam tetrode has already been published.[3] Because, for instance, the Radio Corporation of America sells, at the moment, 17 receiving beam tetrode types and four transmitting types (including certain much publicised ultra-high-frequency vaves), it is not surprising that several mathematical theories have been published. Salzberg & Haeff, of the R.C.A. research staff, have published[4] one which may be taken as typical. Broadly speaking, this analysis consists of an ingenious mathematical treatment of a theoretical space charge due to primary electrons between the screen grid and the anode, and deduces (quite correctly) that there can be a potential minimum. This was done by an extension of well-known space charge theories, such as that published some years ago with respect to triodes by Gill[5] and Tonks.[6]

*Figures refer to bibliography at end of article.

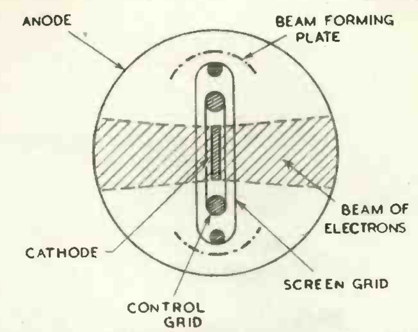


Fig. 1. Diagrammatic sketch of critical distance tetrode showing the electrode arrangement.

It will, however, be shown in the present paper that the operation of the beam tetrode has not in fact been explained at all. The values of potential minimum deduced by the published theories are quite incapable of preventing a dynatron characteristic being produced and do not fit the facts in other ways. In fact, a new approach to the problem, and one involving other factors, is required.

The Properties of Secondary Radiation

Certain facts about secondary electron emission which were published as long ago as 1922 have apparently escaped notice.

Fig. 2 shows typical anode characteristics of a critical-distance valve.

In such a beam tetrode a potential distribution may be expected somewhat similar to that sketched in Fig. 3. The potential minimum Vc corresponds to a retarding potential (1-Va/Vc), which tends to prevent the secondary radiation from travelling back from the anode to the screen grid. (It must not be forgotten, by the way, that secondary radiation of an appreciable amount may travel from the screen grid to the anode).

H. E. Farnsworth[7] published in 1922 the values of retarding potential required between the screen grid and anode to reduce the ratio of primary to secondary current to any desired amount. His method of measurement needed some slight corrections (afterwards made by him in 1928)[8] but is substantially correct.

The following table is plotted from Farnsworth's Fig. 5 of his 1928 paper,[8] and shows a typical result for iron. The results for nickel are much the same.

It is clear that a retarding potential of a very considerable amount will be necessary if, when the anode voltage is raised from the knee value E_B (Fig. 2) to a value equal to the screen voltage, the working surface (shaded) of the anode characteristic is to remain substantially flat.

TABLE

Relationship between retarding potential and secondary electron coefficient for iron.

Retarding potential (1—Vc/Va).	Secondary radiation coefficient (Is/Ip %).
0.99	6%
0.95	6.5%
0.8	8.2%
0.7	9.5%
0.6	12.0%
0.5	13.5%
0.4	16%
0.3	18.5%

Fig. 4 is drawn from Farnsworth's and other information and shows a plot of retarding potentials against secondary radiation coefficient Is/Ip, for various anode potentials.

Fig. 5 indicates a typical distribution of velocity of secondary electron emission from a metal target. It will be observed that there is a peak of emission at a velocity almost equal to the primary impact velocity. It has been shown by Farnsworth and by other workers[9] that this peak gets relatively less in magnitude as the anode voltage (impact energy on the anode) is increased from a small value up to the order of at least 250 volts.

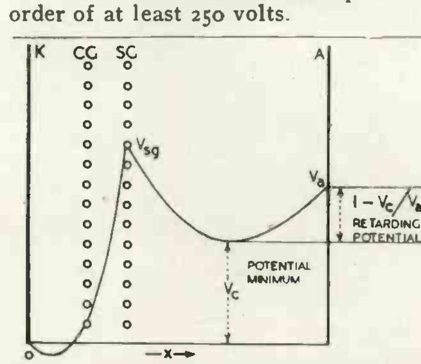


Fig. 3. The potential gradients in the critical distance tetrode. The arrangement of the electrodes is only diagrammatic.

The radiation coefficient Is/Ip for all secondary electrons (disregarding their relative velocities) increases, in the case of a clean nickel anode, from a value of about 0.4 at an anode potential of 50 volts to about unity at an anode potential of 250 volts.

Davisson and Kunsman[9b, 9c] and Tate[9d] show that secondary emission from a plate anode travels almost entirely in the direction from which the primary electrons arrive.

It is clear that unless the retarding potential in a beam tetrode is very considerable (so that the operation is well to the left hand of Fig. 4), as the anode potential increases from a knee value up to a value towards the screen voltage, the secondary electron current will increase rapidly and a dynatron characteristic will be found. For instance, if the knee of the anode characteristic is imagined to be at 50 anode volts, and if the retarding potential averages about 0.2, then the order of about 30 per cent. secondary radiation current will flow from the anode to the screen grid at anode potentials intermediate between the knee and screen voltage. The result will be a very pronounced dynatron type dip in the characteristic. The only way in which linear characteristics can be explained is by a comparatively complicated mechanism involving the use of retarding potentials of fairly high values at the left-hand end of Fig. 4. A theory of this kind is in course of preparation.

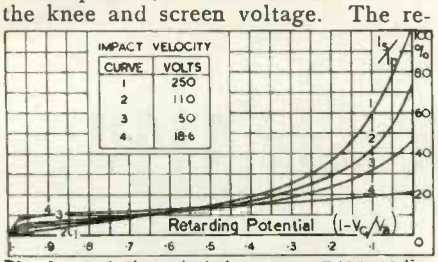


Fig. 4. Relationship between secondary radiation coefficient Is/Ip and retarding potential for various impact velocities on the anode. Impact velocities are expressed in terms of anode volts (after Farnsworth and others).

The problem is not merely one of preventing the passage of secondary radiation at one particular value of anode current, anode voltage and screen voltage. It is, on the contrary, that of maintaining a flat working surface to the characteristic over wide variations of these quantities.

The previous analyses, if quantitatively applied, produce a result such that, even if the potential minimum is sufficient at the knee voltage E_B (in Fig. 2), it is hopelessly insufficient when the operating locus travels down the load line in the direction of the arrow, the anode current drops and the anode voltage rises.

Referring quantitatively to the previous theories (of which that due to Salzberg & Haeff is taken as typical), these workers published a graph (their Fig. 7) which shows the values of potential minimum for various values of a dimensionless parameter P_a. P_a includes in itself such factors as the square root of the anode current and the reciprocal of the 3/2 power of the screen grid voltage. The anode current is therefore proportional to P_a^2. It is possible from this information to plot quantitatively the anode current characteristics predicted by the Salzberg & Haeff theory. This has been done in Fig. 6 of the present paper. In the author's opinion, it is difficult to see any resemblance between these predicted characteristics and those of an actual beam tetrode. The very large hysteresis loop effects are not found in the beam tetrode. For instance, at $P_a = 2$, two "knee values" exist at $Va/Vg = 1$, and at $Va/Vg = 0.35$. Apart altogether, however, from this discrepancy, it is clear that for any "knee" to exist at a value of Va/Vg, much less than unity (*i.e.*, for a beam tetrode type characteristic to exist at all), P_a cannot be much greater than say 1.7 to 1.8 or so.

Moreover, again utilising Salzberg & Haeff's Fig. 7, it is possible to draw Fig. 7 of the present paper. This graph shows retarding potential plotted against Va/Vg for various values of Pa which give knees that are below the screen voltage. Excepting for the left-hand end of the P_a = curve (which is instable and therefore useless) the retarding potential does not exceed about 0.15 V_a. A reference to Fig. 4 of the present paper shows that there is no possibility of the valve giving anything else than a pronounced dynatron characteristic. Apart, however, from this point, Salzberg & Haeff's theory fails in another way, because it is clear that, even if at some particular point (such as X on Fig. 7 of the present paper) a certain amount of secondary radiation could be stopped, the slightest change of anode current or anode voltage would upset these conditions completely.

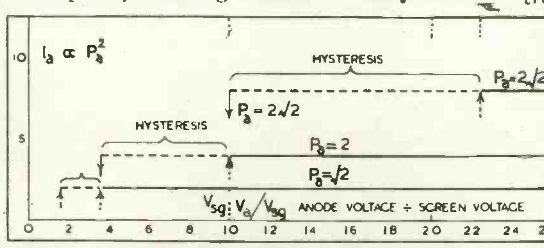


Fig. 6. Anode characteristics according to previous theory Plotted from Fig. 7 of Salzberg and Haeff.

