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Electronic Engineering

INCORPORATING ELECTRONICS, TELEVISION AND SHORT WAVE WORLD

PRINCIPAL CONTENTS

2/- OCT., 1942

Near technical perfection is achieved through use of scientific instruments but the trained eyes of skilled workmen inspect completed units before they are passed along to the pumps

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In the fabrication of plates, sealing of stems and leads, winding of grids...every tiny part must pass the rigid inspection of trained individuals, precision testing devices. At the end of each production line sits a group of hardboiled inspectors. All this checking and testing takes place before Eimac valves reach the vacuum pumps. That's one of many reasons why Eimac valves possess such uniformity of characteristics . . . why their performance records have made them first choice among world's leading engineers.

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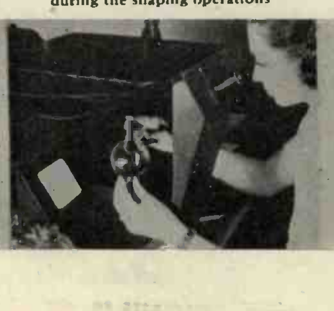
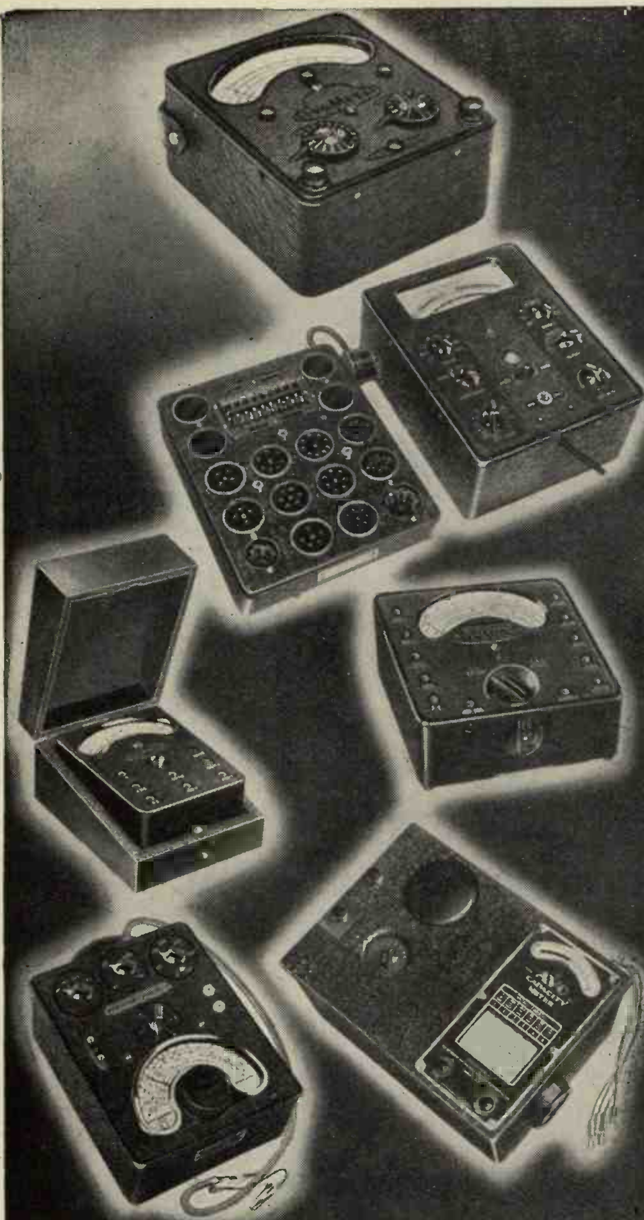
Bead tester utilizes polarized light in search for stress points in glass beads which seal leads to bulbs

Polariscope is here used to inspect glass bulbs for flaws or strain which may occur during the shaping operations



General inspection bench where completed filament stems and assemblies are thoroughly checked for faulty construction



MINISTRY OF SUPPLY

Defence (General) Regulations. 1939 No. 55. S.R. & D. 1942. No. 1770

AN ORDER

RUBBER MUST NOT BE DESTROYED THROWN AWAY OR MIXED WITH REFUSE

The Minister of Supply in exercise of the powers conferred on him by Regulation 55 of the Defence (General) Regulations 1939, has made an Order which provides as follows :—

1 No person shall, except under the authority of and in accordance with a licence granted or a special or general direction issued by the Minister of Supply,

(a) destroy any waste rubber
(b) throw away or abandon any waste rubber
(c) put any waste rubber in a refuse bin or other receptacle used for domestic or trade refuse, or
(d) cause or permit any waste rubber awaiting or in the course of collection or sale to be or become mixed with any material or article other than waste rubber.

Provided that nothing in this Order shall prohibit or restrict the destruction of any waste rubber if and so far as necessary in the course of any process of manufacture, or for the purpose of saving property from immediate danger of destruction or damage by fire.

2 (1) The holder of any licence granted under this Order shall comply with any conditions contained or incorporated in the licence.

(2) If any licence so granted is revoked by the Minister of Supply, the holder of the licence shall forthwith deliver up the licence to the Minister or as directed by him

3 IN THIS ORDER :

"*rubber*" includes reclaimed rubber, liquid latex, gutta percha and balata, and "*waste rubber*" means any worn-out, disused, discarded or waste material or article of the classes or descriptions specified in the Schedule to this Order, but does not include any material or article which is injurious to health or otherwise offensive.

4 This Order shall come into force on the 7th day of September, 1942, and may be cited as the Salvage of Waste Materials (No. 4) Order 1942.

THE SCHEDULE

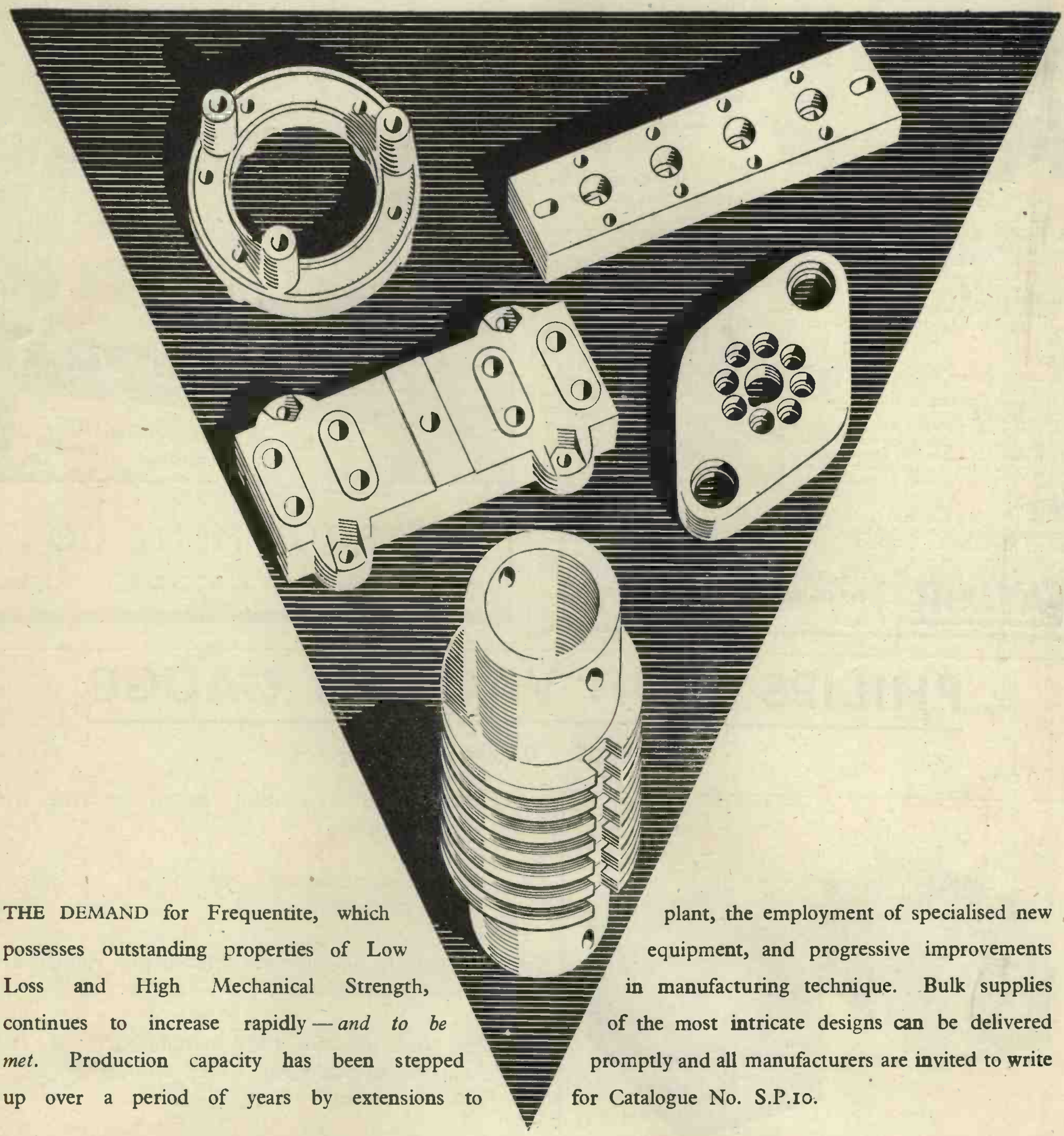
(a) *Articles or materials of any of the following descriptions made wholly or partly of rubber :—*

Balloons • Balls — sports and toy • Bathing-caps • Beds — inflated and sponge • Bulbs — horn, surgical, etc. • Carpet underlay • Catapult strip • Corks and Closures • Corsets — all rubber • Crepe-soled Footwear • Cushions — inflated and sponge • Door-stops • Draught-excluders • Ear Plugs • Elastic Cord and Thread • Electric Cable and Wiring • Football Bladders • Footwear (including Wellingtons and Gum Boots) • Flooring and Tiling • Gloves • Goloshes • Grips — handlebars, etc. • Horse-shoes and pads • Hose and Tubing • Hot-water bottles • Jar-rings • Mats and Matting • Mattresses — inflated and sponge • Milking Rubbers • Pedal-rubbers • Radiator Hose • Rollers • Soles and Heels • Sponge-backed Flooring • Sponges • Squeegees • Stair-treads and nosings • Stopper-rings • Teapot Spouts • Teats and Soothers • Tobacco Pouches • Tubing • Tyres — Pneumatic (Covers, Tubes and Flaps), Solid and Cushion of all types • Upholstery.

(b) *Articles or materials of any other description made wholly or mainly of rubber.*

THIS ORDER CAME INTO FORCE SEPT. 7

FREQUENTITE



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LONDON.

Training Physicists

THE Board of the Institute of Physics has recently issued a memorandum on the Post-war Education and Training of Physicists, prepared by DR. H. LOWERY, of the S.W. Essex Technical College.

As stated in the preface, the memo. has been issued with the object of provoking discussion on the vital problems of the training of physicists and of the education in physics of those not specialising in the subject.

It is Dr. Lowery's experience that the greatest shortage of physicists is found in the industrial development side, and it is frequently found that engineers, chemists and biologists are performing the work of a trained physicist in this sphere.

" A moments consideration will show that this grade of physicist must have a very wide (but possibly not deep) knowledge of his physics and also of industrial processes and practice ; moreover it would appear that the ordinary academical training of " University plus research " is inadequate for the purpose of obtaining the right product."

The author emphasises that the prime function of the Technical College is not the coaching of students for the external degrees of the London University and any tendency to make them into cramming institutions should be resisted.

On the contrary, the modern Technical College is a people's University and should be concerned with vocational studies for the vast number of individuals who will occupy posts in industry covering a wide range of skill—the " backbone of industry."

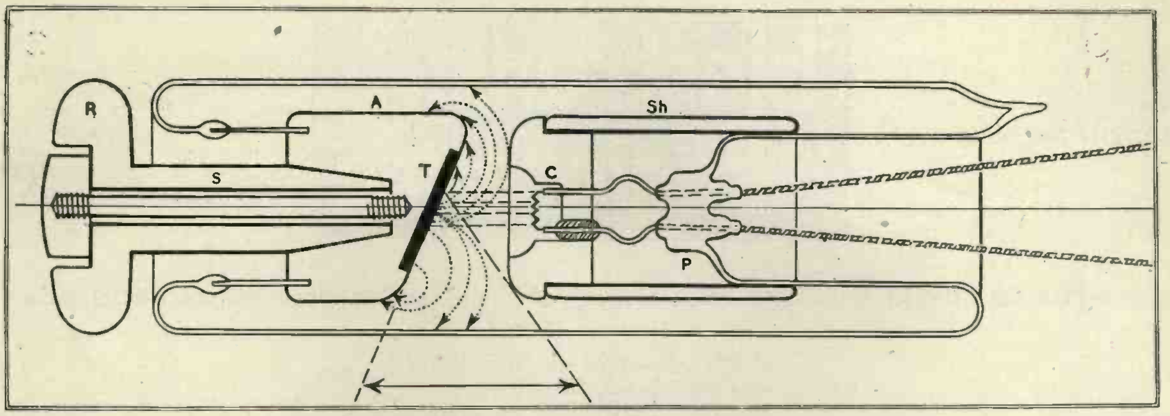
An interesting suggestion is that the teaching of such subjects as geography and physiography should be undertaken by qualified physics teachers in schools, and that courses in appreciation of physics should be given at training colleges for teachers.

Another point which the author stresses is that whatever scheme is specified for training it should include provision for social intercourse and opportunities for discussion so as to enable the individual to express himself freely and to appreciate the other man's point of view—in other words, self-expression must be the keynote of the training.

As Principal of one of the most up-to-date and well-equipped Technical Colleges in the country, Dr. Lowery speaks with authority and his views should be read by all concerned with applied physics, including the radio engineer.

Electronics and physics have a common unit in the electron and it is as much to the interest of the electronic engineer to ensure that the training for his profession is on the soundest lines as it is for the physicist.

The Institute of Physics is arranging for an open discussion to take place on the subject in London on October 12th., at a time and place to be announced later. Those interested can obtain copies of the memorandum from the Secretary of the Institute, Dr. Lang, at the University of Reading, Berks., and written communications on the subject are invited.



The Target

Factors Affecting the Design of Targets for X-Ray Tubes

By A. G. LONG*

THE starting point in the design of X-Ray tubes is the shape and general geometry of the electrodes and these are principally governed by the type of work the tube will be expected to do. The most important of these is the target, as the results obtained from the tube depend on its design and reliability. Medical diagnostic tubes may be called upon to operate under very severe conditions and from some points of view present some of the most difficult problems. The full load conditions are most exacting as under some conditions the tube may be operated at a peak anode voltage of some 100 kV and must be capable of releasing an electronic current of 0.8 amperes or more. These extreme conditions represent electrical loads of no mean magnitude as they are in the neighbourhood of 40 kW or 55 h.p.—sufficient energy to melt a 1 lb. block of copper in five seconds! The time period over which tubes are able to take these loads is governed mainly by the time taken by the target to reach limiting temperature.

Tungsten Targets

Tungsten is principally used for the target material for the following reasons: It has a low vapour pressure at high temperature, and a high melting point. Further, its heat conductivity is reasonably good. The temperature limit for tungsten in good vacuum is in the region of 2,800° C. This figure can only be very approximate as it is modified by a number of relatively controllable factors. The purity of the tungsten used to form the target must be of a very high order, and relatively small amounts of impurity will modify both its physical and general behaviour to a marked degree. Spectrographic means are usually employed to gain knowledge of its impurities and X-Ray back reflexion will show up its physical condition. A beam of X-Rays from a suitable target is passed along a collimator tube of small internal diameter. (Fig. 1). The bundle of X-Rays that leaves the end of the collimating tube is allowed to bombard the face of the material under examination. The bombarding X-Rays are refracted in all directions, but not in equal amounts, and the quantity refracted in various directions will be determined by the crystal faces in the material and their relative positions to the primary X-Ray beam. From the crystallogram it is possible to gain much useful information such as the crystal size and the preferred or random orientation of the crystals. If the crystallogram shows the crystals to have random orientation the material is in a strain-free state and in general mechanically strong, *i.e.*, work free. On the other hand, if the crystallogram reveals the crystals to have preferred orientation, which means that the crystals are lying mainly in one direction the material is in a strained condition and is mechanically weak. It must, of course, be remembered that the examination only extends to a depth of a few or perhaps only one crystal. Fig. 2 is a crystallogram produced by bombarding a copper specimen with X-Rays from a pure copper target. This shows medium grain size with random orientation of the crystals. It has been shown in a most convincing manner that a relatively small amount of even another refractory metal will result in an unsound physical condition. The impurities will broaden the melting point and also set up complex thermal stresses that promote early fatigue failure. If these impurities are included in the target material they will in some cases boil off at temperatures much lower than the target temperature limit, and this will almost certainly mean a release of the gas.