In other words, the Salzberg & Haeff theory shows a hysteresis effect which is not found in actual beam tetrodes; an insufficient retarding potential to prevent the occurrence of a dynatron characteristic, and, finally, no range of anode current and voltage over which suitable values of retarding potential are maintained to produce the very large area of substantially flat working characteristic existing in actual valves (*see* the shaded area in Fig. 2). A large working area is really the principal justification for the existence of the beam tetrode as a commercial device. It will be remembered that this valve was originally put forward,[1, 10] because it possessed a flatter working area of anode characteristic than other multi-grid valves, and therefore had a lower level of distortion for a given power output.

It is an experimental result that no satisfactory beam tetrode critical distance effect (*i.e.*, dynatron-free characteristics) can be produced without the electrons being formed into a beam. (Fig. 1). The use of a focused beam for this purpose was first suggested by the author[2]. It is noticeable that the beam becomes narrower and the electrons more bunched when the control grid is made more negative. This is a very important factor in maintaining the potential minimum when the anode current is reduced in value during operation.

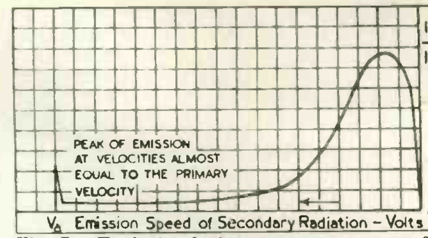


Fig. 5. Typical velocity distribution curve for secondary radiation. Impact velocity indicated by Vp.

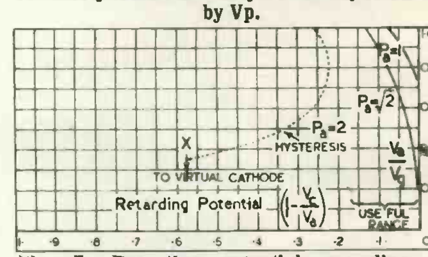


Fig. 7. Retarding potential according to previous theory. Plotted from Fig. 7 of Salzberg and Haeff.

In this paper the author has confined himself to pointing out certain information about secondary radiation which is apparently not too well known; and to a purely destructive criticism of the existing theories of the critical distance beam tetrode. What is hoped will prove to be an adequate theory is in course of preparation.

BIBLIOGRAPHY.

1 Harries, *Wireless World*, August 2nd, 1935.
2 Harries, British Patents 380429 and 385968/1931, and *Wireless Engineer*, April, 1936, Vol. 13, pp. 190-199.
3 Harries, *Wireless Engineer*, April, 1938, Vol. 15, p. 212.
4 Salzberg and Haeff, *R.C.A. Review*, Vol. II, No. 3, January, 1938, pp. 336-374; Dow, "Fundamentals of Engineering Electronics," Chapman and Hall.
5 Gill, *Phil. Mag.*, May, 1925, pp. 993 to 1,005.
6 Tonks, *Phys. Rev.*, October, 1927, pp. 501 to 511.
7 Farnsworth, *Phys. Rev.*, 1922, pp. 358 to 374.
8 Farnsworth, *Phys. Rev.*, March, 1928, Vol. 31, pp. 405-413.
9 (*a*) Haworth, *Phys. Rev.*, July 1st, 1935; (*b*) Davisson and Kunsman, *Phys. Rev.*, Vol. 22, No. 3, September, 1923, pp. 242 and 256; (*c*) Davisson and Kunsman, *Science*, Vol. 54, No. 1,404, November 25th, 1921, p.p. 522 to 524; (*d*) Tate, *Phys. Rev.*, Vol. 17, Series 2, No. 3, March, 1921, pp. 394 to 395. (*e*) Sharman (Communication by Professor Sir E. Rutherford), *Proc. Cam. Phil. Soc.*, Vol. 23, Part 5, January, 1927, pp. 523 to 530.
10 Harries, *Wireless Engineer*, February, 1937, Vol. 14, No. 161, p. 63.
11 Telleghen, British Patent 287958**.

DATA SHEET No. XVIII

The Self Capacity of Coils:– Its Effect and Calculation

Electronic Engineering

EVERY inductance coil has distributed capacities between adjacent turns, and turns that are not adjacent; also every turn has distributed capacities to surrounding earthy objects or the screening can. In addition the coil may be fitted with terminals or tags and there will also provide an additional capacity.

The cumulative effect of the above capacities can in the majority of cases be fairly accurately represented by a fictitious capacity across the whole coil. (Fig. 1a). This fictitious capacity is known as the self-capacity or total distributed capacity of the coil.

While in some cases when the coil is connected across a low-loss condenser the effect of the self capacity may be absorbed in the tuning condenser, in other applications such as filters it may be necessary to provide pure inductance elements under which conditions self-capacities will prove very deleterious. (*See* Fig. 1a and b). While inductance coils are not normally employed at frequencies near their natural resonance $f_o \approx 1/(2\pi\sqrt{LC_s})$, this resonance effect due to the self-capacity will produce absorption effects in multi-band receivers unless precautions are taken to prevent it.

Effect on Inductance and Resistance.

Fig. 2a illustrates the case of an inductance L whose dissipative losses are represented by the series resistance r_1 and the two are shunted by a loss-less self-capacity C_s

The effect of C_s is to alter the apparent reactance of the circuit between the terminals A and B. For frequencies below resonance this circuit may be represented by an effective inductance L_e and resistance r_e in series, given by :—

$$L_e = \frac{L(1-\omega^2LC_s) - C_s r_1^2}{(1-\omega^2LC_s)^2 + \omega^2C_s^2r_1^2} \approx \frac{L}{(1-\omega^2LC_s)} \quad (1)$$

$$r_e = \frac{r_1}{(1-\omega^2LC_s)^2 + \omega^2C_s r^2_1} \approx \frac{r}{(1-\omega^2LC_s)^2} \quad (2)$$

$$f_o = \frac{1}{2\pi\sqrt{LC_s}}\sqrt{1 - \frac{C_s r^2_1}{L}} \approx \frac{1}{2\pi\sqrt{LC_s}} \quad (3)$$

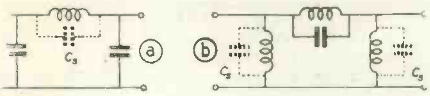


Fig. 1.

So that up to the natural resonance of the coil the effect of increasing frequency is to increase the effective inductance and resistance. At frequencies above the natural resonance the reactance becomes capacitative. When the dielectric of the component capacities of the total self-capacity (such as the coil former, wire insulation, dope, etc.) has not a negligible loss, a series resistance r_2 has to be added to the capacity branch (Fig. 2b) to account for this loss. This loss resistance modifies equations (1) (2) and (3) to

$$L_e = \frac{C_s\left(\frac{L}{C_s} - r_1^2\right) - \omega^2LC_s^2\left(\frac{L}{C_s} - r_2^2\right)}{(1-\omega^2LC_s)^2 + \omega^2C_s^2(r_1+r_2)^2} \quad (4)$$

$$r_e = \frac{r_1 + \omega^4L^2C_s^2r_2 + \omega^2C_s^2r_1r_2\,(r_1+r_2)}{(1-\omega^2LC)^2 + \omega^2C_s^2\,(r_1+r_2)^2} \quad (5)$$

$$\approx \frac{r_1 + \omega^4LC_s^2r_2}{(1-\omega^2LC_s)^2}$$

$$f_o = \frac{1}{2\pi\sqrt{LC_s}}\sqrt{\frac{L-C_sr_1^2}{L-C_sr_2^2}} \quad (6)$$

With dielectrics that will normally be encountered the power factor does not vary rapidly with frequency, so that over the tuning range of a coil the power factor remains almost constant. This implies that r_2 is approximately inversely proportional to the frequency. The second term of the numerator of the second equation (5), *i.e.*, $\omega^4LC_s^2r_2$ which represents the equivalent series loss resistance due to the dielectric will in practice therefore vary only as f^3 over the normal tuning range of a coil.

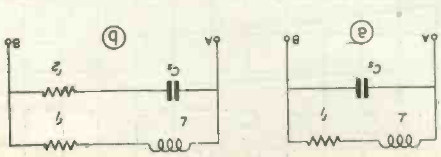


Fig. 2

Just as it has been shown that the resistance r_2 is a function of frequency, so in the case of the coil resistance r_1 it is found that this varies appreciably with frequency even over the normal tuning range of a coil. At the lower frequencies the A.C. component of the copper resistance r_1 increases as f^2 at high frequencies as $f^{\frac{1}{2}}$. Actually it is not the magnitude of the frequency, but the magnitude of the product of the wire diameter times the square root of the frequency that controls the rate of increase. In general therefore the A.C. component of the resistance r_1 varies at a rate proportional to some power of the frequency between $f^{\frac{1}{2}}$ and f^2 depending on the constants and the frequency of operation.

Calculation of Self Capacity.