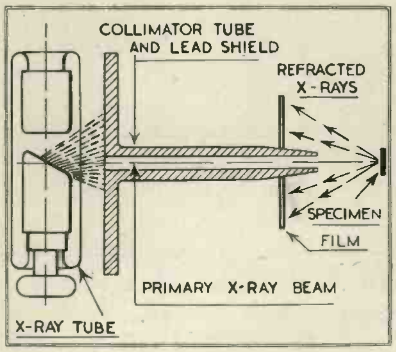


Fig. 1. Method of obtaining crystallograms by back reflexion.

* Messrs. Newton and Wright, Ltd.

The sectional drawing above shows the construction of a typical X-Ray tube with glass envelope. The cathode C is mounted on a glass pinch P, surrounded by a shield Sh. The anode A, in the centre of which is the target T, is sealed in at the opposite end of the tube. S is the anode stem and R the radiator. The path of the secondary electrons from the anode is shown by the dotted lines, the path of the X-Rays by the broken lines.

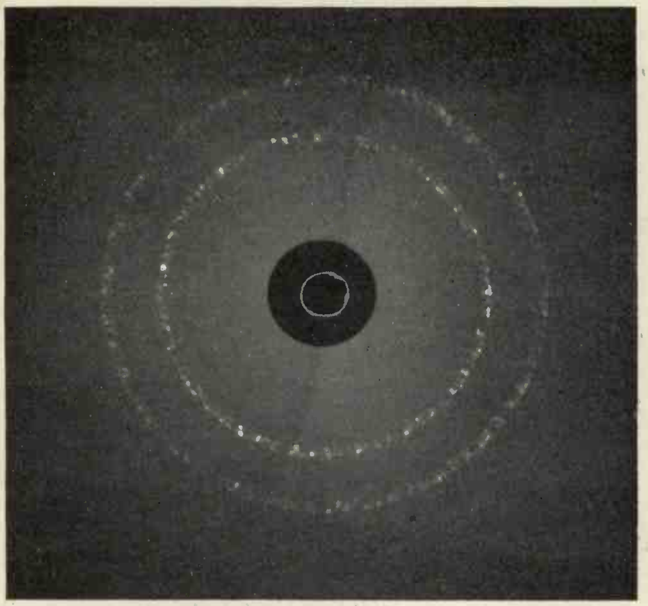


Fig. 2. Crystallogram obtained by the method of Fig. 1.

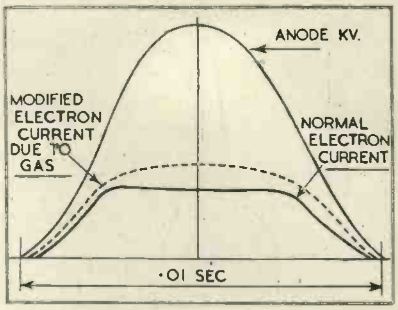


Fig. 3. Effect of gas in increasing emission over a half-cycle of anode voltage.

Effect of Gas.

If the natural clean-up of gas is in excess of or is equal to the speed of release, little damage will result, but if the release of gas is in excess of the natural clean up speed a number of complex conditions can develop. In the first place the presence of gas in the anode-cathode space will promote the generation of positive ions due to their collision with the high velocity electrons in this area. If the number of gas molecules in the anode-cathode space is such as to allow the relatively small electrons to move from cathode to anode without collision the amount of gas is of no importance, but if the gas density (the number of gas molecules present in this area) is sufficient to bring about collisions between the fast moving electrons and gas molecules there is a possibility (if the electron has enough energy), that it will remove an electron from the gas molecule. The removed electron and the bombarding primary electron will in most cases move to the anode, while the positive ion resulting from the collision will move to the cathode. This positive ion having a large mass in comparison with the electron will even with slow acceleration do two things when it reaches the cathode area. In the first place the surface of the filament that is subjected to ion bombardment will increase in temperature and so increase the electron release. (Fig. 3). In the second place the presence of positive ions in the cathode area will modify the space charge condition of the current curve and, should there be a further increase in the number of molecules per unit area, cumulative ionisation may develop. Cumulative ionisation is a condition where ions can release further electrons from further gas molecules by collision. To favour this condition the cathode-to-anode potential drop must be high and also the gas density must be of sufficient value to allow for the necessary collisions. There are many stages between these extremes each with its own particular behaviour. As each stage progresses the pure electron emission becomes decreasingly unimportant until when complete cumulative ionisation sets in the electron emission becomes overshadowed. Under these conditions the impedance is reduced to almost zero value, and currents are no longer limited by filament temperature. Needless to say a tube even approaching this condition is rendered useless. Before leaving this incomplete survey of ionisation conditions it might prove of help to say that the recombination of positive ions with free electrons helps to delay the growth of the process to a large extent.

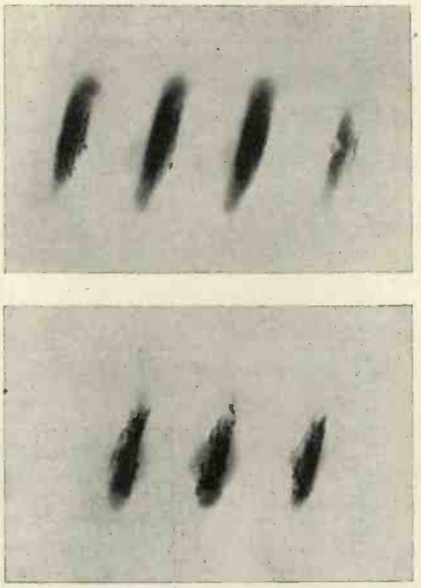


Fig. 4. Photographs of tube filaments. Above: Good filaments. Below: After ion bombardment.

Filament destruction by ion bombardment will take place both by vaporisation of the filament surface due to temperature increase from ion bombardment and also by pure cathode sputtering.

The photographs of Fig. 4 show a series of coiled filaments before and after ionic bombardment. The roughening and pitting can be clearly seen in the lower group of filaments.

A small release of gas would normally be interpreted as operation instability. This instability should not be confused with an increase in tube current near the end of a short-period running time at heavy load which is caused by the target increasing the filament temperature by direct radiation of heat. This condition is quite normal and can always be observed when tubes are used under conditions approaching full load for short time duration. Observations of this condition can be undertaken by a cathode ray oscillograph, using current deflec-

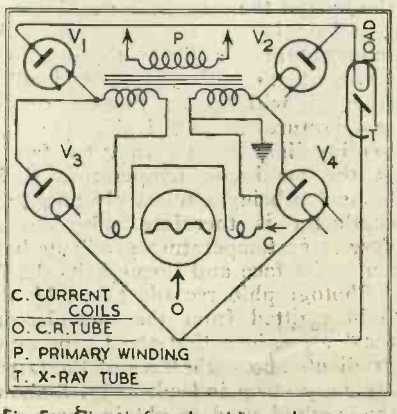
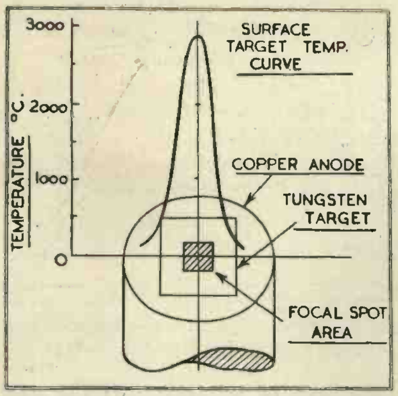
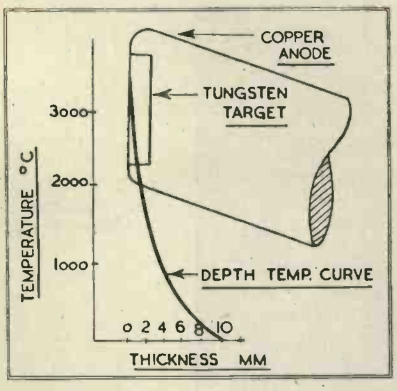


Fig. 5. Circuit for observing tube current.





Figs. 6 (left) and 7 (right). Temperature curves of a tungsten-copper target showing variation across the surface and with target thickness.

tion connected in the mid point of the H.T. transformer secondary. The tube filament transformer should be closely stabilised against voltage variation, or a drop in tube filament current will tend to off-set the observed results. A suggested circuit is given in Fig. 5. In cases where it is not possible to connect into the H.T. transformer secondary the deflecting coils can be connected into the H.T. transformer primary though there will be considerable distortion of the observed wave shape due to the magnetising current, etc.

The physical condition of the tungsten target will in many ways modify the safe maximum loading. At high temperature the physical structure undergoes a marked change, the crystal size increasing and the number of crystal boundaries decreasing. The importance of this physical change will be apparent from the following: When the tungsten target is receiving bombardment from high velocity electrons over only a part of its area, heat is generated in this area at an extremely rapid rate.

Figs. 6 and 7 give indication of the temperature gradients encountered with X-Ray tube targets. Fig. 6 showing the temperature gradient across the face of the target after a high current load for one sec., and Fig. 7 the temperature gradient through the target depth with the same loading time. It will be seen that the copper temperature is just inside the safe loading limit and the tungsten face is at the maximum temperature limit. Tungsten being a relatively poor heat conductor is therefore subjected to very steep temperature gradients both across its face and through its depth.

Photographic records taken by the light emitted from the incandescent focal area show that the temperature gradients across the face of the target are very steep indeed. Fig. 8 shows the method used to obtain these pictures; the lead glass filter prevents the X-Rays from reaching the photographic plate in the camera and the polar screen prevents to some degree the light radiation from the tube filament over-exposing the record. A record on line A gives the normal X-Ray view and a record on line B gives the almost true view. It will now be evident that the stresses produced by unequal thermal expansion are very severe.

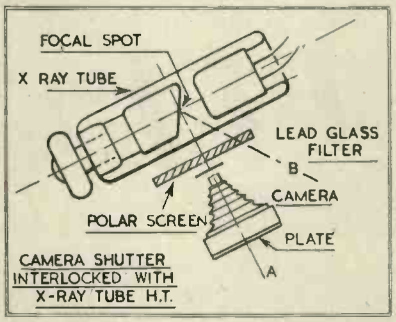


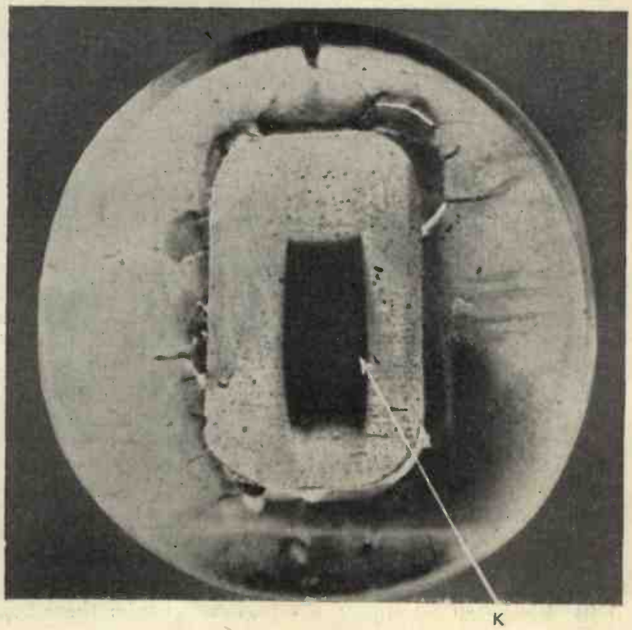
Fig. 8. Method of obtaining photograph of target.

Copper-Tungsten Targets

It is standard practice in most tubes to attach the tungsten target to a copper block which forms the anode proper. The thermal contact between the copper and the tungsten must be sound and moreover must be able to withstand considerable strain resulting from the differential expansion of the two materials. Copper has been selected chiefly on account of its rapid heat conduction, though its melting point is unfortunately low, approximately 1,080° C. The ratio of heat conduction to that of tungsten is in the neighbourhood of 3-1.

Using pure copper it is important that the back of the tungsten target must not be allowed to reach a temperature of 1,080° C. or the copper will melt and set up a high mechanical strain due to excess thermal expansion with no room for movement. This condition often leads to the lifting of the tungsten target which brings about the destruction of the thermal contact, cracks the main copper body and releases a quantity of gas. Fig. 9 shows an example of target destruction due to copper expansion from an overload melt, and the cracks around the tungsten target can be clearly seen. In the bottom left-hand corner molten copper has escaped from behind the target. The mechanical damage to the tungsten face producing roughening can be seen in the area marked K. This roughening is in the main due to

Fig. 9. Photograph of tungsten target showing damage by expansion of the copper.



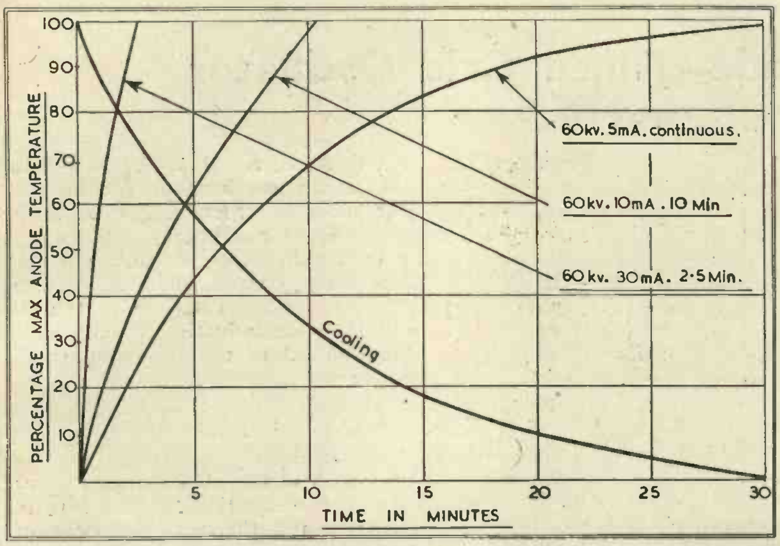


Fig. 10. Heating and cooling curves of target under various load conditions.

mechanical work resulting from electron bombardment. The thickness of the tungsten target must therefore be arranged so that the temperature at the back of the target will not reach the limit temperature for copper at maximum load. It might be considered that the risk of damage would be reduced by thickening the tungsten face, but since heat must be conveyed away from the target area as soon as possible, it will be seen that this defeats its own object. The tungsten thickness is governed by the use to which the tube will be put. If a given number of kW seconds are put into the target area to increase the temperature to 2,800° C. from room temperature, it is obvious that the same loading in kW seconds cannot be immediately repeated, as in all probability the starting temperature will be in the neighbourhood of 600° C. and under these conditions the temperature limit would be exceeded. In practice the thickness of the tungsten target is so arranged as to allow for a repeat of maximum load conditions in a reasonable time, *i.e.*, the interval time met with in practice. Alternatively, if the target were made thin to improve the heat conduction circuit and thereby increase the loading, this apparent improvement would steepen the temperature gradient through the tungsten and bring the copper into the danger zone before the limit temperature was reached at the tungsten face. The thickness of the tungsten is thus an important factor in the design of the tube. Mention was made earlier of the change in crystal size with temperature. This process is not a function of temperature alone, but is also related to time, and is cumulative. This change is accompanied by work fatigue brought about by the thermal stresses which in turn are the results of temperature gradients between the focal area and the surrounding metal, and if these stresses are not relieved a fracture may take place at the boundary of the bombardment area. This fracture is also in part due to irregular grain growth as the grain growth in the bombardment area is not of the same order as that in the surrounding tungsten.

Design

To design a target and be able to predict its maximum loading with a margin for overload, the following main items must be known.

Maximum kW seconds loading (short time period).
Size of focal spot area to be used.
Expected rest time between maximum loading (medical tube).
Screening conditions (long time periods 10-15 minutes).

kW-Time curves can be produced which are the result of the heating and cooling curves, and one such curve is shown in Fig. 10. Radiographic technique calls for a maximum X-Ray output from a minimum focal area, since all other factors remaining constant, the smaller the focal area the better the definition on the radiograph. This is, of course, the worst possible condition from the electronic design point of view, and one which the X-ray tube designer continually meets. For example, the rough figure for tungsten loading is 200 watts per sq. mm. for a period of one second, which gives a focal area of 4.1 mm. × 12.3 mm. for a 10 kW loading.

Heat loss by radiation from the focal area is so small a percentage of the total energy to be negligible from the point of view of loading, but it is of importance from the point of view of change in filament temperature as mentioned earlier. The emission of free electrons from tungsten commences at a temperature as low as 1,200° C. and from this it will be evident that a tube operating as its own rectifier (*i.e.*, no valves in the circuit) will reach a temperature limit condition long before the normal temperature limit. The target of a tube operating as a self-rectifier will never reach the tungsten limit temperature though the mechanical work on the target will be more than doubled for a given kW second loading. By careful design the loading over the bombarded area of the target can be unbalanced to good effect, and this condition is achieved by arranging the cathode focus so as to increase the electron density at the edges. This produces more heat at the edge of the target which can be conducted away across the face (Fig. 11) and under these conditions the normal loading for a given area can often be extended. Many methods are in use to conduct the

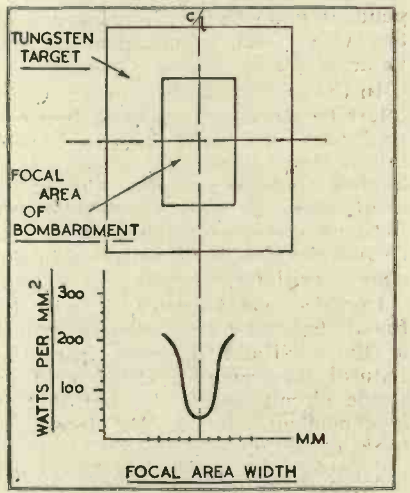
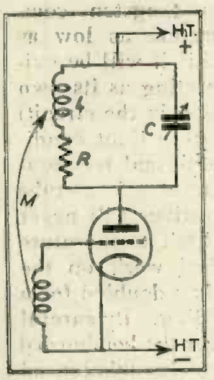


Fig. 11. Variation of power dissipation across the surface of the target.

generated heat away from the target head, the principal being as follows: (a) A radiator can be fitted to the stem of the copper anode which dissipates the heat to air. (b) Liquid cooling with water or oil and a combination of these is often used, or, in some tubes the radiator and tube are immersed in oil. It should be noted that approximately 99 per cent. of the energy put into the X-Ray tube is converted into heat, and the remaining 1 per cent. into useful X-Rays.

An Analysis of
The Tuned Anode—Tuned Grid Oscillator

By G. P. POLLARD, B.Sc.*



In the case of low or medium frequency oscillators, an oscillatory current started in an $L.C.R.$ circuit in the anode lead of a valve by attaching the H.T. + lead can be maintained by mutual coupling between the coil of the tuned circuit and a coil placed in the grid circuit, provided M in the figure above is of the correct sign and is equal to or greater than :

$$\frac{CR}{g_m} + \frac{L}{\mu}$$

The actual frequency of the oscillatory current will be

$$\frac{1}{2\pi}\sqrt{\frac{1+\frac{R}{r_a}}{LC}}$$

and is therefore dependent upon the A.C. resistance of the valve, but independent of the value of M.

In the case of high frequency oscillators the necessary feedback between grid and anode circuits can be obtained by utilising the grid-anode interelectrode capacity of the valve itself, the effect of which since it forms a virtual open circuit at low frequencies, has been neglected in the above oscillator.

In this case however, it will be found that not only is the frequency of the oscillator different from the natural frequency of either grid or anode circuit separately, but that it is dependent upon K, the interelectrode capacity.

To determine the frequency of the oscillator and the necessary relationship between the circuit values in order that oscillations may take place, we can replace the oscillator circuit shown in Fig. 2 by the equivalent A.C. circuit of Fig. 3, because the oscillator can be regarded as an amplifier supplying its own grid input.

Since the oscillator frequency will be different from the natural frequency of either of the tuned circuits, if Z_A and Z_g denote their respective impedances, either may be of the form $R + j\omega L$ or $R - j/\omega C$, depending on the relative values of natural frequency and oscillator frequency for either circuit.

Using the notation shown in Fig. 3, and applying Kirchhoff's laws to the circuit gives :—

$$\mu V_{gf} = I_a r_a + (I_a - I_o)Z_A$$

$$\therefore I_a(r_a + Z_A) - I_o Z_A = \mu V_{gf} \qquad (1)$$

$$(I_a - I_o)Z_A = I_o(Z_g - j/\omega K)$$

$$\therefore I_a = \frac{Z_A + Z_g - j/\omega K}{Z_A} I_o \qquad (2)$$

$$V_{gf} = -I_o Z_g \qquad (3)$$

where $\omega = 2\pi \times f$ and f = oscillator frequency.

From the above equations by eliminating I_a, I_o and V_{gf}, the relationship between f, f_A and f_g for oscillations to be possible and also the relationship which must appertain between the circuit values are obtained.

The elimination of I_a, I_o and V_{gf} gives the relationship

$$Z_A Z_g(\mu+1) + r_a(Z_A + Z_g) - j/\omega K(r_a + Z_A) = 0 \qquad (4)$$

To simplify the application of this equation it can be assumed that the nature of the tuning of the anode and grid circuits with respect to f, can be determined by considering the reactance components alone of Z_A and Z_g, since the including of resistance in these circuits whilst modifying the actual frequency of oscillation will not, in itself, render a circuit oscillatory which has been proved by consideration of reactance alone to be incapable of oscillation.

Bearing in mind that to produce an oscillatory circuit including K the reactance of at least one of the circuits must be inductive, we can apply equation (4) to the following conditions :—

(a) $Z_A = R_1 \quad Z_g = j\omega L_2$
(b) $Z_A = j\omega L_1 \quad Z_g = R_2$
(c) $Z_A = j\omega L_1 \quad Z_g = -j/\omega C_2$

In each of the above cases the real part of the equation (4) is a positive quantity and therefore cannot equal zero.

(d) $Z_A = -j/\omega C_1 \quad Z_g = j\omega L_2$

Substituting in equation (4) gives

$$\frac{L_2}{C_1}(\mu+1) + jr_a\left(\omega L_2 - \frac{1}{\omega C_1}\right) - j/\omega K(r_a - j/\omega C_1) = 0$$

which is satisfied if

$$\frac{L_2}{C_1}(\mu+1) - \frac{1}{\omega^2 C_1 K} = 0$$

$$\text{i.e., } \omega^2 = \frac{1}{KL_2(\mu+1)}$$

$$\text{and } r_a\left(\omega^2 L_2 - \frac{1}{C_1}\right) - \frac{r_a}{K} = 0$$

$$\text{i.e., } \omega^2 = \frac{1}{L_2}\left(\frac{1}{K} + \frac{1}{C_1}\right)$$

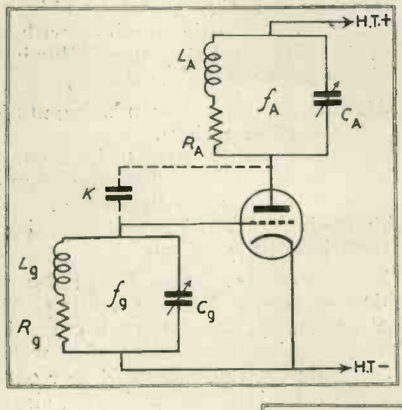
Equating these values of ω^2 gives

$$\frac{1}{L_2K(\mu+1)} = \frac{C_1+K}{L_2C_1K}$$

$$\therefore \frac{1}{\mu+1} = \frac{C_1+K}{C_1} = 1 + \frac{K}{C_1}$$

which is an impossible relationship.

The only remaining possibility to give a real solution to equation (4) is that the frequency of the oscillator will be such that both Z_A and Z_g will be of the form $R + j\omega L$.



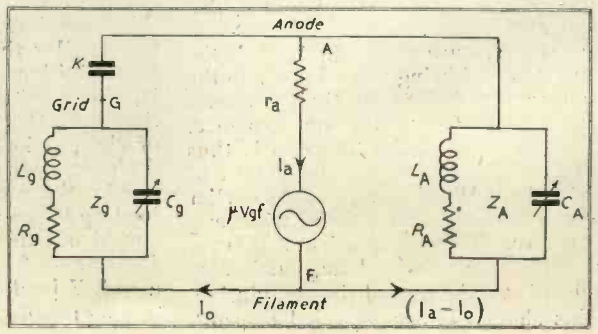


Fig. 2 (above), Circuit of tuned anode-tuned grid oscillator and Fig. 3 (right) the equivalent circuit.

*Northampton Polytechnic Institute, London.

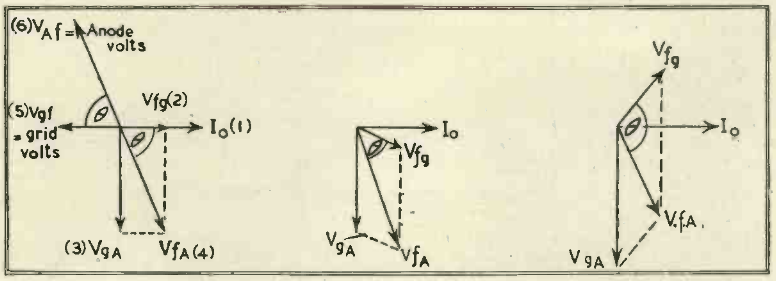


Fig. 4. Vector diagrams for the grid circuit of Fig. 3.

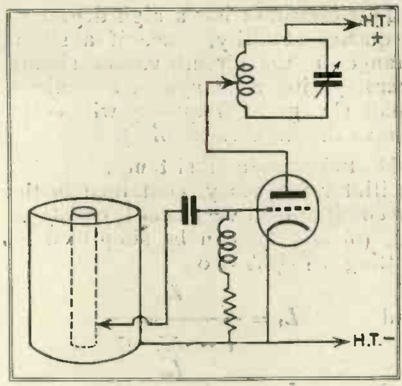


Fig. 6. Simple oscillator with ¼-wave transmission line in the grid circuit.

Assume $Z_A = R_1 + j\omega L_1$
and $Z_g = R_2 + j\omega L_2$

Substituting these values in equation (4) gives

$$(R_1 + j\omega L_1)(R_2 + j\omega L_2)(\mu + 1) + r_a\left\{R_1 + R_2 + j(\omega L_1 + \omega L_2)\right\} - j/\omega K(r_a + R_1 + j\omega L_1) = 0$$

which will be satisfied if, neglecting second powers of resistance

$$\omega^2 = \frac{R_1 + r_a}{K[(\mu+1)(R_1L_2 + R_2L_1) + r_a(L_1 + L_2)]} \quad (5)$$

and

$$\omega^2 = \frac{r_a(R_1 + R_2) + \dfrac{L_1}{K}}{L_1L_2(\mu + 1)} \quad (6)$$

Equating these values of ω^2 gives

$$\frac{r_a(R_1 + R_2)(L_1 + L_2)K}{\mu L_1 L_2} = 1 - \frac{L_1}{\mu L_2} - (\mu + 1)\frac{L_1 R_2}{\mu L_2 r_a}$$

and if $(\mu+1)L_1R_2 \ll \mu L_2 r_a$, which would appertain in practice, the condition for oscillations becomes

$$(R_1 + R_2)(L_1 + L_2)K = \frac{L_1}{r_a}(\mu L_2 - L_1) \quad (7)$$

These equations show that the oscillator frequency will be less than the natural frequency of either of the tuned circuits and that to alter the frequency of oscillations, the tuning of both circuits must be altered so that equation (7) will be satisfied at the new frequency.

Vectorial Analysis

If we assume that the variations of current and voltage in the oscillator circuit are purely sinusoidal, we can construct a vector diagram for the oscillator which will indicate the necessary tuning of the grid and anode circuits with respect to the oscillator frequency which render oscillations possible, since simple oscillator theory indicates that for oscillations to be possible a component of both grid volts and anode current must be in antiphase to anode volts.

Consider first the grid circuit.

(a) Resistively tuned : $f = f_g$.
(b) Capacitatively tuned : $f > f_g$
(c) Inductively tuned : $f < f_g$.

Fig. 4a, b and c. It will be seen that only in case (c) could the $\angle\theta$ which is equal to the angle between the vectors representing grid and anode volts, be greater than 90°.

Thus oscillations will be possible only when the grid circuit is inductively tuned with respect to f.

If we now construct a vector diagram (Fig. 5) for the oscillator circuit as a whole, the relationship between the vectors representing V_{fA} and $(I_a - I_o)$ to make the voltage V_{fg} in antiphase to μV_{gf} will indicate the necessary tuning of the anode circuit.

From the equivalent A.C. circuit

$$\mu V_{gf} = I_a r_a + V_{fA}$$

and

$$V_{fA} = V_{fg} + V_{gA}$$

whilst the current I_o will lag on voltage V_{fg} by $\angle\phi_g$

The relationship between the vectors representing grid volts (1), anode current (9) and anode volts (7) shows that oscillations are possible, and since V_{fA} leads $(I_a - I_o)$ by $\angle\phi_A$ less than 90°, the anode circuit must be inductively tuned with respect to the oscillator frequency.

If $f = f_g(1 - x)$ and x is a small fraction, circuit theory indicates that $\tan \phi_g = 2Qx$.

If therefore the Q factor of the grid circuit be made as high as possible by the use of a quarter wavelength of transmission line and if the coupling of this line to the grid of the valve be loose, an oscillator can be constructed the frequency of which will be almost wholly controlled by the grid circuit.

Simple circuits of this type of oscillator are shown in Figs. 6 and 7. In Fig. 6 a quarter wavelength of concentric feeder is used, and the grid attached as near the voltage node as possible.

A push-pull form of T.A.T.G. oscillator employing resonant lines in both grid and anode circuits is shown in Fig. 7.

The wider pipes in the grid circuit reduce the ohmic resistance at high frequency and hence increase the value of Q (in spite of the decrease in $\sqrt{L/C}$.)

The position of the sliders on the grid pipes should be as near to the shorting strip, S, as possible.

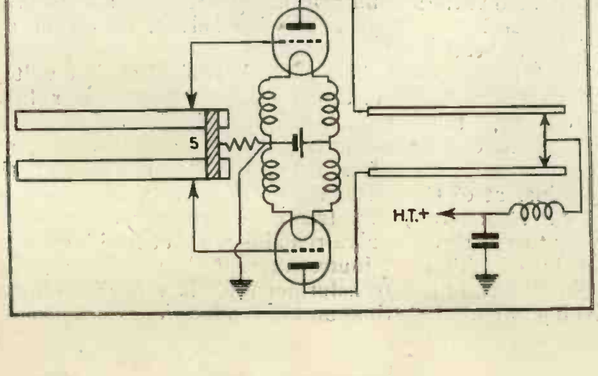
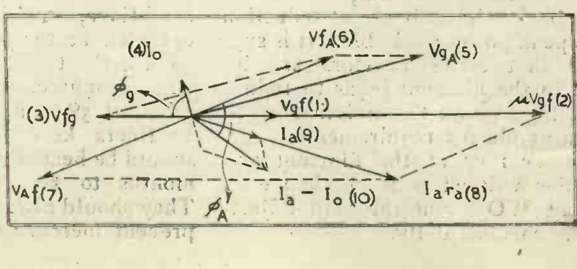


Fig. 5 (below). Vector diagram for complete oscillator circuit.

Fig. 7 (left). Push-pull T.A.—T.G. circuit using resonant lines.



This oscillator has a high degree of frequency stability, since if a slight change in the circuit values should occur, giving a change in ϕ_g, only a small change of frequency will occur due to the high value of Q.

If the ohmic resistance, at the oscillator frequency, contained in the tuned circuits is very small, equations (5), (6) and (7) can be simplified by putting $R_1 = R_2 = 0$

and $$L_1 = \frac{L_A}{1 - \omega^2 L_A C}$$

and $$L_2 = \frac{L_g}{1 - \omega^2 L_g C}$$

in which case the two values of ω^2 are :

$$\omega^2 = \frac{1}{K(L_1 + L_2)} \quad \text{... (5a)}$$

and $$\omega^2 = \frac{1}{KL_2(\mu + 1)} \quad \text{... (6a)}$$

and the condition for oscillation reduces to

$$L_1 = \mu L_2 \quad \text{..................... (7a)}$$

If Z_A and Z_g are pure inductances we find that $I_a = 0$ and the oscillatory current $I_0 = -\frac{V_{gf}}{j\omega L_2}$ is the current flowing in the series circuit comprising L_1 and L_2 and K across the H.T. battery.

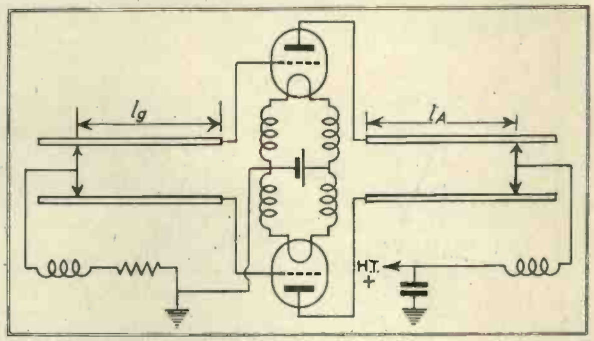


Fig. 8. Circuit using resonant lines as tuned circuits.

In the simplest form of oscillator using resonant lines as tuned circuits, a quarter wavelength of short circuited line is used for each tuned circuit. (Fig. 8).

If λ = wavelength of the oscillatory current then

$$Z_g = j\omega L_2 = jZ_0 \tan \frac{2\pi l_g}{\lambda}$$

and $$Z_A = j\omega L_1 = jZ_0 \tan \frac{2\pi l_A}{\lambda}$$

where Z_0 = characteristic impedance of the lines, assumed constant for both circuits.

From equation (7a) we obtain the relationship

$$\tan \frac{2\pi l_A}{\lambda} = \mu \tan \frac{2\pi l_g}{\lambda}$$

for oscillations to take place.

If I_a is small enough to be neglected, i.e., the ohmic resistance in the tuned circuits is negligible then

$$V_{fA} = \mu V_{gf} = -\mu\omega L_2 I_0$$

$$= -\frac{L_1}{L_2} \cdot \frac{L_2}{K(\mu + 1)L_2} I_0$$

$$= -\frac{I_0}{K(\mu + 1)} \cdot \frac{L_1}{L_2}$$

If a meter be included in the grid circuit so that I_0 can be kept constant, then the output voltage from the oscillator will be proportional to the ratio of the effective inductances of anode and grid circuits at the frequency of the oscillator.