There is no known accurate method of calculating the self capacity of a coil; however, Palermo (*Proc. I.R.E.*, July, 1934) has given a relation for short coils by means of which a useful estimate can be made of the value of C_s when the coil length is at the most not longer than the diameter of the coil. Palermo gives :

$$C_s = \frac{\pi D}{3.6\,\cosh^{-1}(p/d)}\ \mu\mu F \quad (9)$$

$$= \frac{\pi D}{3.6\,\log_e\left[\frac{p}{2d} + \sqrt{\left(\frac{p}{2d}\right)^2 - 1}\,\right]}\ \mu\mu F \quad (10)$$

where D is the diameter of the coil winding between wire centres in cms. p is the pitch and d is diameter of the bare wire. (p and d must be in the same units). The above equation neglects the effect of the wire covering and coil former on the self capacity and will therefore in general give too low a value.

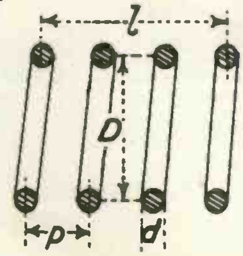
Equation (9) and (10) have been plotted on Data Sheet No. 19 for values of (p/d) up to 2.5 and coil diameters between 2 and 20 cms. As C_s is directly proportional to D the self capacity for larger coils may be easily read off the curves.

5·90

THE SELF-CAPACITY OF SHORT SOLENOIDS

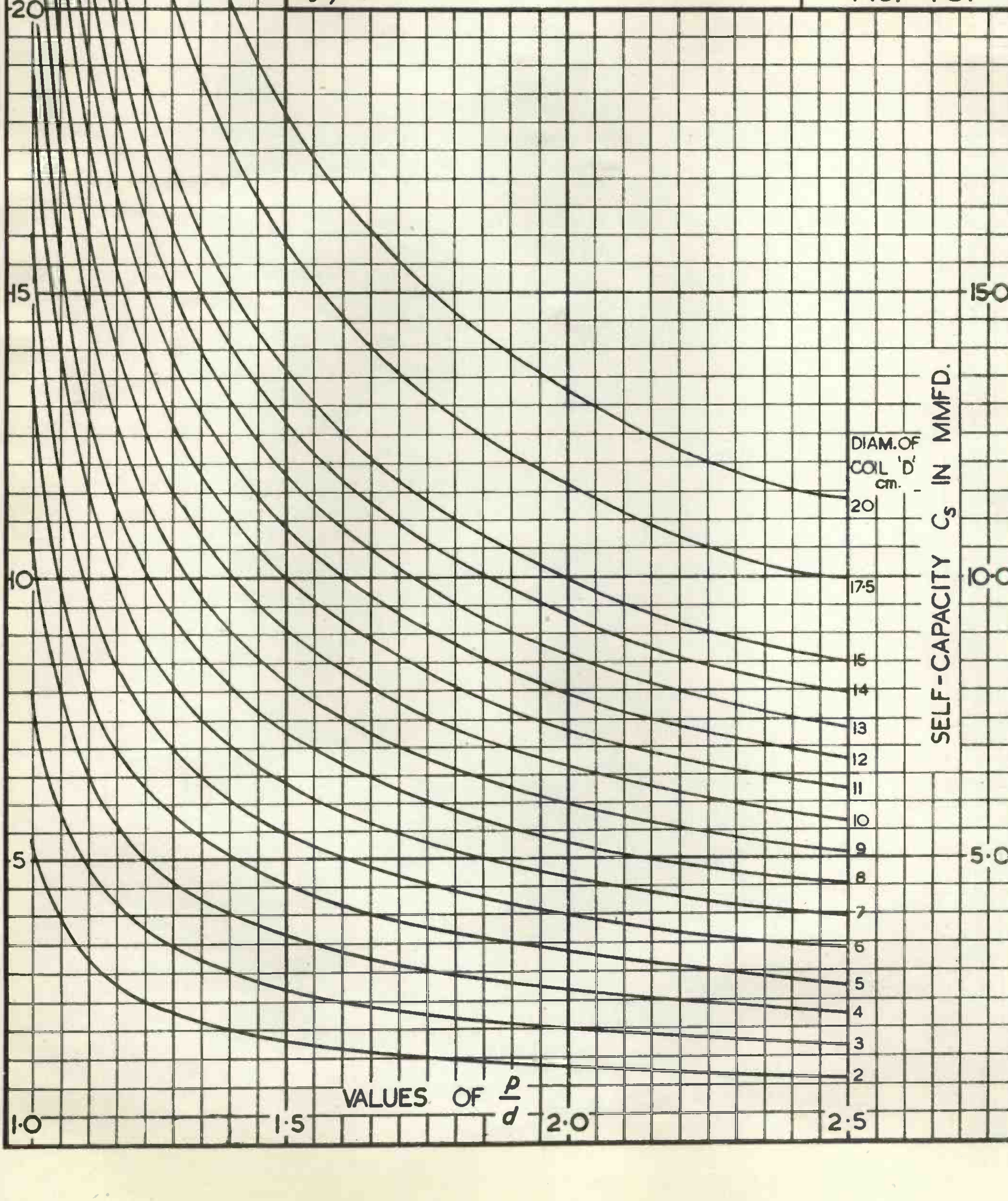
$$C_s \doteq \frac{\pi D}{3{\cdot}6 \cosh^{-1} \frac{p}{d}} \ \mu\mu F.$$

$l \not> D$



Electronic Engineering

DATA SHEET No. 18.



DATA SHEET XIX

Circuit Noise due to Thermal Agitation

Of all sources of noise present in a radio receiver the most fundamental and inevitable is that due to thermal agitation in conductors.

Thermal Agitation Noise is due to the free charge of any conductor being in random motion in equilibrium with the thermal motion of the molecules of the conductor, and this flow causes a random voltage to be developed across the terminals of the conductor.

The mean square value of this voltage is given by

$$\overline{E^2} = 4\,kT\int_0^{\infty} R(f)\,df \quad .. \quad (1)$$

where T = absolute temperature of the generalised impedance $R(f)$ in degrees Kelvin.

k = Boltzmann's constant 1.37×10^{-23} watt sec. per degree.

$R(f)$ = Resistive component of any generalised impedance at frequency f.

If the noise-producing resistance or impedance $R(f)$ is followed by an amplifier whose voltage amplification at any frequency f is given by $A(f)$ then the mean square noise voltage at the output terminals of the amplifier is given by

$$\overline{E^2_T} = 4\,k\,T\int_0^{\infty} R(f)|A(f)|^2 df \quad .. \quad (2)$$

If the frequency response of the amplifier is such that the voltage amplification is $|A|$ between the frequencies f_1 and f_2 and negligible at all other frequencies (*i.e.*, perfect band pass amplifier: *see* Fig. 1) then equation (2) may be simplified to

$$\overline{E^2_T} = 4\,k\,T|A|^2\int_{f_1}^{f_2} R(f)\,df \quad .. \quad (3)$$

Interpretation of R(f).

In equations (1) (2) and (3) ($R(f)$ represents the resistive component of any generalised impedance at a frequency f. The simplest possible case is that of a pure resistance R (*see* Fig. 1a) where $R(f) = R$ at all frequencies and equation (3) may be further simplified to

$$\overline{E^2_T} = 4\,k\,T|A|^2 R\,\Delta f \quad .. \quad (4)$$

where $\Delta f = (f_2 - f_1)$ = pass band of amplifier in c.p.s.

The case of a pure resistance can only be approximated in practice when the pass-band of the amplifier is in the low spectrum at other frequencies the stray capacities across the resistance must be taken into account (Fig. 1b). The resistive component of the impedance of the parallel RC combination is given by

$$R(f) = R \div (1 + \omega^2C^2R^2) \quad .. \quad (5)$$

and (3) becomes

$$\overline{E^2_T} = 4\,k\,T|A|^2 R\int_{f_1}^{f_2}\frac{df}{1 + \omega^2C^2R^2} \quad (6)$$

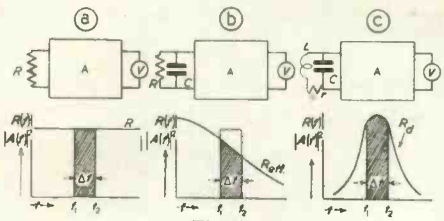


Fig. 1.

when the relative pass band is very narrow (*i.e.* $2\left(\dfrac{f_2 - f_1}{f_2 + f_1}\right)$ very small) expression (6) reverts to expression (4).

The most common source of thermal noise in receivers is due to the first tuned circuit (Fig. 1c). In normal circuits the dissipation loss of the tuning condenser is negligible compared to that of the inductance, and we can therefore concentrate all the resistance in the inductance branch. Under these conditions the resistive component $R(f)$ of the impedance is given by

$$R(f) = \frac{r}{(1 - \omega^2LC)^2 + \omega^2C^2r^2} \quad .. \quad (7)$$

and

$$\overline{E^2_T} = 4\,k\,T|A|^2 \int_{f_1}^{f_2}\frac{r\,df}{(1 - \omega^2LC)^2 + \omega^2C^2r^2} \quad (8)$$

However, in (7) the resistance r is itself a function of frequency. If as before the relative pass band is very small (7) can be considerably simplified for when the circuit is tuned to resonance,

i.e. $(1 - \omega^2LC) = O$ and

$$R(f) = 1 \div \omega^2_0C^2r = R_{\text{dyn}} \quad (9)$$

where R_{dyn} is the **Dynamic Resistance** of the circuit at the resonant frequency $\omega_0 = 2\,f_0$.

The resultant noise is therefore

$$\overline{E^2_T} = 4\,k\,T|A|^2 R_{\text{dyn}}\,\Delta f \quad .. \quad (10)$$

While the thermal noise output of any system may be computed by evaluating the integral of (2), in most cases the result may be estimated from experimentally obtainable information.

Taking again the case of the tuned circuit we can consider this from the point of view that the thermal noise voltage is wholly produced by the series resistance r of the inductance L. In the case of an amplifier with a relatively narrow pass band f this noise voltage is stepped up to a value V_1 by magnification of the tuned circuit $Q = (\omega_0 L)/r = 1/\omega Cr$

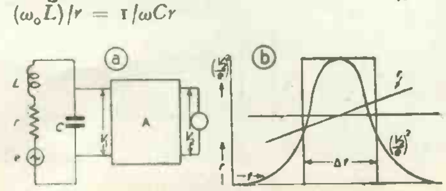


Fig. 2.

The mean square noise voltage at the output of the amplifier is then

$$\overline{E^2_T} = 4\,k\,T\,r\,(1/\omega_0 Cr)^2|A|^2\Delta f$$
$$= 4\,k\,T\,R_{\text{dyn}}|A|^2\,\Delta f \quad .. \quad .. \quad (11)$$

as before.

In the more general case where the relative pass band of the amplifier is not very small, the square of the voltage step-up of the tuned circuit is given by

$$1 \div (1 - \omega^2LC)^2 + (\omega Cr)^2 \quad .. \quad (12)$$

The product of this multiplied by the square of the voltage gain of the amplifier can be obtained by injecting a small signal from a very low resistance source (such as a Q-meter) in series with the inductance L and measuring the output voltage V_2 (*see* Fig. 2a). $(V_2/e)^2$ is then plotted against frequency (Fig. 2b) and the equivalent band width Δf is given by the width of the rectangle having the same height as and an area equal to that enclosed by the $(V_2/e)^2$ curve. As the resistance r will vary at some rate proportional to functions of $f^{\frac{1}{2}}$ and f^2 it is possible to estimate the equivalent value of r over the pass band by erecting another rectangle of width Δf for the curve of r against f. (Fig. 2b).

In many cases it is more important to be able to compare the relative values of the noise voltages due to circuit thermal and valve shot effects at the grid of first valve. To obtain the mean square thermal noise voltage at the first valve grid, we divide the height of the $(V_2/e)^2$ rectangle by the square of the amplifier gain at resonance $V_2/V_1 = |A|^2$. As the height of the new rectangle so obtained is $(1/\omega_0 Cr)^2$ the mean square noise voltage when $r \propto f$ is

$$\overline{E^2_T} = 4\,k\,T\,R_{\text{dyn}}\,\Delta f \quad .. \quad .. \quad (13)$$

In general the R.M.S. thermal noise voltage at the first grid will be given by

$$E_T = 7.4 \times 10^{-6}\,T^{\frac{1}{2}}R^{\frac{1}{2}}\,(\Delta f)^{\frac{1}{2}}\ \textit{microvolts} \quad (14)$$

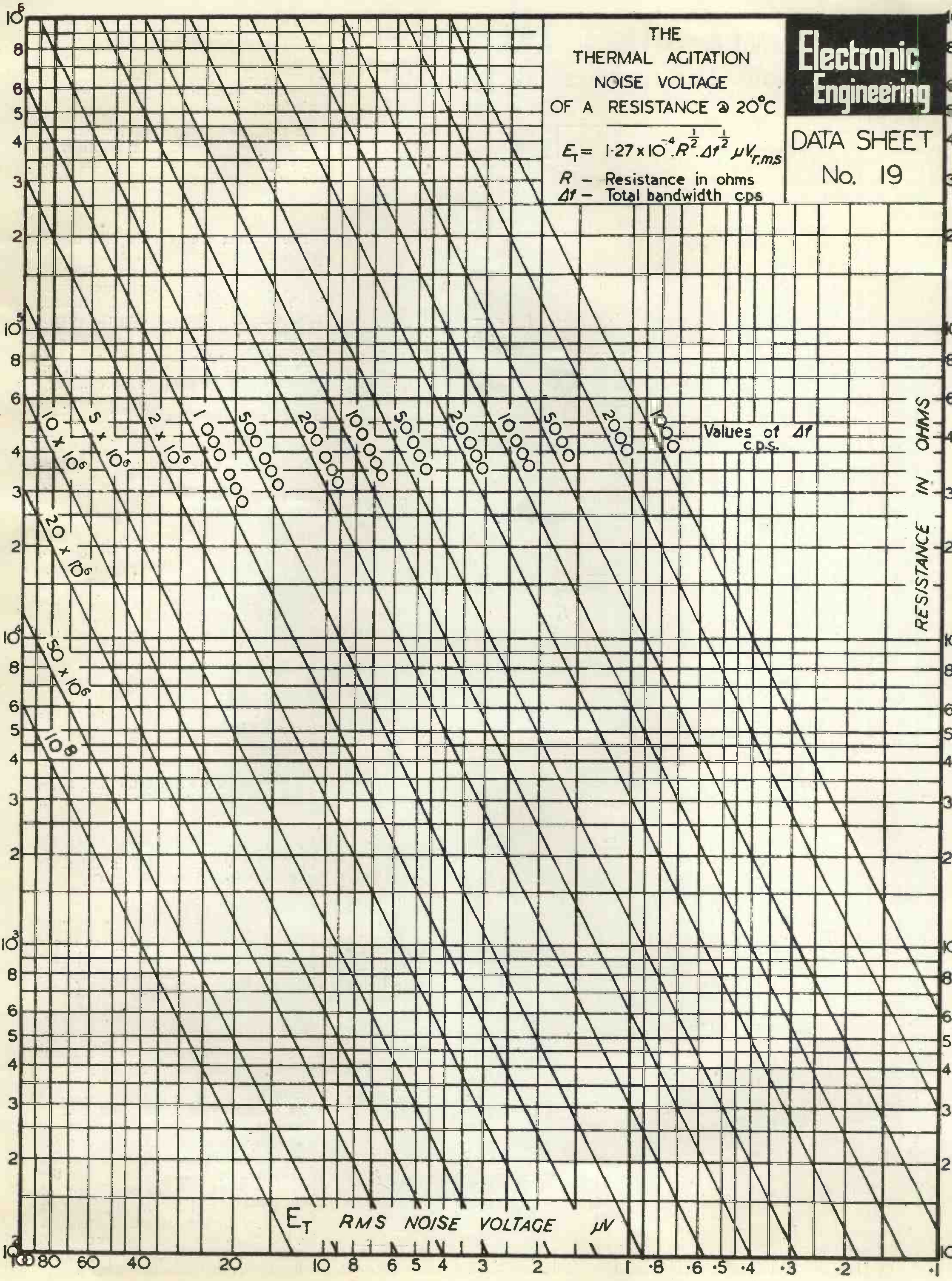
where Δf is in c.p.s and R in ohms. At a room temperature of 20° C.,

$T = (273 + 20)$ °K and

$$E_T = 1.27 \times 10^{-4}\,R^{\frac{1}{2}}\,(\Delta f)^{\frac{1}{2}}\ \textit{microvolts} \quad (15)$$

The approximate method given above is quite satisfactory in the usual case where the pass band is small compared with the mid-frequency. For greater accuracy where $R(f)$ may vary appreciably, the $(V_2/e)^2$ curve should be multiplied by the ordinates of the $R(f)$ curve and the rectangle erected about the resultant curve to determine Δf.

Equation (15) has been plotted as a series of curves on Data Sheet No. 19 where the R.M.S. thermal noise in μV across any resistance R is given for values of Δf between 1,000 c.p.s. and 10,000,000 c.p.s. Thus for example the noise produced by a resistance of 10,000 ohms would be 1.27 μV when it is followed by an amplifier with an effective pass band Δf = 10,000 c.p.s. and 28 μV when Δf = 5,000,000 c.p.s.