Prolonging Valve Life

THE technical supervisor of the American stations WOR-W71NY, Mr. C. H. Singer, has contributed useful suggestions for prolonging the life of transmitting valves in an article reproduced in *Radio* (June, 1942).

Reviewing the more important factors that enter into the life of valves he shows how careful attention to these will eliminate many of the abuses that cause premature burnouts or loss of emission.

The filament voltage of the valve is of the greatest importance, and in tungsten filament transmitters the life is halved by increasing the filament voltage by 5 per cent. over its correct rating. Conversely, by underrunning the filament by 5 per cent., the life can be doubled, with only slight reduction in the peak power available. It is suggested that series resistors are inserted in the filament leads to reduce the voltage where the emission available is ample for requirements.

The accuracy of the filament and grid bias voltmeters is checked each week at WOR and the zero adjustment is checked daily.

In the case of thoriated tungsten valves the underrunning of the filament is governed by the peak currents drawn, and this type of valve should not be operated below the manufacturers rating as a general rule.

The recommended cure for loss of emission due to overload is the running of the filament at 30 per cent. above normal voltage (with the anode and grid voltages cut off) for approximately 10 minutes, after which the filament is run for 1-2 hours at normal voltage.

With mercury vapour rectifiers the filament voltage should never be allowed to drop below the rated value; if anything, 1 per cent. above is suggested. A five minute pre-heating period is recommended for normal conditions, but if the rectifiers are operated at temperatures lower than 20° C., the time should be increased. One manufacturer recommends 10 mins. at 5° C. and 15 mins. at 0° C. Rectifiers kept in stock as spares should be heated up every two or three months to keep them in condition. They should always be kept upright to prevent mercury splashing on to the cathode or anode.

The author strongly recommends the use of a " tube reconditioner " for transmitters which have become gassy, either from shelf life or during running, and gives details of a high voltage equipment which has given valuable service at the station.

This consists of a high voltage transformer to suit the rating of the valves employed, which is fitted with an adjustable magnetic core for varying the secondary voltage and a spark gap across the secondary winding.

The valve is connected across the high voltage winding, the filament and grid being strapped, and the voltage is applied betwen the anode and filament.

The valve is flashed on low voltage for a minute, the transformer being switched off each time that an arc tends to develop across the gap, and after running satisfactorily for this short time, the voltage is raised and the treatment continued until the valve remains without flashing for ten minutes after the last flashover. The transformer must be switched off each time an arc tends to develop.

Colour Television

By P. C. GOLDMARK, J. N. DYER, E. R. PIORE and J. M. HOLLYWOOD

The following paper is reproduced in abbreviated form from the original which appeared in the *Proceedings of the Institute of Radio Engineers,* April 1942, and covers the development of colour television by the Columbia Broadcasting System, N.Y. from 1940 to the present time. Daily colour transmissions were broadcast from WCBW from June 1941 for the purpose of testing the practicability of the system described in this paper.

IN the colour-television system under discussion the following terms will be used:

Colour-field frequency: the highest vertical scanning frequency employed in the system.

Colour-frame frequency: colour-repetition time per second; *i.e.,* tri-chromatic-repetition rate per second, corresponding to the colour-field frequency divided by 3.

Colour-picture frequency: number per second of the coincidence of one and the same primary colour with one and the same area of the image.

Frame-frequency: identical to the term used in monochromatic television, *i.e.,* completion of the scanning of the entire picture area per second in black and white.

Before the choice for a final system was narrowed down, several alternatives were considered. These all had in common sequential, additive-colour scanning where the primary-colour impulses of varying ratio, following in rapid succession, are integrated by the observer's eyes. The three primary colours employed were red, blue and green, the characteristics of which will be discussed later. Rotating colour disks or drums in front of the pick-up device and the receiving tube, suitably synchronised and phased, produced the colour analysis at the transmitter and the synthesis at the receiver.

It became evident that in order to increase definition interlacing had to be introduced. This led to a system (No. 3 in Table I) which has a colour-field frequency of 120, a colour-frame frequency of 60, and colour-picture frequency of 20 per second. Due to a 2 : 1 interlace ratio the frame frequency remains at 50 per second and the number of lines at 343. This system gave freedom from flicker with

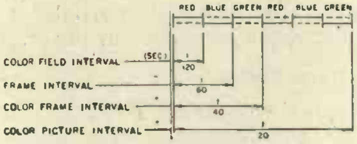


Fig. 1. Diagrammatic representation of CBS System 3.

screen brilliancies up to 2 apparent foot-candles and showed no interline flicker. It was subsequently chosen as the most satisfactory compromise for the present 6 megacycle band at the same time increasing the number of lines to 375 which corresponds to the 525 lines used in monochromatic television.

System 4 is a compromise between Systems 3 and 5 inasmuch as the colour-picture frequency is 15 with a corresponding colour-frame frequency of 60 and colour-field frequency of 180 per second. The frame frequency of 45 per second will permit a number of

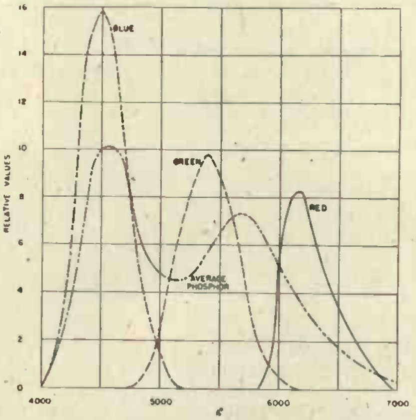


Fig. 2. Colour characteristics at the receiver: filters Nos. 47, 58 and 26 combined with the phosphor. The fourth curve represents an average phosphor.

lines approximating to 525 $\sqrt{30/45}$. In this system flicker even at the highest brilliances is eliminated; however, the interline flicker still appears somewhat excessive.

System 5 uses the same horizontal-scanning frequency as monochromatic television; but utilises quadruple interlacing to increase the field frequency to 120 per second. Thus flicker conditions are satisfactory and resolution is excellent. However, due to the low colour-picture frequency (10 per second) interline flicker appears excessive.

In order to avoid the so-called "line-crawling" effect the quadruple interlacing in Systems 4 and 5 is of the staggered type, the sequence of lines being 1, 3, 2, 4 instead of 1, 2, 3, 4. Conditions in these two systems are aggravated by the fact that colour-field frequency of 180, and 120 per second respectively, being a multiple of the power-supply frequency, would show a distinct breakdown of the line structure emphasising a raster of approximately 100 lines in case 60 cycle components were not completely eliminated from the vertical scanning.

The final decision in favour of System 3, with System 4 as a close second, had to be made in view of the discouraging results, confirmed by other experimenters, in attempting to reach a satisfactory solution of the quadruple interlacing problem in general. The system is shown diagrammatically in Fig. 1.

Colour Characteristics at the Receiver

In the television system under discussion the primaries at the receiver are determined by the colour filters, red, green and blue, and the fluorescent material in the tube. Unfortunately, monochromatic primaries can only be obtained at the sacrifice of light intensity. Thus one finds that in television as in certain colour-reproducing processes, a compromise must be found between light intensity and the best choice of primaries. In addition, there is a restricted choice in available phosphors. The decay time of the fluorescent powder used in the receiving tube must be such that its intensity becomes negligible after one colour-field period.

TABLE I.

System	3	4	5
Colour fields—(*c*)	120	180	120
Colour frames	40	60	40
Frames—(*f*)	60	45	30
Colour pictures	20	15	10
Interlace ratio—(*c*/*f*)	2 : 1	4 : 1	4 : 1
Lines per frame, corresponding to 441 black and white, nearest practical value	315	350	441
Horizontal (line) frequency	18,900	15,750	13,230
Lines per frame, corresponding to 525 black and white, nearest practical value	375	450	525
Horizontal (line) frequency	22,500	20,250	15,750
Colour breakup conditions	Satisfactory	Satisfactory	
Interline flicker conditions	,,	Doubtful	Unsatisfactory
Picture flicker conditions	,,	Satisfactory	Satisfactory

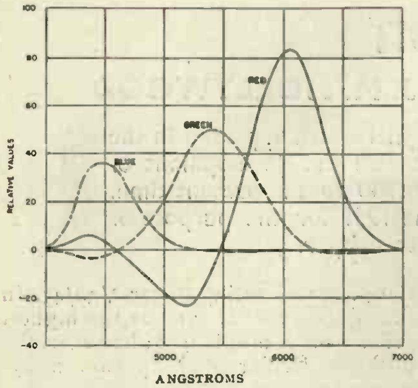


Fig. 3. Colour characteristics at the transmitter: calculated on the basis of receiver characteristics represented in Fig. 2.

Of the commercially available phosphors the zinc and calcium sulphides possess sufficient luminescent efficiency and also satisfy the decay requirements. The luminescent spectrum of the phosphor must cover the entire range of the three filters in order to provide a light source for each primary and the precise character of the spectrum desired is contingent upon the choice of filters.

The filters finally chosen for use at the receiver were Wratten No. 26 for red, No. 47 for blue, and No. 58 for green. The emission curves for the phosphor mixture used for the experimental tubes combined with filters Nos. 47, 58 and 26 are given in Fig. 2.

The range of colours obtainable with this choice of phosphors and filters is as large as is encountered in colour photography. The white produced with three equal signals appears somewhat bluish. Very recently, however, satisfactory "white" tubes were made in the laboratories, which show consistently good colour characteristics and permit transmitter operation with equal blanking pulses.

Transmitter Colour Characteristics

While the performance of the receiver was based on the colour theory of vision, the study of the colour characteristics at the transmitting end of the system had to be guided by the desirability of producing all colours encountered in nature. At the receiver three properly chosen narrow bands in the spectrum were sufficient; at the transmitter, the bands must be wide enough and sufficiently overlapping to produce a signal from every colour. The exact character of the three spectral curves at the transmitter is determined by the filter and phosphor combination at the receiver. The general relationship between the colour characteristics of the transmitter and the receiver can be derived from an analysis by Hardy and Wurzburg[1] in connexion with photographic reproduction in colour, and the resulting colour characteristics at the transmitter obtained from calculations are shown in Fig. 3. It will be noted that these curves show the characteristics usually found in colour matching problems—the existence of negative colour values for perfect matching. The present system has no mechanism for introducing negative values, but they are partially compensated for by the colour mixer.

Camera Tubes

Before comparing the results contained in Fig. 3 with actual operating conditions, it is best to consider briefly the colour characteristics of the tubes used at the transmitting end. They are of two types; the dissector, used for slides and motion pictures, and the orthicon, used in the studio and in outdoor pickups. One of the problems in dissector operation is the elimination of signals produced with infra-red radiation. The infra-red contaminates all colours as it passes freely through the red, blue and green filters. Originally the standard dissector was used with cæsiated silver-oxide cathode surface.

This surface has a minimum colour response in the green portion of the spectrum while it shows rising tendencies both towards the blue and the infra-red regions. In order to utilise this tube a carbon arc was used as a light source, combined with the infra-red absorbing Corning glass filter No. 978, 2 millimetres thick.

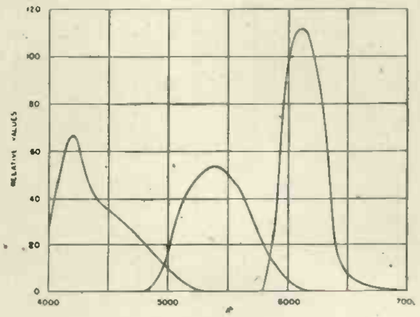


Fig. 4 (above). Colour characteristics at the transmitter: the curves combine the standard dissector, the carbon arc, 2 mm. of Corning glass No. 978 and Wratten filters Nos. 25, 47 and 58.

The filters that were used in this setup were Wratten Nos. 47, 58 and 25. Fig. 4 is a graphic representation of the results.

The dissector used at present is the so-called daylight dissector (developed especially for colour television by the Farnsworth Television and Radio Corporation) with a maximum in the green portion of the spectrum falling off towards the blue and red end.

This dissector was also used with a carbon arc, but with a Corning filter No. 978 only 1 millimetre thick. The colour filters were again Nos. 47, 58 and 25.

In both cases a water cell was used to secure protection for the slides and film. There is no question of the superiority of this type of tube over the standard dissector for this work. The signal-to-noise ratio is improved partly because of the greater photo-electric response in the pertinent portion of the spectrum and also because of the reduction in thickness of the infra-red filter.

Flicker in Colour Television

The well-known Ferry-Porter[2, 3] law states that the critical frequency is proportional to the logarithm of the illumination intensity.

In the sequential colour-television system under consideration the worst flicker condition would occur if only the primary colour with the highest luminosity were received while the other two primary colours were suppressed completely.

Such a condition hardly ever occurs in practice unless the colour camera picks up a green object, the limits of its chromaticity being between 5,400

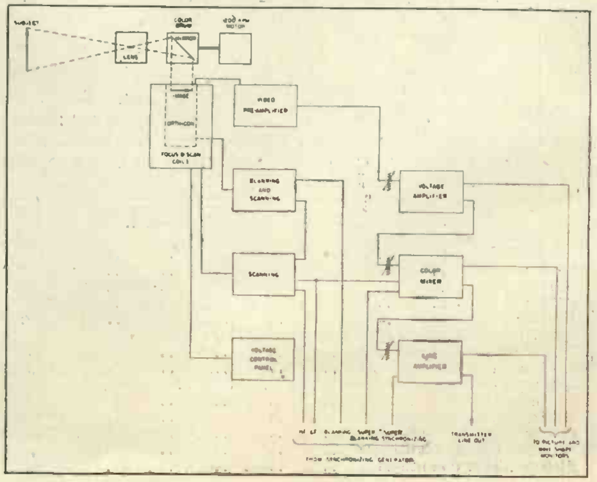


Fig. 5. Block diagram of orthicon pick-up camera.

and 5,800 angstrom units (*see* Fig. 4). The flicker frequency for this case would be 40 per second.

Applying Talbot's law[3] to the special case of colour television as discussed here, the apparent brilliance of the image at the receiver would be

$$I = \frac{1}{T_c}\int_0^T L(t)dt$$

where T_c is the duration of a complete colour cycle (colour frame) and T the duration of a colour field. $L(t)$ is the decay function of the screen material. This is assumed to be exponential, with a luminosity not greater than 1/10th of the initial brilliance after the duration of one field period (T).

Thus for the transmission of green between 5,400 and 5,800 angstrom units the apparent brilliance of the received picture becomes

$$I_g = \frac{1}{T_c}\int_0^T Y_g L(t)dt$$

where the luminosity at the receiver has been expressed in terms of Y_g a component representing the combination of the receiver's green filter and the screen material and the decay with time of the screen material.

In order to determine at what apparent brilliance a 40-cycle television picture will just begin to show flicker we consult the curves given by Engstrom[4] which were obtained by using a special film which corresponded to certain decay characteristics of screen materials. It was decided to choose film No. 1, the decay characteristic of which is sufficiently fast to correspond to a screen material usable in colour television, where at the end of one frame the screen brilliance decays practically to zero, which shows that a repetition frequency of 40 per second will permit a screen illumination of 1.8 apparent foot-candles.

So far we have considered the most unfavourable case from a flicker point of view. More favourable conditions will occur if white with three equal electrical impulses, during the red, green and blue periods is transmitted and received.

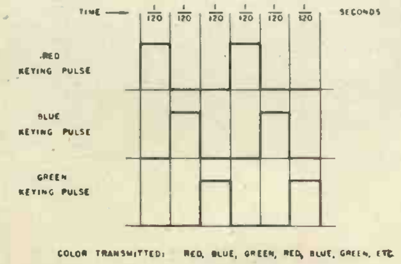


Fig. 8. Colour mixer pulse diagram.