THE
THERMAL AGITATION
NOISE VOLTAGE
OF A RESISTANCE @ 20°C
$E_T = 1{\cdot}27 \times 10^{-4} . R^{\frac{1}{2}} . \Delta f^{\frac{1}{2}} \ \mu V_{r.m.s}$
R – Resistance in ohms
Δf – Total bandwidth c.p.s
Electronic Engineering
DATA SHEET
No. 19
Values of Δf c.p.s.
10 x 10⁶
5 x 10⁶
2 x 10⁶
1 000 000
500 000
200 000
100 000
50 000
20 000
10 000
5000
2000
1000
20 x 10⁶
50 x 10⁶
10⁸
RESISTANCE IN OHMS
E_T RMS NOISE VOLTAGE μV
100 80 60 40 20 10 8 6 5 4 3 2 1 ·8 ·6 ·5 ·4 ·3 ·2 ·1

Stable Admittance Neutralisation

IN resistance-coupled amplifiers a limit is set to the degree of amplification by the presence of stray capacity shunted across the anode-load resistance. A method for reducing the stray capacity, which may be quite large, as it is, for example, when contributed to by the capacitative component of admittance of a subsequent valve, has been proposed by Herold, of R.C.A. As originally suggested by Herold* the method suffered from the disadvantage of instability. Briefly, it was proposed by the use of an auxiliary valve possessing a negative mutual conductance to supply the capacity current that would otherwise require to be supplied by the valve working into the stray capacity. The auxiliary valve is arranged so that the stray capacity forms part of its input circuit admittance. Any potential applied to its grid appears in view of its negative conductance as a potential in like phase upon its anode, and if the latter is the larger, a current is fed from anode to grid when an impedance is connected between the two. If the impedance is capacitative a capacitative current will flow to the grid, and if of the right magnitude, can supply exactly the current that would normally be taken by the stray capacity from the preceding valve. Thus, if this right magnitude could be achieved, as far as the preceding valve is concerned, the stray capacity would be neutralised. It is found, however, that before neutralisation becomes complete, instability sets in, due to the feedback capacity which inevitably forms part of the anode load of the auxiliary valve, bringing about a phase lag in the anode potential and thus a certain degree of regenerative feedback in the grid circuit which may be sufficient to neutralise the positive resistive component of input admittance.

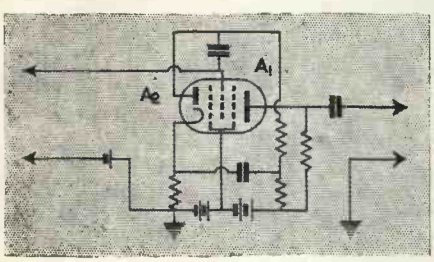


Fig. 2. The circuit of Fig. 1 adapted for balanced input.

It is also found, however, that a highly useful degree of neutralisation can be obtained by working very near to the oscillating point; but this is not a commercial proposition, since small fluctuations in the valve constants due to fluctuations in supplies, for example, are likely to occur in normal circumstances, which will cause the circuit to pass over into the oscillating state. It has now been suggested by Herold that this difficulty can be overcome by stabilising the auxiliary valve by the proper application to it of a sufficient degree of anti-phase, that is to say, purely degenerative, feedback.

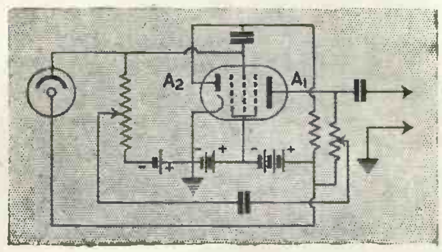


Fig. 1. Neutralisation by means of a special valve.

Various ways in which the stabilisation can be carried out are indicated in the figures shown.

Fig. 1 shows an arrangement in which a special valve co-operates with a photo-cell, where for the sake of uniform sensitivity over a wide range of frequencies, it is quite clear that stray capacity should be removed as far as possible from across the load resistance of the photo-cell. The special valve is of a dual character, and provides two outputs, one with a positive and one with a negative transconductance. It consists of a cathode surrounded by two screening grids, between which lies a control grid, and outside these grids lies the anode A_1 having a positive transconductance associated with it. the anode indicated as A_2 is located between the cathode and the innermost screening grid, and actually consists of two conductors arranged on either side of the cathode. In operation, as the potential of the control grid becomes increasingly negative, less electrons pass on to the anode A_1, but more are reflected back to the anode A_2. Thus the anode A_2 has associated with it a negative transconductance, and by connecting it to the control grid through the condenser C_m neutralisation of capacity associated with the control grid can be carried out to a large degree. Between the anode A_1 and the control grid it will be noticed that there is a degenerative feedback connexion through the condenser C_0, and this has the necessary effect of stabilising the current in the valve.

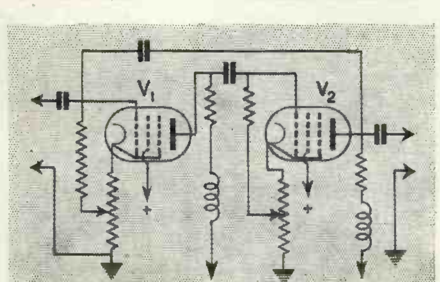


Fig. 4. The special valve of Fig. 1 replaced by two valves in cascade.

Fig. 2 simply shows a circuit of the kind as described, but adapted for working with a balanced input.

Fig. 3 indicates a slight modification, feedback being obtained by a resistance connected in the cathode lead of the valve. So that degenerative feedback is not applied to the auxiliary portion of the valve, that is to say, so that there is no negative feedback from the anode A_2, a condenser is arranged to by-pass current from the anode A_2 directly to the cathode, so that it does not flow through the cathode resistance.

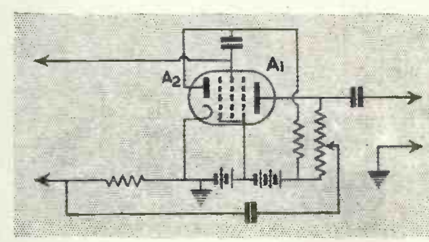


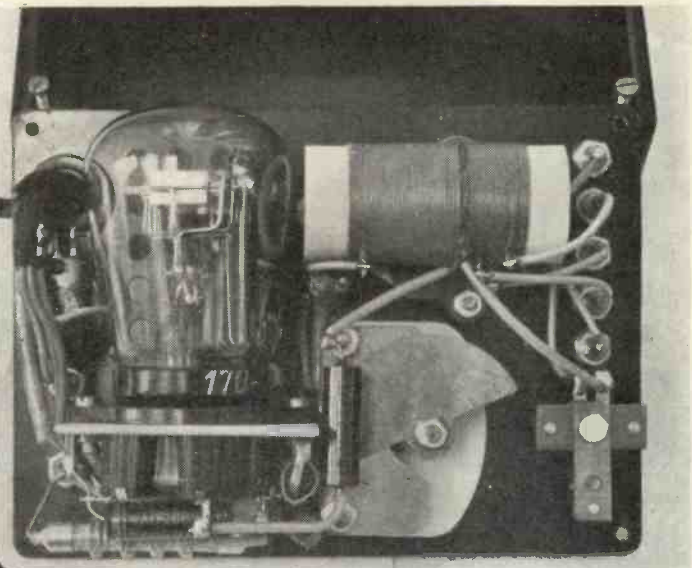
Fig. 3. A modified circuit in which feedback is obtained from the cathode resistance.

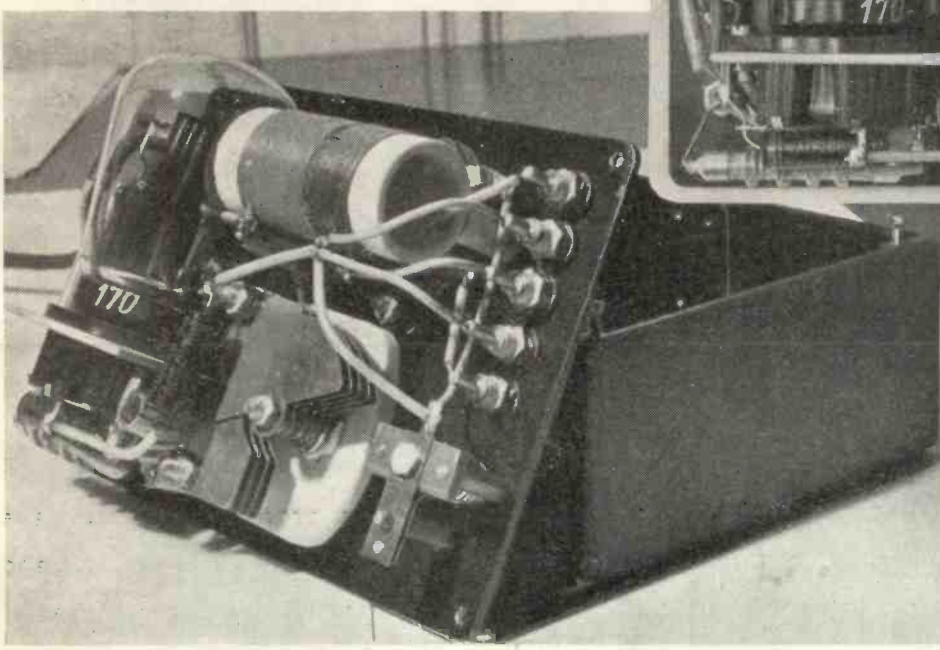
Fig. 4 illustrates an arrangement where the functions of the special valve are performed by an arrangement of two valves, V_1 and V_2, in cascade. Both valves have applied to them negative feedback to stabilise their characteristics, this feedback being obtained by cathode resistance, and the capacity-neutralising currents are fed back from the anode of V_2 to the control grid of V_1, through condenser C_m.