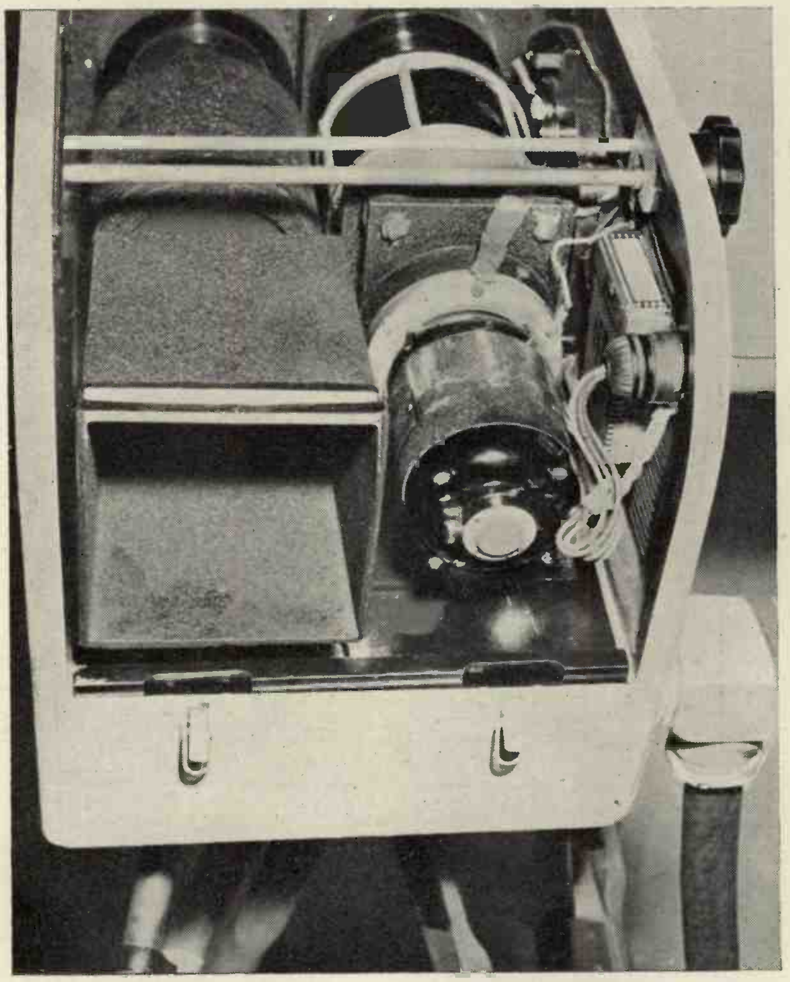


Fig. 6. Orthicon colour camera ; inside view showing the filter drum with synchronous driving motor.

The apparent brilliance of the receiver screen increases only in the ratio of 34.3/23 when white is transmitted, even though it is produced by three equal electrical impulses, one during each colour field. Since in this case a light impulse is received during each of the colour fields, flicker conditions improve rapidly. However, the apparent brilliance should not be higher than 1.8 times 34.3/23 which is 2.7 apparent foot-candles if one wishes to make sure that in the singular case of the transmission of a narrow band of green no flicker is present.

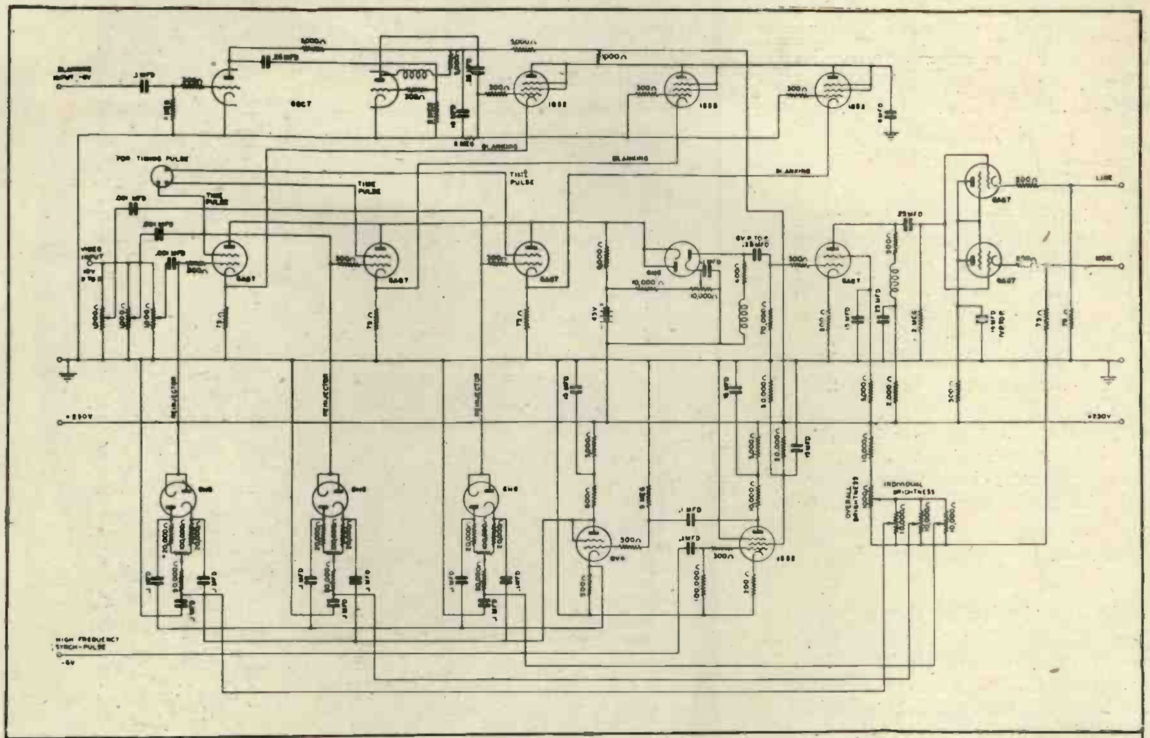
Colour-television pictures produced with the aid of rotating filters do not deteriorate appreciably in the surrounding illumination due to the fact that the room light which passes through the filters twice is attenuated by the square of the filter loss factor, while the picture itself is only attenuated by the first power of the filter factor.

Equipment for Colour Television

Certain electrical requirements must be met by pickup tubes if they are to be used in colour television. It is important that the signals produced during any one colour field should not be adulterated by a signal left over from a previous field. Storage-type camera tubes must, therefore, be designed so that the entire electrical charge on the mosaic is removed within one field period.

A constant black level must be established in the camera tube, and spurious signals such as " shading " should be absent. The dissector is the only commercial camera tube that meets all of the above requirements though its usefulness is limited to the transmission of film or slides due to its low sensitivity.

The orthicon as modified for colour television with lower mosaic capacitance was developed through the cooperation of the R.C.A. Radiotron Division and has been found to pro-



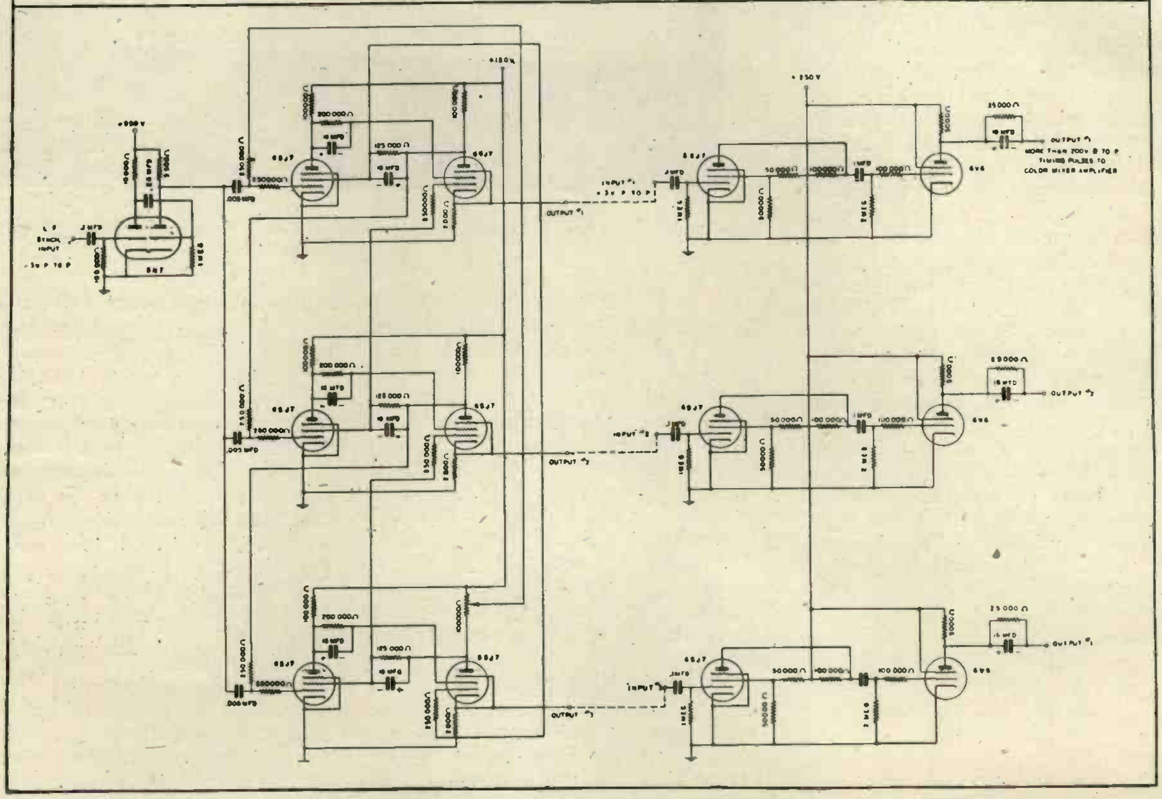


Fig. 7 (top). Circuit diagram of colour mixing amplifier.

Fig. 9. Circuit diagram of colour mixing pulse generator.

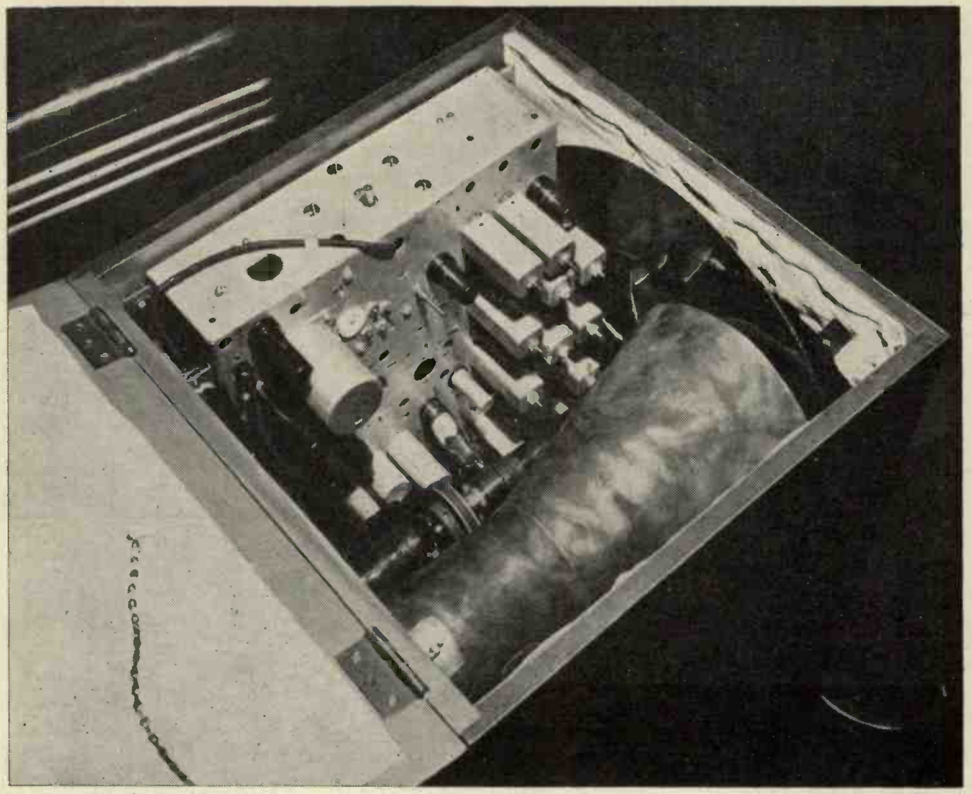


Fig. 11. The interior view of a 9″ colour television receiver showing filter disk and cathode ray tube (lower right).

duce very acceptable colour pictures with incident light of 150 foot-candles on the subject. A certain amount of "hangover" which may be defined as the amount of signal remaining on the mosaic after the scanning beam has completed one field, appears to be unavoidable, but is only troublesome at lower light levels.

The gamma of a camera tube need not necessarily be unity, as correction may be made for any particular characteristic later on, if desired. In general, a television system employing a linear pickup tube such as an orthicon, will have an overall gamma higher than unity, due to the cathode-ray tube. A reduction in the gamma may be more satisfactorily made with tubes of the dissector type, where the noise is negligible in the black portions of the picture, than in a tube of the orthicon type, where the noise is determined by the impedance of the tube and the first amplifier stage.

The introduction of colour does not change many of the design requirements which are ordinarily encountered in monochromatic television studio equiqment. Certain factors, however, are worthy of mention. The colour field frequency of 120 per second necessitates freedom from 60-cycle hum in the synchronising generator and scanning equipment and, to a lesser legree, in video amplifiers. Sixty cycle components present in the synchronising generator or scanning equipment cause loss of interlace and in the video equipment cause flicker at a 20-cycle rate resulting from the beat between the 60-cycle hum and the 40-cycle picture components. Hum may be eliminated by operating the equipment from a 120-cycle power source.

Good low-frequency response is necessary in video amplifiers to pass the 40-cycle picture components properly. The video control equipment for colour is somewhat more complex than for black-and-white transmission, as it seems advantageous to control the gain and possibly the background of each colour independently, as previously mentioned.

A block diagram of a colour-television system using an orthicon camera tube is shown in Fig. 5 and a view of the camera is shown in Fig. 6.

As previously mentioned, it is essential that the black level be established at the camera since manual control of the direct-current level for each colour would seem a tremendous task. This is done by applying the blanking pulses to the grid of the orthicon and to the cathode of the dissector.

The video amplifiers in a colour-television camera channel are conventional except for the colour mixer. Manual control of gain and brightness for each colour are achieved by the equipment shown in Fig. 7. The colour-mixer amplifier may be described as an electronic switch combined with three separate amplifiers, each with its own gain and brightness controls. The video signal is switched by means of suitable timing pulses (Fig. 8) applied to the screen grids of the 6AG7 switching amplifier tubes. The pulses are so timed as to operate each amplifier in succession, turning on one as another is turned off. This switching occurs during the blanking period, and switching transients are removed by subsequent clipping of the recombined sigals. Blanking is injected on the cathode of each switching amplifier tube and the individual brightness of each colour is adjustable by bias controls. The switching pulses are generated by the "ring frequency divider" circuit shown in Fig. 9.

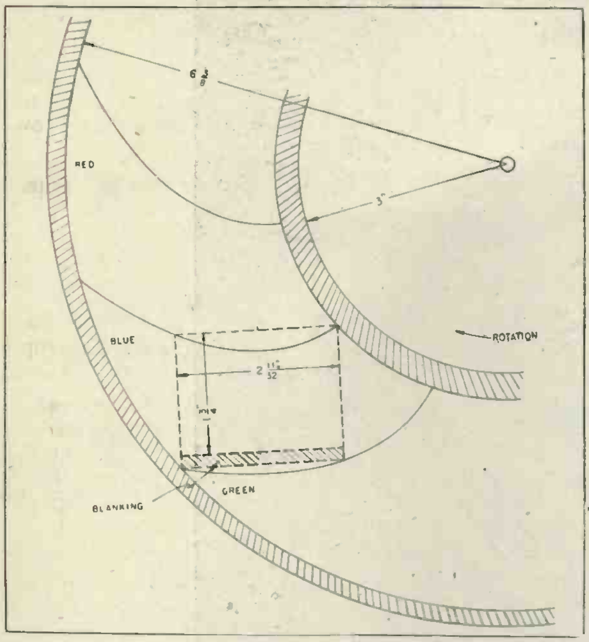


Fig. 10. Drawing showing portion of filter disk for orthicon camera.

Filter Disks

The non-storage dissector is more easily adapted to a filter disk, since it is necessary that the optical image on the cathode be of the correct colour only at the point which is being scanned. The orthicon, on the other hand, being of the storage type, requires that only one colour can be present in the optical image for one complete scanning-field period, prior to the actual scansion of a given point.

Fig. 10 shows a filter disk designed for the orthicon, where the contamination of colours due to the curvature of the filter segments is not more than 10 per cent. A filter drum accomplishes the same purpose with less contamination and less space. The drum is phased so that the shadow of the

(*Concluded on page 213*)

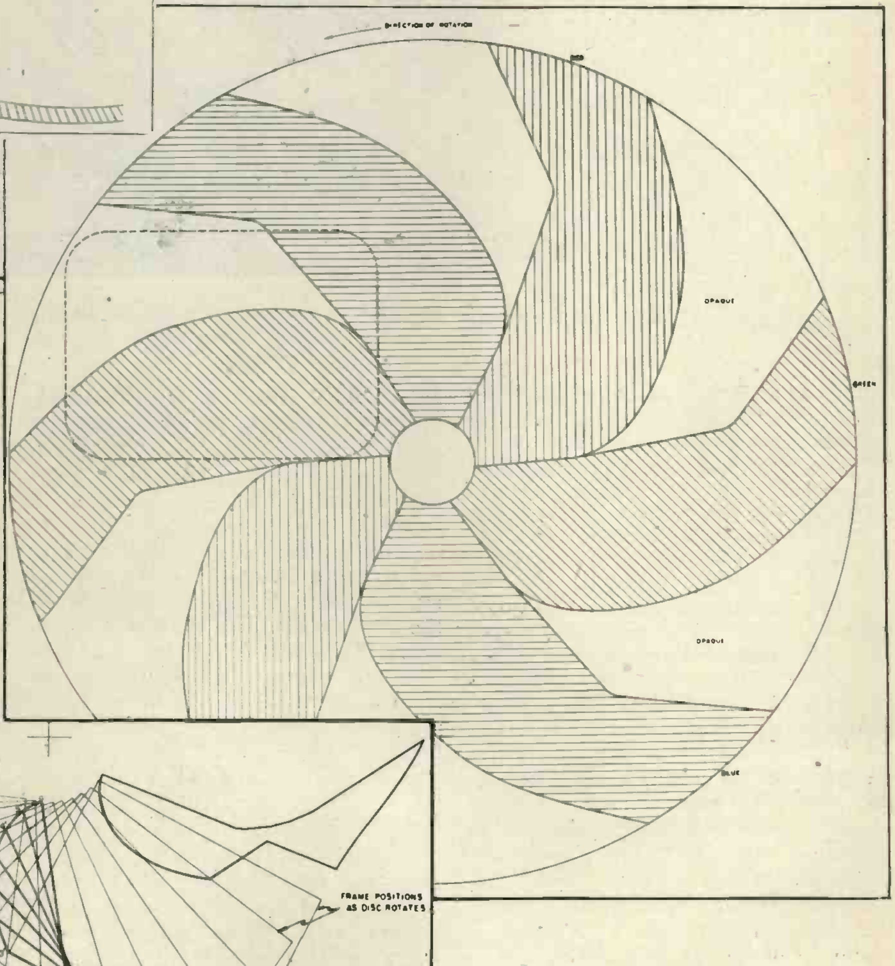


Fig. 13 (right). Typical receiver filter disk design.