With circuits such as have been described, it has been found possible to reduce stray capacity to such an extent that an increase in gain of some forty times can be achieved as compared with conventional circuits. This increase in gain is equivalent to an additional stage of amplification, and in the arrangement of Fig. 1 it has been found possible to employ a load resistance for the photo-cell of 23 megohms for a cut-off frequency of 20,000 cycles per second.

* *Proc. I.R.E.*, November, 1937, p. 1399.

A GERMAN PORTABLE TRANSMITTER





The photographs show two views of the interior of a single valve portable transmitter used by enemy agents.

The aerial tuning coil is wound on a Steatite former and connexion to the batteries is made by the multicore cable seen on the left of the upper photograph.

The valve used is the Telefunken KL2.

(By permission of the Ministry of Aircraft Production)

BOOK REVIEWS

Thermionic Valves in Modern Radio Receivers.

A. T. Witts, 213 pp. 135 figs. (Pitman—10/6 net.)

This is the second edition of this book, which first appeared in 1936. Notes have been added on the output tetrode and beam valves in addition to negative feedback. An entirely new chapter on mains rectifiers has also been included in this edition.

The book deals with the theory and practice of valves and valve circuits in a thorough and practical manner, commencing with the theory of thermionic emission and dealing in turn with detectors, high- and low-frequency amplifiers, output valves and frequency changers. The chapter on push-pull output stages is particularly good and covers Class A and Class B operation and the variety of paraphase circuits.

Some of the diagrams would be improved by the addition of typical component values, but there is an excuse for omitting these in a book which aims to deal with fundamentals rather than specific design.

The criticism sometimes made by American reviewers that there are insufficient diagrams in British textbooks cannot be levelled at those of Mr. Witts. There is a diagram on every other page, which adds considerably to the value of the book to the student.

Radio Upkeep and Repairs for Amateurs.

By the same Author. 212 pp. 136 figs. (Pitman—6/6 net.)

The popularity of Mr. Witts as an author is shown by the fact that this is the fifth edition, the first appearing in 1933.

Chapters are included on common faults and how to clear them, testing components, care of batteries and accumulators, and fitting pickups to receivers.

A new and useful chapter has been inserted at the beginning on test equipment and how to construct it, giving full details on the assembly of a universal meter and test box. A second new chapter deals with design of mains receivers.

The book is full of useful suggestions which will be appreciated by the service engineer in addition to the amateur—in fact, the last two words of the title might be omitted from future editions!

Radio Engineers Pocket Book.

F. J. Camm. 132 pp. (Geo. Newnes—3/6 net.)

Many readers who collected the tables and data sheets issued some time ago in *Practical Wireless* may not appreciate that this pocket book contains that information and more in a convenient booklet form. In addition to the usual radio design data the book contains tables of physical constants, logs, and mechanical data (B.A. sizes, twist drills, etc.).

The copper wire and eureka wire tables are particularly useful with the inclusion of a column of "inches per ohm" and "pounds per ohm." On p. 84, however, in the table of musical frequencies, the author has not altered the scale to conform with the new standard A 440 (C′ 523) which was agreed internationally in 1939. No doubt this will be corrected in the next edition.

An Improved Hard Valve Time Base

by M. G. Saunders

ONE of the first successful hard valve time bases was the well-known "squegger" circuit of Watson-Watt, Appleton and Herd. The main disadvantages lay in a slightly more complex layout and difficulty in synchronisation. Further, the output voltage was small, necessitating amplification in most cases.

Reid* has described a modification of the "squegger" which has a far greater output voltage, is more stable and can utilise the high negative voltage across the cathode-ray tube to charge the time base condensers through a resistance. The basic circuit is shown in Fig. 1 and for further details the article referred to should be consulted.

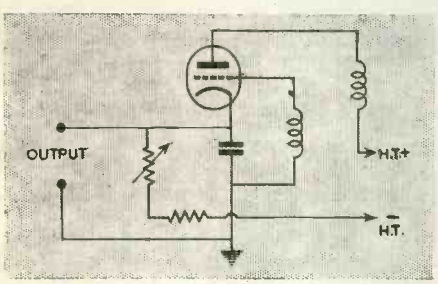


Fig. 1.

With a pentode (Fig. 2) in place of the triode, Reid used the suppressor grid as a point at which to inject the synchronising signal. The suppressor grid is connected to the cathode condenser, however, and it is across this that the sweep voltage develops, resulting in feed-back to the synchronising source and if this is the vertical amplifier it causes serious intermodulation. To overcome this the following circuit, with a separate synchronising valve may be used. (Fig. 3).

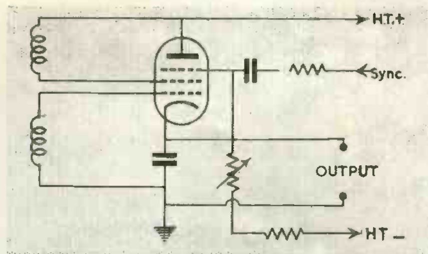


Fig. 2.

This circuit is a further modification of Reid's in certain respects—the time base valve is now a triode and the anode coil is parallel fed, with an H.F. choke as the anode impedance. The synchronising valve is fed from a resistance common to the two valves and acts as a modulator, an impulse on the grid varying the anode current and so the voltage drop across the resistance. This in turn varies the sweep voltage and hence the sweep frequency; with a sweep frequency similar to (or a multiple of) the synchronising frequency there will be strong locking between the two.

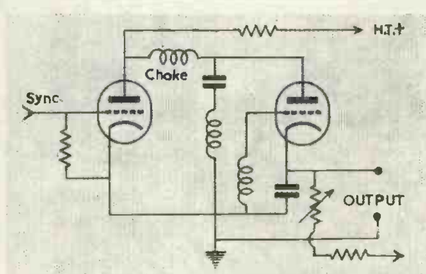


Fig. 3.

A circuit used in practice is shown in Fig. 4. Any valves will do provided that V_1 will oscillate at high frequencies; V_2 can quite well be a small output pentode. Coil L consists, in this case, of 8-10 turns of 16 gauge copper wire, tapped to earth at a point to give maximum sweep voltage. The choke is an Eddystone U.S.W.H.F. type, R_1 some 5,000 ohms, R_2 to suit the sync. valve and R_3 a potentiometer to vary the sweep voltage.

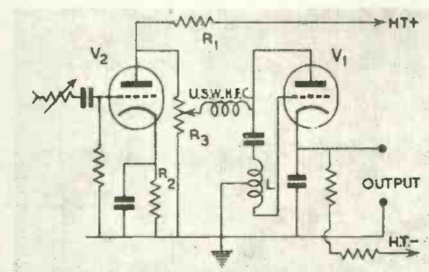


Fig. 4.

The sweep voltage is high, up to half or more of the anode voltage. The speeds can vary from the lowest to two or three megacycles with careful design; the synchronising voltage is very low with negligible power consumption, locking sweep frequencies twenty to thirty times slower, and finally the fly back is practically invisible except at excessive brilliance.

*(*Wireless World,* April 14, 1938, p. 334).

Frequency Modulation

(*Continued from page 585*)

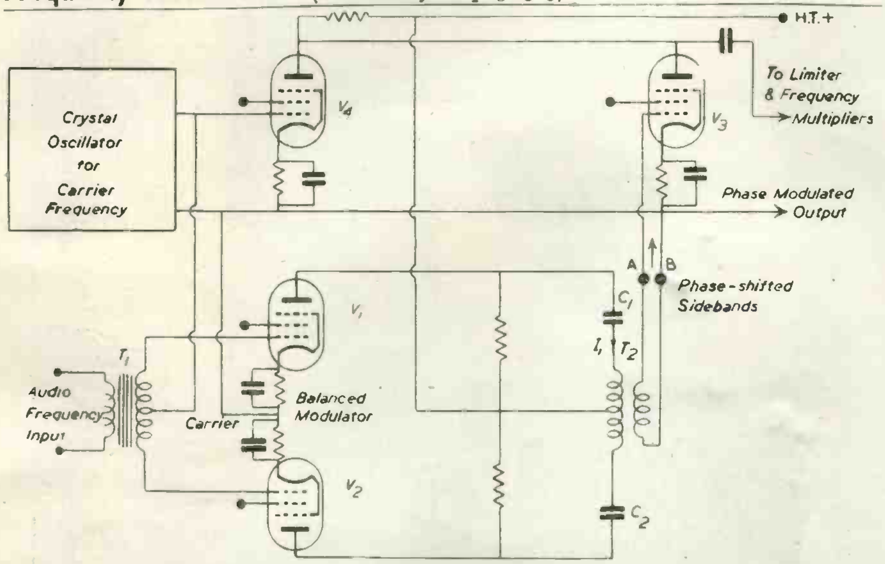


Fig. 5. Circuit diagram showing production of phase modulation.

"central" component is applied to another hexode V_3 (acting as a frequency changer) together with the output from a very stable crystal oscillator (multiplied up if necessary). The difference frequency is passed to a discriminator, in this case two circuits tuned about 2 kc/s above and below the difference frequency and two detectors giving opposing D.C. voltages which provide bias for V_2 through the grid leak R_1. When the original carrier frequency is correct, the difference frequency is exactly centred between the two discriminator circuits and there is no D.C. voltage across XX. If the carrier frequency component wanders, a voltage, positive or negative, is produced and automatically adjusts the D.C. bias on the reactance valve to correct for this. The carrier frequency component has then practically the same stability as the crystal oscillator.

Phase modulation can be achieved by separating the two sidebands of an amplitude modulated carrier from the carrier and passing them through a phase adjusting network which places their resultant at 90° to the carrier vector as shown in Fig. 4. This method does not completely suppress amplitude modulation (a limiter can be incorporated to remove amplitude modulation) and the phase change is not exactly proportional to the modulating voltage amplitude, but if its maximum value is limited to about ± 25° distortion of the modulation does not exceed 5%[3]. Now if 30 c.p.s. modulation is to produce a carrier frequency deviation of +75 kc/s at 40 Mc/s, a phase change of

$$143{,}200^\circ \left[\frac{75{,}000 \times 360}{30 \times 2\pi}\right] \text{ is required, so}$$

that the phase change of 25° must be multiplied 5,720 times, *i.e.*, the original carrier frequency must be 40/5,720 = 7 kc/s. Actually it is rarely necessary to cater for full modulation at 30 c.p.s. and a more normal carrier frequency would be 70 kc/s followed by frequency multiplier stages stepping up to 40 Mc/s. A circuit showing this type of phase modulator is given in Fig. 5. The carrier frequency, obtained from a crystal controlled master oscillator is fed to two channels, one which amplifies and one which modulates, suppresses the carrier and phase changes by 90° the sideband resultant with respect to the original carrier. The modulator is of the balanced push-pull type; the carrier frequency, being connected in the common grid lead, produces cancelling voltages in the anode and is suppressed. The sideband frequencies obtained by interaction between the carrier voltage and the audio voltages applied through the centre tapped transformer T_1, add at the output circuit, the primary of the centre tapped transformer T_2. The inductive reactance of the latter is cancelled by condensers C_1 and C_2 so that the output circuit acts as a resistance and the sideband current I_1 in the primary is in phase with the carrier grid input. The required 90° (or 270°) phase shift of the sidebands is, however, realised at the secondary because the voltage across AB is 90° (or 270°, depending on the direction of winding of the secondary with respect to the primary) out of phase with respect to the current I_1.

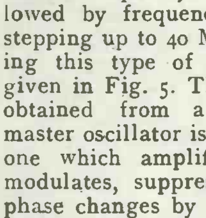


Fig. 4.

The sideband voltage is amplified in V_3 and mixed with the original carrier frequency in the anode of V_4, after which the resultant phase modulated wave is amplified and passed through frequency multipliers bringing the carrier up to its required ultra high frequency value.

Bibliography.

1 "Amplitude, Phase and Frequency Modulation," H. R. Roder, Discussion, *Proc. I.R.E.*, May, 1932, p. 887.
2 "A Method of Reducing Disturbances in Radio Signalling by a System of Frequency Modulation," E. H. Armstrong, *Proc. I.R.E.*, May, 1936, p. 689.
3 "Armstrong's Frequency Modulator," D. L. Jaffe, *Proc. I.R.E.*, April, 1938, p. 475.
4 "Reactance Tube Frequency Modulators," M. G. Crosby, *R.C.A. Review*, July, 1940, p. 89.
5 "A New Broadcast Transmitter Circuit Design for Frequency Modulation," J. F. Morrison, *Proc. I.R.E.*, October, 1940, p. 444.

Television Picture Storage

(*Continued from page 580*)

mobilised images are successively focused. Finally, the photo-electric screen on to which the wave images are projected, could be used in the form of a photo-electric control grid to control electron beams issued by a thermionic surface and passing through this grid.

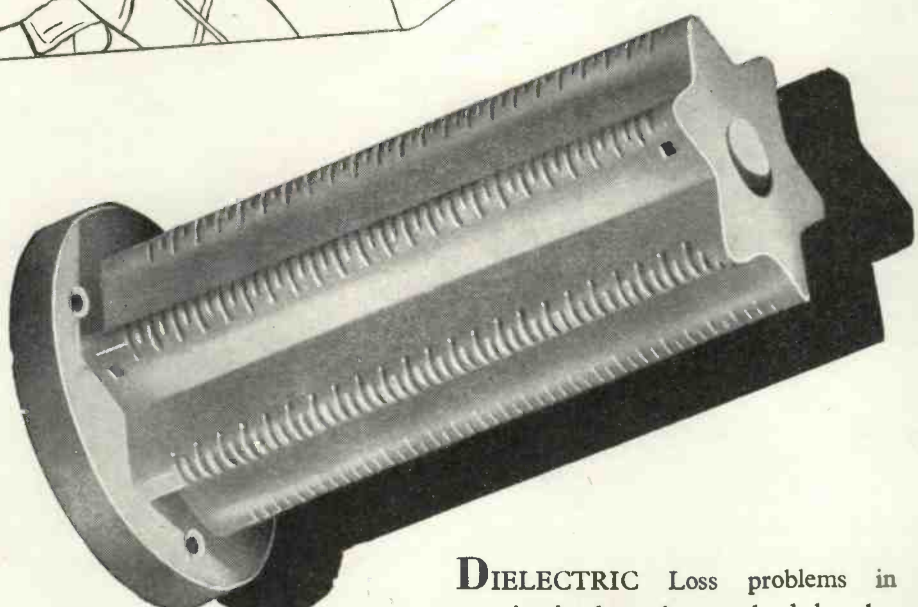
Using these latter means it will be possible to obtain a considerably higher light output from the fluorescent screen and/or to use a considerably smaller light flux as well as illumination on the photo-electric screen.

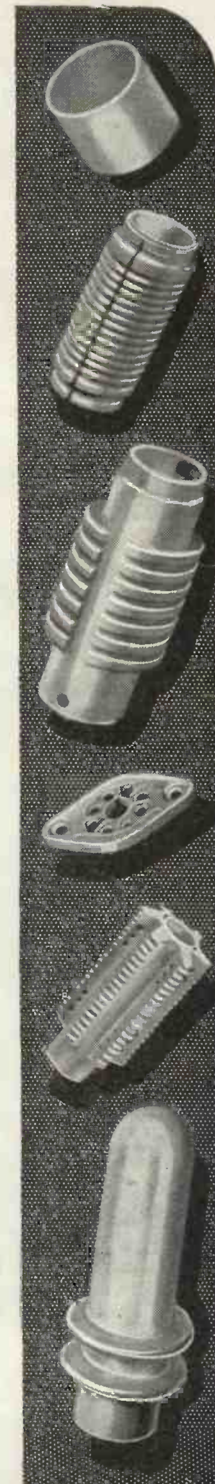
The system just described in principle is equivalent to a multi-beam cathode-ray tube of several hundred independent scanning beams. At any moment an area of the picture screen representing a considerable part of the picture line is struck by a fan-shaped beam of the exciting electrons. The total beam current, instead of being concentrated in a spot of element size, is distributed over this oblong area of several hundred elements, which is swept over the picture screen, exciting each element for a time period of the order of the line scanning period without any compromise to sacrifice definition. Consequently, even with much higher beam currents than feasible in single-spot cathode-ray tubes, it is possible to remain safely below the screen saturation. This will result not only in an increased life of the screen and in a fully defined picture the brightness of which, as quantitatively outlined above, will be considerably increased, but also improve the tone gradation in the high-lights since the image screen is no longer saturated.

Bibliography.

1 A. H. Rosenthal, *Proc. I.R.E.*, May, 1940, *Electronics and Television*, February and March, 1940, pp. 52, 117.
2 A. H. Rosenthal, *Electronics*, October, 1941, p. 41.
D. M. Robinson, *Proc. I.R.E.*, Aug., 1939.
3 DuMont Proposals, *Electronics*, February, 1940, p. 22.
4 *Television*, R.C.A. Institutes Technical Press, October, 1937.
5 E. Bruche, O. Scherzer, *Geometrische Electronen Optik*, pp. 130, 199, Berlin, 1934.

"Everything O.K. Sir !"