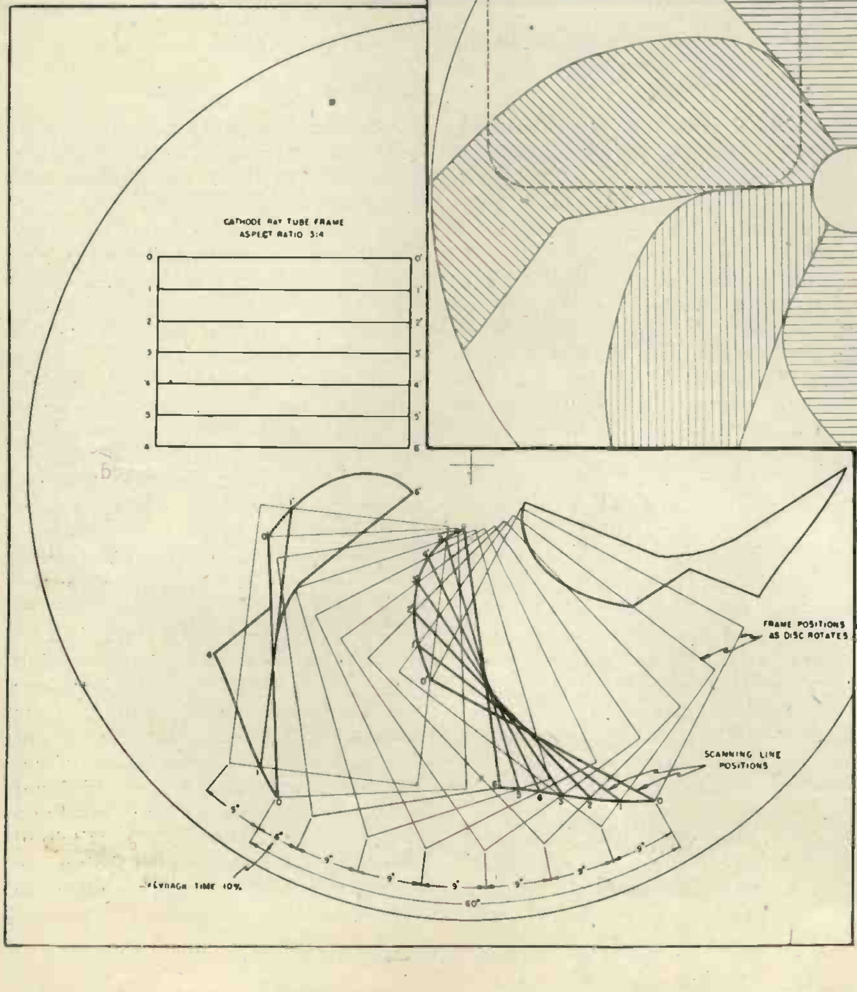


Fig. 12 (left). Drawing showing the generation of proper shape of filter disk.

DATA SHEET XXXVII.

Choke Input Filter for Bi-Phase Rectifier Circuit*

THIS Data Sheet deals with the effect of a choke of finite inductance in a bi-phase rectifier circuit.

The object of the use of a choke input filter is (a) to obtain good regulation of output voltage for varying load currents, and (b) to reduce the ratio of peak-to-mean rectifier current to a lower value than is possible with a condenser input filter.

In order to obtain the best regulation it is essential to ensure that the current in the choke never drops to zero.

The circuit on Data Sheet 37 shows a bi-phase rectifier with a choke input filter circuit L_1C_1. For simplicity, assume that a further inductance L_2 is interposed between the load R and the choke L_1. This inductance L_2 is assumed to have a sufficiently high value to ensure that the relation.

$$4\pi f L_2 \geqslant \frac{1}{4\pi f C_1} \quad \ldots \quad (1)$$

is satisfied (where f is the frequency of the mains supply).

Fig. 1 illustrates the voltage across the points marked A, B in the circuit diagram. This waveform consists of a D.C. component V_d plus a number of harmonics, the analysis of which is given by:

$$v = V\,(2/\pi + 4/3\pi \cos 2\omega t - 4/15\pi \cos 4\omega t + \text{etc.} \quad \ldots \quad (2)$$

where V is the peak value of voltage.

The current through L_1 for a constant voltage amplitude will be inversely proportional to the frequency provided that the circuit is not near resonance. The current through L_1 due to the fundamental ripple frequency $2f$ will therefore be 10 times larger than that due to the second harmonic $4f$ and 35 times larger than that due to the third harmonic.

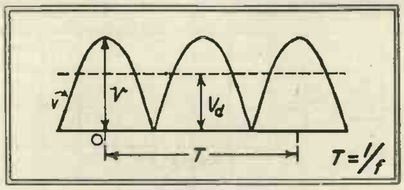
We can ensure that the current in the choke never drops to zero by making the current due to the fundamental ripple voltage always less than the D.C. load current I_d

This may be expressed (provided Eq. (1) is satisfied) by the relation:

$$I_d = \frac{2V}{\pi R} \geqslant \frac{4V}{3\pi Z} \quad \ldots \quad (3)$$

$$\text{or} \qquad Z/R \geqslant 2/3 \quad \ldots \quad (4)$$

where Z is the impedance of the circuit L_1C_1 at the frequency $2f$, and



$$\frac{Z}{R} = \frac{2\omega L_1}{R} - \frac{1}{2\omega C_1 R} = 2a - \frac{1}{2b} \quad \text{say} \quad (5)$$

$$\text{from which} \qquad a \geqslant \frac{4b+3}{12b} \quad \ldots \quad (6)$$

Or, if we express L_1 in henries, C_1 in μF and R in kilo-ohms, then for 50 c/s mains:

$$\frac{L_1}{R} \geqslant \frac{4C_1R + 9.55}{3.76C_1R}$$

henries per kilo-ohm (7)

This relation is shown by the full line curve on Data Sheet 37 and from this it will be seen that the highest value of inductance is required when drawing the lowest current. If the load resistance is gradually increased beyond the value which satisfies Eq. (5) and (6) then the output voltage V_d will rise above the D.C. value of $0.637\,V = 0.9V_s$, the circuit finally operating as a condenser input filter at very low currents.

When the inductance L_1 is reduced to the lowest value satisfying (5) or (6) the peak current through L_1 is equal to twice the mean load current I_d and therefore each rectifier is operating with a peak-to-mean current ratio of 4 : 1.

If it is desired to reduce the peak-to-mean ratio below this value a higher value of inductance must be chosen for L_1.

Denoting the desired peak-to-mean ratio by m, the expression (3) becomes

$$I_d = \frac{2V}{\pi R} \geqslant \frac{2}{m-2}\left(\frac{4V}{3\pi Z}\right) \quad (8)$$

and (7) becomes:

$$\frac{L_1}{R} \geqslant \frac{8C_1R + 9.55(m-2)}{3.76C_1R(m-2)}$$

henries per kilo-ohm (9)

Expression (9) has been plotted on Data Sheet 37 for $m = 3.5$ and $m = 3$. As the value of $m = 4$ represents the limit of good regulation, it cannot be used in practice for design purposes (except as an extreme) due to the inevitable tolerance on the value of L_1C_1 and R. The value of $m = 3.5$, however, provides a good general basis for design, giving a reasonably low value of inductance for L_1.

When the condenser size is limited the figures

$$C_1R = 9\ \mu\text{F} \times \text{kilo-ohms} \quad (10)$$

and

$$L_1/R = 1.7 \text{ henries per kilo-ohm} \quad (11)$$

will provide a reasonable compromise for general design with a mains supply of 50 c/s.

To use the curves for any other mains supply frequency f the ordinates of the C_1R and L_1/R scales should be multiplied by the ratio $50/f$. Thus with 60 c/s mains the above figures become 7.5 μF × kilo-ohms and 1.4 henries per kilo-ohm.

In order to prevent instability and uneven loading of the rectifiers, Dunham* recommends that the resonance frequency of the filter circuit should not be within less than 10 per cent. of the frequency of the supply mains. This restriction is shown by the boundary $\omega^2L_1C_1 = 0.81$, and beyond this boundary (VII) the curves I, III and V are shown dotted.

In the range of the curves shown, the filter section L_1C_1 (with relation (1) satisfied) provides a smoothing action. If we denote the peak amplitude of the fundamental ripple frequency ($2f$) by V_r, then from (2) at the points A, B the ratio $V_r/V_d = 2/3$. The peak amplitude of the fundamental ripple voltage V_r after passing through the filter L_1C_1 (points D.F. on the circuit) is shown on the Data Sheet by the curves II, IV and VI as a ratio V_r/V_d.

The case where the second section of the filter is omitted, *i.e.*, $L_2 = 0$ and $C_2 = 0$, considerably complicates the equations unless $\omega C_1 R \gg 1$, when the relations are unchanged. Thus, equation (10) becomes:

$$a \geqslant \frac{b}{1+4b^2} \pm \sqrt{\frac{4}{9(m-2)^2} - \frac{1}{4(1+4b^2)^2}} \quad (12)$$

which is shown on the Data Sheet 37 by broken lines.

The load resonance boundary is also shifted downwards. For further information, reference should be made to Dunham's article (*see* previous footnote).

* This circuit is also called a Single-Phase Full-Wave Circuit.

* "Some considerations in the design of Hot-Cathode Mercury Vapour Rectifier Circuits," C. R. Dunham, *Jour. I.E.E.*, Sept. 1934.

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DESIGN OF CHOKE INPUT FILTER FOR BI-PHASE RECTIFIER CIRCUITS – 50 c/s.

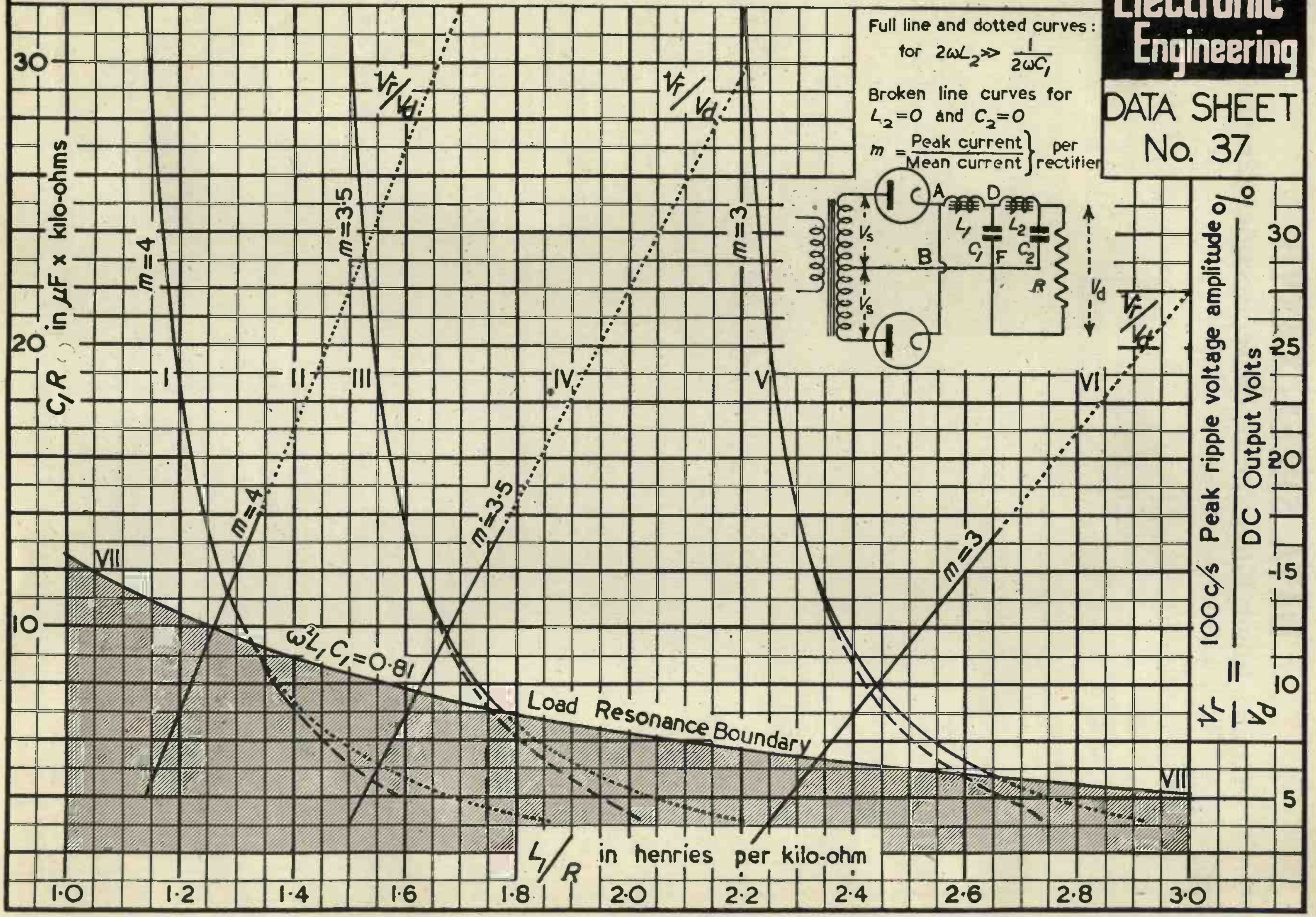
Electronic Engineering

DATA SHEET No. 37

Full line and dotted curves: for $2\omega L_2 \gg \frac{1}{2\omega C_1}$

Broken line curves for $L_2 = 0$ and $C_2 = 0$

$m = \frac{\text{Peak current}}{\text{Mean current}}$ per rectifier



DATA SHEET XXXVIII

Condenser Input Filter Power Supplies

THE generalised graphical representation of the performance of rectifiers working into a condenser input filter is rather an intractable problem. For certain conditions, simplifying assumptions can be made which enable a reasonably simple presentation to be produced. One such case is that of half-wave high voltage low power supplies (Fig. 1a and 1b) which are extensively used in both television and laboratory equipments to provide the H.T. volts for cathode ray tubes. These power supplies usually provide outputs of the order of $\frac{1}{4}$ to 1 mA at a voltage of some 2,000 to 6,000 volts. The low load currents justify the simplifying assumption of a zero rectifier volt drop and the output voltage can then be assumed to have a wave form as illustrated in Fig. 1c and 1d.

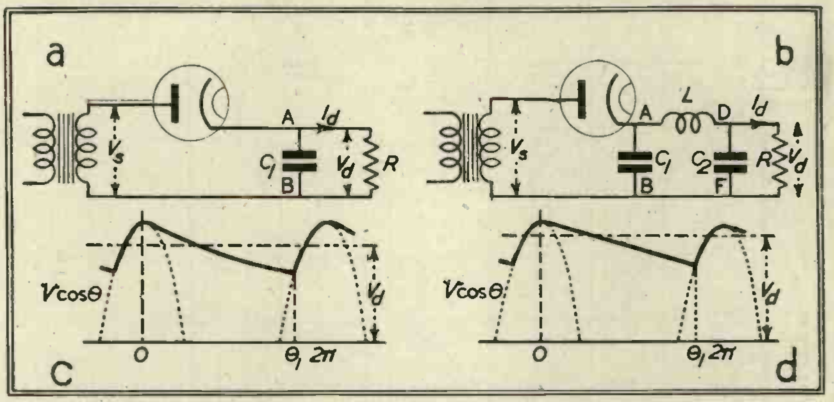


Fig. 1

Fig. 1a represents the most popular circuit which is used in the form illustrated or with an additional resistance-capacity smoothing section according to the magnitude of the load current I_d employed. Fig. 1b illustrates a less frequently used circuit more suitable for higher load currents where the additional smoothing section consists of an inductance L_1 and condenser C_2.