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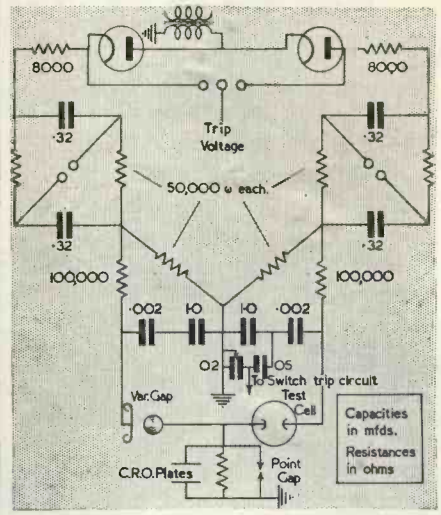
ABSTRACTS OF ELECTRONIC LITERATURE

Measurement

Measurement of Dielectric Pre-Breakdown Current
(Race)



The relatively long-time constants of even highly sensitive indicating instruments make the latter unsuitable for measuring current surges of the order of milli-seconds. The author describes the adaptation of the cathode-ray tube to indicate current changes just preceding breakdown of heptane and various oils. The apparatus, which is capable of applying 200 micro-second impulse voltage waves with gradually increasing unidirectional voltages to the specimens, is described, and results of such tests tabulated.

—*El. Engg.*, August, 1941, page 855.*

The Equivalent Characteristics of Valves Operating in Feedback Circuits
(J. H. Pratt)

A graphical method of determining the effects of feedback upon the characteristics of a valve amplifier is described, depending upon the fact that the characteristics of valves may be considered to be changed by feedback.

The method of obtaining valve characteristics with feedback from the normal characteristics is described, and a number of examples of "feedback characteristics" for various percentages of positive and negative feedback of both the voltage and current type are illustrated.

—*R.C.A. Review,* Vol. VI, No. 1, July, 1941, page 102.

Circuits

4-Valve Band-Switching Receiver for A.C. Mains
(S. K. Lewer)

A receiver is described in which the novel features incorporated are—selector switches in tandem and a ganged condenser with both large and small sections.

The receiver gives complete coverage from 100 kc/s to 26 Mc/s has a directly calibrated dial. A full circuit diagram is given together with a complete list of components.

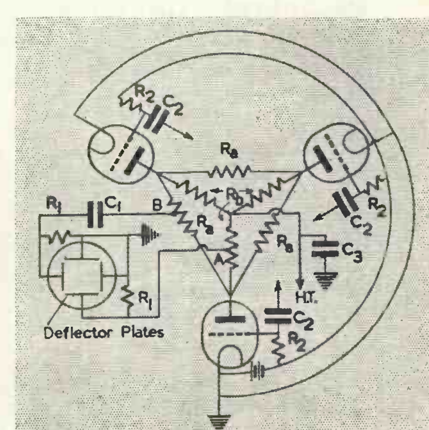
—*T. and R. Bulletin,* Vol. 17, No. 5, November, 1941, page 160.

Electro-Medical

Three-phase Electrocardiograph Mixing Circuit
(W. H. Jordan)

An electrical mixing circuit which permits the use of an ordinary cathode-ray tube having two sets of deflecting plates was suggested by the circuits used to change three-phase A.C. to two-phase A.C., and was tested by applying three-phase A.C. to the grids and observing a circular pattern on the screen of the cathode-ray tube. The resistance net-work was tapped at two places and connected to the vertical and horizontal deflecting plates. The position of the taps was chosen as shown in order to obtain the correct phase and amplitude relationships, i.e., when three-phase A.C. is applied to the grids, two-phase A.C. will be obtained on the deflecting plates. Pentode type valves were used to increase the gain.

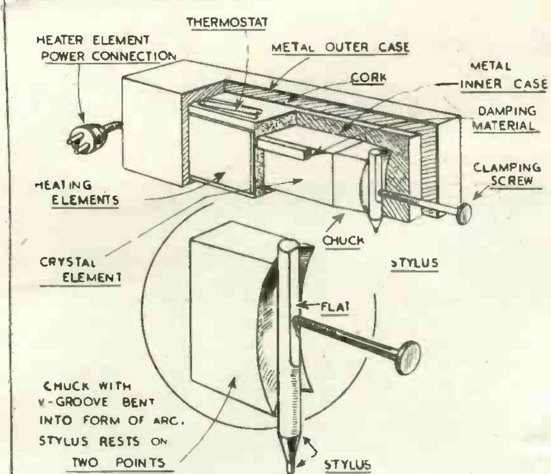
High gain amplifiers similar to those described by Hollman and Mann were used in the input circuit and were Y-connected to the conventional three-body leads. The output of the three amplifiers is connected to the three grids of the mixing circuit as shown in the diagram.



R1 and R2	1 meg.		
Ra and Rb.	50,000 ohms.		
C1 and C2	2μFd.		
C3	8μFd.		
Tappings	(A) 0.58 Rb.	(B) 0.33 Ra.	

—*Rev. Sci. Inst.*, Vol 12, No. 9, September, 1941, page 449.

* Supplied by the courtesy of Metropolitan-Vickers Elec. Co., Ltd., Trafford Park, Manchester.



A New Temperature-controlled Crystal Cutting Head

by

D. W. ALDOUS

AN investigation into the effects of temperature changes upon the various elements in the disk recording process, with special reference to the effects of temperature on electromagnetic and crystal cutting devices, was undertaken some time ago by the technical staff of The Brush Development Co., Cleveland, Ohio, U.S.A., directed by Dr. S. J. Begun.

It was found that the number of factors affecting cutting-head performance could be considerably reduced by maintaining the temperature of the cutting-head constant. This will stabilise the characteristic of all the variables inside the head, and leave the external resistance component as the only variable, representing the variations due to depth of groove-cut and differences in record material. If this remaining variable resistance component, for the frequency range considered, is small compared with any of the other three components, it can have little control upon the overall characteristic of the cutting-head.

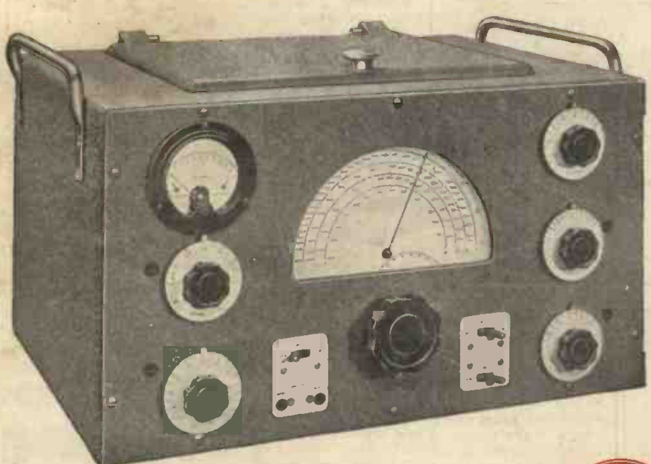
As devices utilising Rochelle salt crystals are the special interest of The Brush Development Company, the investigations were continued with the object of producing a crystal cutting-head possessing temperature stability. The practical outcome of these experiments was a design of crystal cutting-head with the desirable feature of temperature stability obtained by means of a built-in heating element with thermostat control. (*See Fig.*)

This temperature-controlled cutting-head consists of a four-ply crystal element mounted in a metal casing on "koroseal" pads. The "koroseal" pads are the damping and not only provide a stiff mounting support but, because of their damping characteristic, practically eliminate the natural frequency peak of the cutting-head The stylus chuck, which has a V-groove bent into the form of an arc thus allowing the stylus to rest on two points, is connected to one end of the crystal element. The stylus is pressed against the V-groove by a set-screw mounted in the head casing. The axis of the set-screw is located on the neutral axis of the torsional motion of the crystal element. Two heating pads are fitted close to the head housing sidewalls, controlled by a thermostat which rests upon and has good metallic connection to the cutting-head housing. The heater coils are designed for operation from the usual power supply. The complete unit is enclosed in another heat-insulated metal shielding to ensure that any draught or air motion in the recording room will not affect the temperature stabilisation of the crystal inside the housing.

After connecting the heater coils to the power supply the temperature of the cutting-head will rise and usually above the limit set for the thermostat. A definite time is required for heat equalisation, due to heat conduction, thus causing this temperature overshooting, but soon the thermostat operates and within 10 to 12 minutes the cutting-head arrives at the proper temperature. The crystal element itself introduces a small amount of distortion which changes with temperature, but this distortion is small at and above 35 deg. C. and accordingly a temperature of that value has been adopted for stabilisation purposes. No hum problems have been encountered, as electro-magnetic induction cannot take place in a crystal unit and, where electro-static influences are concerned, the cutting-head is completely screened in its housing.

The capacity of the crystal until is 0.004 μF. and to extend the range for constant-amplitude recording to 9,000 c/s the head should be fed from a 4,000 ohms impedance with 6 to 8 watts of output power. If a constant-velocity characteristic is desired for a turnover point of 500 c/s, the necessary output power from an amplifier may be as low as 2 to 3 watts. Portable equipment, for exterior recordings, fitted with this new cutting-head, can now operate with the same stability as studio equipment.

Band-pass crystal filters are available on both the EDDYSTONE '358' & '400'



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of the Eddystone "358" Communication Receiver, and its counterpart the medium frequency Model "400" are available with circuits incorporating band-pass crystal filter units. Full details will be sent on request. Both these special models are supplied to priority order only.

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