With the assumption of a perfect transformer choke and rectifier, the voltage across the condenser C_1 follows the thick line curves of Fig. 1c and 1d. In the case of Fig. 1a the voltage across C_1 rises to the peak amplitude V of the applied voltage and then discharges exponentially, while in the case of Fig. 1b the discharge is linear if the inductance L_1 is of a sufficiently high value.

Let the transformer secondary voltage be E_s volts r.m.s. the instantaneous voltage v is then

$$v = V\cos\theta \qquad \ldots \quad (1)$$

where V is the peak amplitude and equal to $\sqrt{2}E_s$ and $\theta = (2\pi f)t = \omega t$.

For the circuit Fig. 1a the voltage across the condenser C_1 after reaching V for $\theta = 0$ discharges with an instantaneous value

$$v_d = V\exp\left(\frac{-t}{C_1R}\right) = V\exp\left(\frac{-\theta}{\omega C_1R}\right) \qquad (2)$$

The condenser goes on discharging until $V_d = V\cos\theta$. Let the angle θ which satisfies the above relation be θ_1 then

$$\exp\left(\frac{-\theta_1}{\omega C_1R}\right) = \cos\theta_1 \qquad (3)$$

Also during the discharge period the average voltage is equal to

$$\frac{1}{\theta_1}\int_0^{\theta_1} V\exp\left(\frac{-\theta_1}{\omega C_1R}\right)d\theta \qquad (4)$$

$$= \frac{V\omega C_1R}{\theta_1}\left[1-\exp\left(\frac{-\theta_1}{\omega C_1R}\right)\right] \qquad (5)$$

or applying (3)

$$= \frac{V\omega C_1R}{\theta_1}(1-\cos\theta_1) \qquad (6)$$

During the charging period, *i.e.*, from $\theta = \theta_1$ to $\theta = 2\pi$ the average voltage across C_1 is

$$= \frac{1}{2\pi-\theta_1}\int_{\theta_1}^{2\pi} V\cos\theta\,d\theta \qquad (7)$$

$$= \frac{-V\sin\theta_1}{2\pi-\theta_1} \qquad \ldots \quad (8)$$

and the average voltage over the complete cycle is

$$V_d = \frac{V}{2\pi}\left\{\omega C_1R(1-\cos\theta_1)-\sin\theta_1\right\} \qquad (9)$$

If we designate by the symbol "η" the voltage rectification efficiency: that is the ratio of V_d/V then

$$\eta = f\,C_1R(1-\cos\theta_1)-\frac{\sin\theta_1}{2\pi} \qquad (10)$$

By the use of equations (3) and (10) it is possible to plot η in terms of ωC_1R or C_1R and the result is given by the dotted curve on Data Sheet No. 38 for a supply frequency of 50 c/s.

In a similar manner we have for circuit Fig. 1b that the instantaneous volts during the discharge period are given by

$$v_d = V - \frac{I_d}{C_1}\cdot\frac{\theta}{\omega} \qquad \ldots \quad (11)$$

and equation (3) is therefore replaced by

$$\frac{C_1V}{I_d} = \frac{\theta_1}{\omega(1-\cos\theta_1)}$$

or

$$\omega C_1R = \eta\frac{\theta_1}{1-\cos\theta_1} \qquad (12)$$

also the average volts during the charging period are still given by (8), while during the discharge period

$$v_d = \frac{V(1+\cos\theta_1)}{2} \qquad \ldots \quad (13)$$

and therefore η in (12) is given by

$$\eta = \frac{1}{2\pi}\left\{\frac{\theta_1(1+\cos\theta_1)}{2} - \sin\theta_1\right\} \qquad (14)$$

From equations (12) and (14) the full line curve for $\eta = V_d/V$ has been drawn on Data Sheet No. 38.

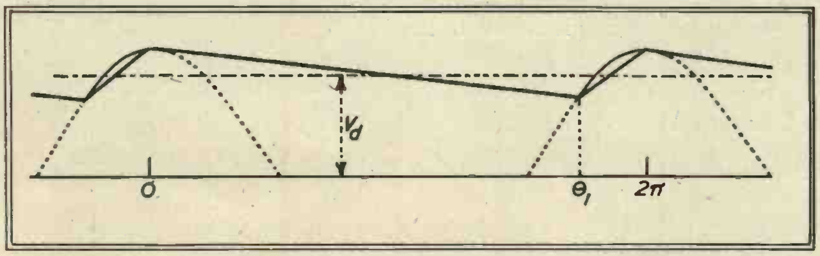
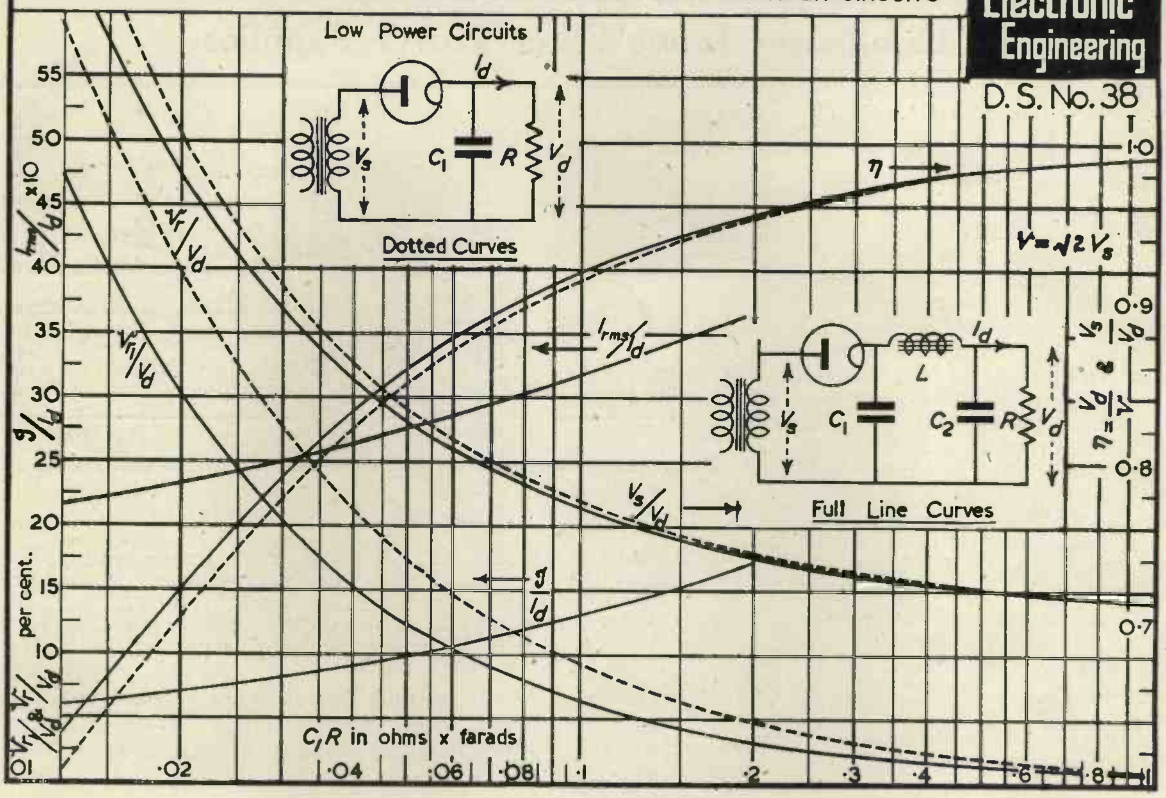


Fig. 2

THE PERFORMANCE OF HALF-WAVE CONDENSER INPUT FILTER RECTIFIER CIRCUITS



Ripple Voltage

If the voltage across the condenser C_1 of circuit iFg. 1a is applied to a C.R. tube then the ripple will affect the instantaneous sensitivity of the tube and probably the focus. The magnitude of the effects will be proportional to the ratio of the peak-to-peak amplitude of the ripple (*i.e.*, one-half the peak-to-peak amplitude) and is given by

$$V_r = \frac{V(1 - \cos \theta_1)}{2} \qquad (15)$$

and expressed as a percentage of V_d

$$\frac{V_r}{V_d} = \frac{100\pi(1 - \cos \theta_1)}{\omega CR(1-\cos \theta_1) - \sin \theta_1} \% \qquad (16)$$

This expression is shown by the dotted curve on the Data Sheet. For large values of $\omega C_1 R$, say of 50 and over, expression (16) may be further simplified to:

$$\frac{V_r}{V_d} \approx \frac{90}{2fC_1R} \% \quad \ldots \quad (17)$$

which for 50 c/s mains gives

$$\frac{V_r}{V_d} \approx \frac{0.9}{C_1R} \% \quad \ldots \quad (18)$$

For some applications the actual value of V_r is required and this may be expressed from (17) by

$$V_r \approx \frac{0.9\, I_d}{2fC_1}$$

or at 50 c/s

$$\approx 9 \text{ volts per mA per } \mu\text{F} \quad \ldots \quad (19)$$

When, however, a further smoothing circuit follows the reservoir condenser C_1 such as in Fig. 1b (or if an additional resistance capacity filter stage is added to Fig. 1a) then we must consider the peak amplitude of the fundamental component of the ripple voltage across C_1 in order to be able to calculate the final ripple voltage across C_2 (see Data Sheets Nos. 35 and 36). In order to estimate the fundamental component it is necessary to approximate the waveform shown in Fig. 1d by the wave shown in Fig. 2.

The analysis of this triangular waveform has been given in Data Sheet No. 29 and by its use the full line curve of V_{r1}/V_d was calculated where V_{r1} is the amplitude of the fundamental ripple across the condenser C_1 of circuit Fig. 1b. (The second harmonic has an amplitude of approximately one half the fundamental).

In a similar manner the fundamental component of the ripple voltage of circuit Fig. 1a may be estimated, but an approximate figure will be obtained by multiplying the ordinates of the dotted V_r/V_d curve by 2/3, which is given near enough by the V_{r1}/V_d curve for C_1R values above 0.15.

Equation (19) is then replaced by (20) where V_{r1} is the peak amplitude of the fundamental (50 c/s) ripple voltage.

$$V_{r1} \approx 6 \text{ volts per mA per } \mu\text{F} \quad (20)$$

if the r.m.s. value is required the figure 6 should be replaced by $4\frac{1}{4}$.

While the above approximations are reasonably satisfactory for voltage relations they are not really suitable for current relations as the peak anode current $\hat{I}$ is so greatly affected by remaining reactances and resistances. Approximate curves have, however, been plotted on the Data Sheet, of the ratios $\frac{I_{R \cdot M \cdot S \cdot}{}^*}{I_d}$ and $\frac{\hat{I}}{I_d}$ to give an indication of their magnitude, and as such the curves may be used for both circuits.

While all the curves have been drawn for a supply frequency of 50 c/s they may be used for any other frequency f by multiplying the C_1R scale by the ratio $50/f$.

* $I_{R \cdot M \cdot S \cdot}$ = R.M.S. Anode current.

Designing a Resistance-Loaded Push-Pull Invertor

By R. FEINBERG, Dr.Ing:, M.Sc:.*

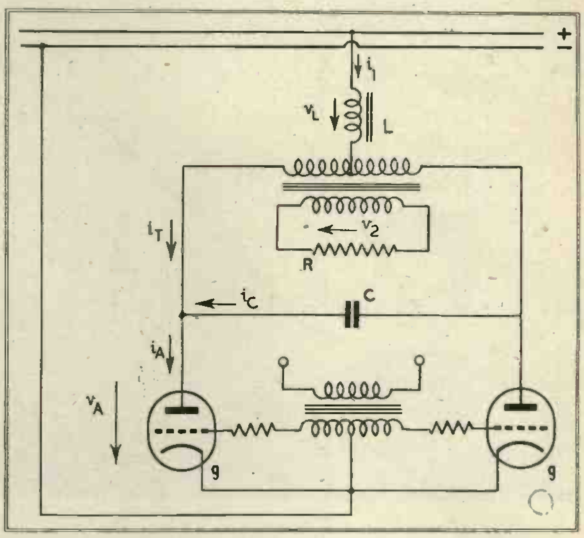


Fig. 1. Circuit diagram of the invertor circuit.

THE push-pull invertor with gas-filled triodes whose operation and mathematical theory have been described elsewhere† is a thermionic valve circuit for static conversion of direct current to alternating current. Current conversion is obtained at a high degree of efficiency owing to the relatively small power loss in the gas-filled valves.

Fig. 1 shows the circuit diagram of the resistance-loaded invertor, the records of Fig. 2 represent some typical forms of alternating current, the records of Fig 3 show the direct current supplied to an invertor under various load conditions, and Fig. 4 indicates voltages and currents in an invertor circuit.

Designing an invertor means calculating the values of the circuit parameters, viz., inductance L, capacitance C, and transformer ratio m, *i.e.*, turns ration of secondary to primary. (*See* Fig. 1). The basic data in the case of a resistance-loaded invertor are power output P, power loss in inductor and transformer $P' = a.P$, alternating voltage V_2, frequency f, D.C. supply voltage E_1, and discharge-working voltage V_d of a gas-filled triode. In addition the values of the characteristic circuit factors

$$d = \frac{m^2}{4CRf} \quad \dots \quad (1)$$

and

$$k = \frac{R^2 C}{m^4 L} \quad \dots \quad (2)$$

have to be decided upon, d and k determining the waveform of alternating current and direct current. A rectangular form of alternating current as in Fig. 2(a) is obtained with $d > 5$, a triangular form as in Fig. 2(b) with $d < 0.5$ and $k < 1 + (\pi/2d)^2$, and a sinusoidal current form as in Fig. 2(c) with $d < 1$ and $k = 1 + (\pi/d)^2$. Smooth direct current as in Fig. 3(a) requires $k < 1$. The slightly rippled curve of Fig. 3(b) is the result of $k > 1$, and the heavily rippled curve of Fig. 3(c) is obtained with $k = 1 + (\pi/d)^2$, where $d < 1$.

Before proceeding to calculate the circuit parameters the value of d has to be checked for stable invertor operation. The diagram of Fig. 5 gives a critical time value t_v as a function of d and frequency f. The condition for stable invertor operation requires that t_v must not be smaller than the grid-control recovery time plus safety margin which is for small valves of the order of 0.1 to 2 milliseconds.

The values of the circuit parameters C, L, and m are found from the formulæ

$$C = \frac{P}{16\, d f E_1'^2 I_{10}'} \quad \dots \quad (3)$$

$$L = \frac{E_1'^2 I_{10}'}{k d f P} \quad \dots \quad (4)$$

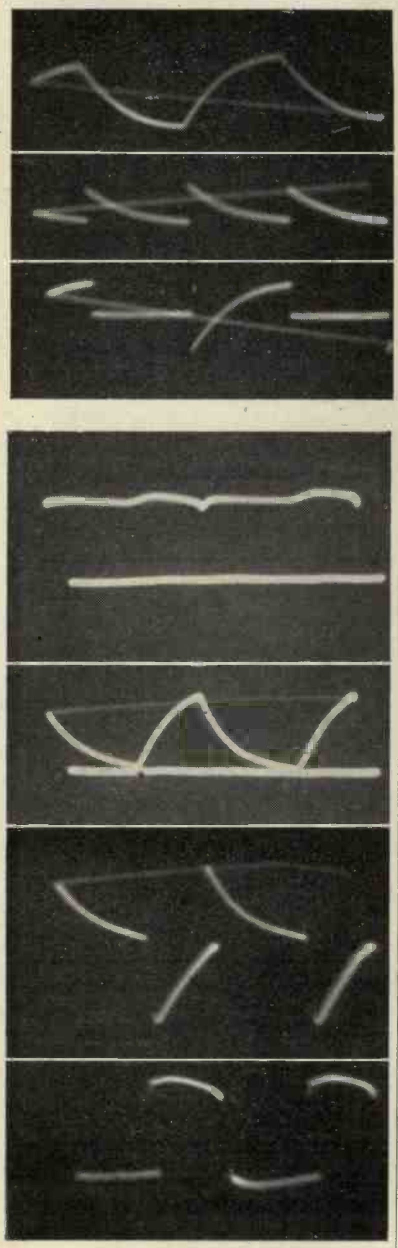


Fig. 4. Oscillograms of the voltages and currents at various points in the circuit. The lettering against each oscillogram corresponds to the lettering on the circuit diagram.

* Electrotechnics Department, University of Manchester.

† "Static Conversion of Direct Current to Alternating Current with Grid-Controlled Mercury-Arc Mutators." *Journal I.E.E.*

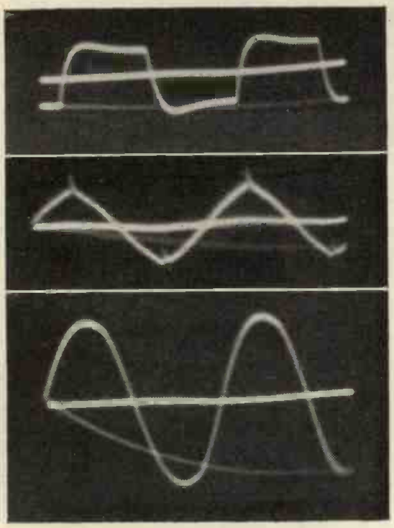
a
b
c


Fig. 2.

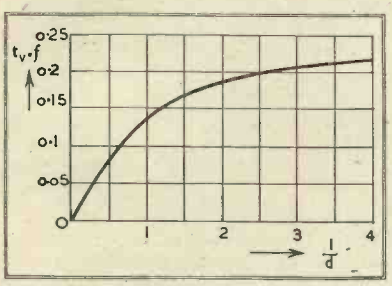
$$m = \frac{V_2}{2E'_1\sqrt{I'_{10}}} \quad \ldots \quad (5)$$

where

$$E'_1 = \frac{E_1 - V_d}{1 + a} \quad \ldots \quad (6)$$

and, if $k < 1$,

$$I'_{10} = 1 + \frac{n}{d}\,\frac{\cosh d - \cosh n}{n \sinh d - d \sinh n} \quad (7)$$



with

$$n = d\sqrt{(1 - k)} \quad (8)$$

if $k = 1$,

$$I'_{10} = 1 + \frac{1}{d}\,\frac{\cosh d - 1}{\sinh d - d} \quad \ldots \quad (9)$$

if $k > 1$

$$I'_{10} = 1 + \frac{\nu}{d}\,\frac{\cosh d - \cos\nu}{\nu \sinh d - d \sin\nu} \quad (10)$$

with

$$\nu = d\sqrt{(k - 1)} \quad \ldots \quad (11)$$

It should be noted that the formulæ for C, L, and m are derived with the assumption that the iron magnetisation of transformer and inductor are below saturation, and that leakage of flux in the transformer is fairly small.

Example

Basic data:

$P = 5$W, $P' = 0.5$W ($\therefore a = 0.1$)
$E_1 = 100$v, $V_d = 33$v ($\therefore E'_1 = 67/(1.1) = 61$v.)
$V_2 = 500$v, $f = 50$ c/s.

Current waveform:

Alternating current to be sinusoidal, therefore $d = 1$ and $k = (1 + \pi^2) = 10.9$.

Check for stability of invertor operation:

Grid-control recovery time: $t_g = 1$ millisec.
Safety margin: $t_s = 0.5$ millisec.
Critical time from Fig. 5: $t_v = 0.14/50$ sec. $= 2.8$ millisec.
Checked: $t_v > t_g + t_s$, *i.e.*, condition for stable invertor operation is satisfied.

Calculated:

$I'_{10} = 3.17$.
$C = 0.53$ μF.
$L = 4.3$ H.
$m = 2.3$.

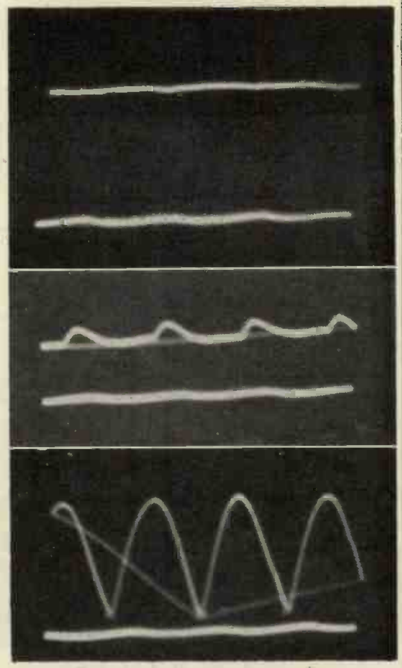
a
b
c


Fig. 3.

SALVAGE

Catalogues, Instruction Sheets, and Circuit Diagrams which are collected and filed for reference, mount up to a surprisingly large quantity in a comparatively short space of time.

There are probably catalogues in your files which are now out of date together with obsolete circuit diagrams. These would play a vital part in the war effort as paper salvage helps to make munitions.

Will you help by sorting your files at the earliest opportunity and add all you can to the salvage sack ?

Mathematical Calibration of Tuning Dials

IN *Communications* for April, 1942, R. L. Drake and R. R. Schmidt describe a method for the mathematical calibration of superhet tuning dials which they state has proved satisfactory in practice on automatic D.F. receivers.

It is pointed out that the dial calibration arrived at by this method is a true average of the tuning curves that are obtained from production variations in ganged condensers, and that the mathematical design of the dial enables it to be put in hand as soon as production is started without waiting for the completion of various radio and electrical design details.

The following data are required:

(a) Frequency range to be covered.
(b) Incremental capacity of tuning condenser.
(c) Degrees of rotation of pointer.
(d) Ratio of mid-band to i.f. frequency.

The frequency ratio of each band must be calculated and if possible a ratio should be selected that will cover all bands, thus eliminating trimmers and simplifying calculation.

The capacity ratio (C_{max} to C_{min}) is equal to the square of the frequency ratio F_r, and C_{max} and C_{min} are given by computing the actual values of capacity in the tuning condenser plus the stray capacities.

C_{max} is then equal to $\Delta_c$max. + C_{min}, where $\Delta_c$max. is the incremental change in capacity.

$$\Delta_c\text{max.} = C_{min}\,(F_r^2 - 1) \text{ or}$$

$$C_{min} = \frac{\Delta_c\text{max.}}{(F_r - 1)}$$

The incremental change in capacity is found for every 5 per cent. of rotation of the condenser spindle and the value is added to C_{min}.

The capacity ratio C_r at a given point on the dial is then

$$\frac{C_{min} + \Delta_c}{C_{min}}$$

and since $\log F_r = \frac{1}{2}\log C_r$, we have:
$\log F_r = \frac{1}{2}\log(C_{min} + \Delta_c) - \log C_{min}$

Knowing the upper limit of each band, the frequency for any point on the dial will be $F_k = F_{max}/F_r$, F_r being the frequency ratio for that particular per cent. of rotation of the condenser.

The frequency for each 5 per cent. rotation of the condenser can thus be determined and a curve plotted for per cent. rotation v. frequency for each band.

Practical Notes on Receiver Design

Part IV—(Conclusion)

By G. T. CLACK

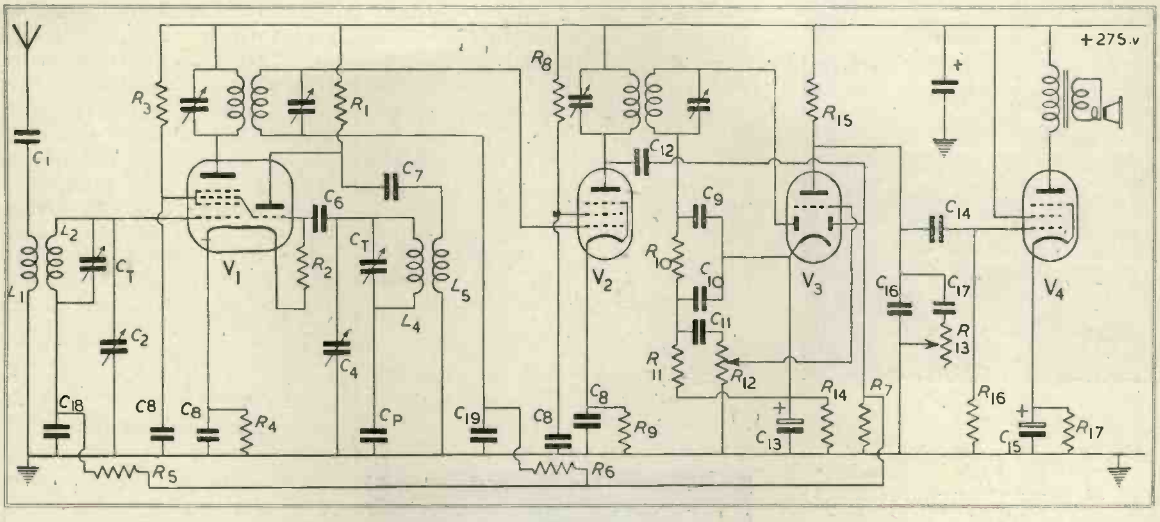


Fig. 34.

R_1 40,000 ohms	R_6 1.0 megohm	R_{11} 0.5 megohm	R_{16} 0.5 megohm	C_4 500 pF	C_9 100 pF	C_{14} 0.01 μF
R_2 40,000 ohms	R_7 1.0 megohm	R_{12} 0.5 megohm	R_{17} 200 ohms	C_P Padder	C_{10} 100 pF	C_{15} 50 μF
R_3 100,000 ohms	R_8 100,000 ohms	R_{13} 0.5 megohm	C_1 50 pF	C_6 50 pF	C_{11} 0.01 μF	C_{16} 0.001 μF
R_4 100 ohms	R_9 300 ohms	R_{14} 1,000 ohms	C_2 500 pF	C_7 500 pF	C_{12} 50 pF	C_{17} 0.02 μF
R_5 1.0 megohm	R_{10} 0.25 megohm	R_{15} 50,000 ohms	C_T Trimmers	C_8 0.1 μF	C_{13} 25 μF	C_{18} 0.1 μF

A.F. Stages

THE last section of the receiver under discussion consists of an A.F. amplifier and output valve as illustrated in Fig. 34. Both sensitivity and response requirements depend upon the character of the signal across R_{12} which forms part of the diode circuit discussed earlier. The output power depends upon individual requirements and determines the type of valve capable of supplying the requisite wattage to the loudspeaker. With a valve of the PenA4 or EL33 class operating from an H.T. line of 270 volts an output of 2½ to 3 watts is obtained which is quite sufficient for average requirements.

From a knowledge of the input voltage to the first stage and output wattage from the last stage, the number and type of A.F. stages can be determined. In general only two stages are necessary and a typical circuit is shown, where the triode section of V_3 functions as a voltage amplifier and the pentode V_4 as the output power valve.

Output Transformers

These are essentially power transformers used at alternating flux densities much lower than the usual mains transformers, and provide the most convenient method of correctly matching the valve to its external load. The simple formula to calculate the ratio of primary to secondary impedance is N^2 = Optimum Load/Speaker Impedance which does not take into account any transformer losses and is only true for an ideal transformer. The windings have resistance, and as each turn cannot occupy the same position in space there is a small magnetic leakage between primary and secondary windings. There is also capacity between sections of each winding and between primary and secondary, together with eddy current losses in the core. The latter two faults are not serious in output transformers, but call for considerable attention in intervalve transformer design. The resistance present in both windings together with the core and copper losses result in an efficiency of 80-90 per cent. for a good transformer. At the lower frequencies, both resistance and primary inductance reduce the efficiency still further, and a similar loss takes place at the higher frequencies above 3,000 c/s due to magnetic leakage.

The following points must be considered in the design of transformers for use in the anode circuit of a single valve.

(a) Primary inductance to be as large as possible to avoid loss at the lower frequencies.

(b) Leakage inductance to be as low as possible.

(c) D.C. resistance must be kept low.

(d) The transformer must present a primary impedance equal to the recommended optimum load for the valve in use.

Unfortunately, to fulfil (a) makes (b) and (c) difficult, furthermore the anode current of the valve increases the difficulty as it becomes necessary to introduce a gap into the magnetic circuit to prevent undue saturation; thus the design can therefore only be a compromise and with the usual small transformer a falling off in response is experienced at both ends of the A.F. range.

Matching

Assuming the normal case where there are less turns on the secondary than on the primary winding, then the secondary voltage, E_s, will be less than the voltage, E_p, across the primary by the turns ratio of the windings. Representing the turns ratio as N, then

$$E_s = \frac{E_p}{N} \text{ or } I_s = I_p \times N$$

This is so, for discounting losses the product of the primary current and voltage is equal to the product of the secondary current and voltage.

The impedance $Z_s = E_s/I_s$ may be written as

$$Z_s = \frac{\dfrac{E_p}{N}}{I_p N} = \frac{\dfrac{E_p}{I_p}}{N^2}$$

As E_p/I_p is the primary impedance, the last equation can be rewritten as $Z_s = Z_p/N^2$ or $Z_p = N^2 Z_s$. Taking an example where the optimum load is given as 8,000Ω and the speech coil impedance is 2.6Ω, then

$$N = \sqrt{\frac{Z_p}{Z_s}} = 65:1.$$

When designing a transformer allowances can be made for losses by including the primary and secondary D.C. resistance in the calculation, or if convenient, it is worth while arranging for a tapped secondary or an experimental transformer so that ratios above and below the calculated values are available. Under test conditions measurements will indicate the best ratio for maximum undistorted output.

Paper interleaved layers and thorough impregnation with either bakelite varnish or bitumen is necessary in view of the high voltages that exist at the anode connexion for maximum output conditions. The voltage for a correctly loaded secondary can reach twice the applied D.C. anode volts and possibly four times for an open circuit secondary load which obviously calls for high insulation properties between windings and frame.

1st. A.F. Stage

It is possible to compute the stage gain required from V_3 so that when a predetermined signal input voltage is applied to the aerial circuit the output stage is fully loaded. It was suggested previously that the maximum output should occur at a point where A.V.C. takes control and reference to Curve I, Fig. 33, will show that this occurs when a modulated signal of 45 μV is applied to the aerial circuit. From the previous figures given for H.F./I.F. gain and a detector efficiency of 80 per cent. this produces an A.F. signal at the junction of R_{11} R_{12} of 0.175 volts r.m.s. for a 30 per cent. modulated signal. In practice the maximum output will be secured for smaller inputs as the modulation level of a broadcasting station will in all probability be higher than 30 per cent.

Using the output pentodes mentioned earlier a peak input signal of 3.5 volts will be required to produce about 2-2½ watts, in which case a gain of 20 must be realised between the grids of V_3 and V_4. The value of the anode load for V_3 can be obtained by plotting a load-line on the static anode characteristic curves for the par-

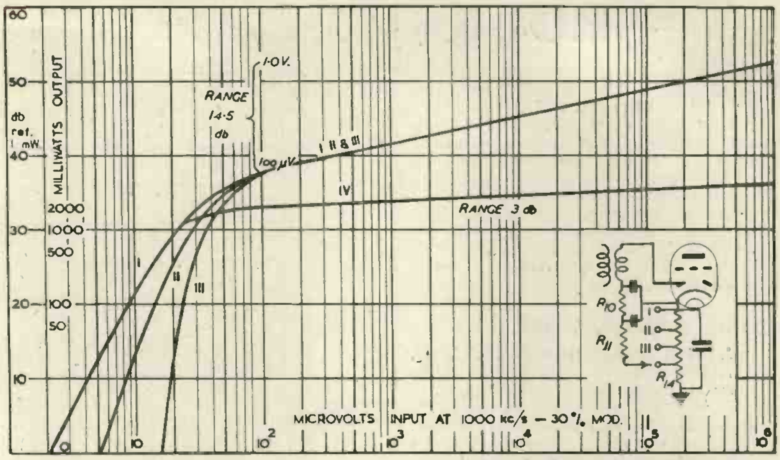


Fig. 33. A.V.C. characteristics. (i) For two controlled stages of conventional receiver with —3 v. delay. (ii) as (i) but with —1 v. at signal diode. (iii) With —1.5v. (iv) Is the same as (i) but with A.V.C. also applied to the 1st A.F. stage.

ticular valve used, or obtained from the fact that since the A.C. voltage generated in the output circuit of the valve is in series with its anode resistance, R_p, the A.C. voltage developed across the load resistance, R_L depends upon the relative values of R_p and R_L. The amplification is then given by μ (voltage gain) = $\frac{g_m R_p R_L}{R_p + R_L}$, or in terms of $R_L = \frac{\mu}{\left(g_m - \frac{\mu}{R_p}\right)}$

Using a valve with an anode resistance of 13,500Ω and a mutual conductance of 2mA./V, the value of R_{15} is 50,000Ω but the anode loading, R_L is in the order of 39,000Ω due to the shunting effect of the grid and tone control circuits which are effectively in parallel with R_{15}. The condensers $C_{14}C_{16}C_{17}$ contribute little towards this as calculation will show that R_{14}, R_{13} and R_{16} in parallel come to about 41,000Ω. From the previous formula the voltage gain works out to 20 and this amplification is obtained under working conditions. A much higher amplification is possible from the first A.F. stage, but then the signal handling capacity of the valve is reduced, and the possibility of distortion increased accordingly. The aim is then to limit the amplification to the required amount in order to keep penultimate distortion at a minimum.

Tone Control

Tone control is necessary in all A.F. amplifiers as individual tastes and operating conditions vary. Three major reasons for its inclusion are (i) to compensate for lack of balance in the broadcast signal, (ii) reduce the effect of interference due to heterodynes, record scratch and other forms of electrical noises (iii) to compensate for local acoustic imperfections. To carry out the job effectively, tone compensation circuits require separate continuously variable bass and treble controls so that any desired aural response characteristic is obtainable.

For the normal commercial receiver this is ruled out primarily on the grounds of cost, and only the simplest type of control is to be found in most receivers.

In Fig. 34, the tone control circuit follows the customary practice of using a condenser in series with a variable resistance, C_{17} R_{13}, which functions as a variable impedance between limits determined by the magnitude of the reactance of C_{17} relative to R_{13}. The circuit is effectively in shunt with the anode load, R_{15}, and so reduces the anode loading, hence the amplification, as the frequency of the amplified signal increases. Despite the fact that it can only reduce the top response, an apparent bass boost takes place as the acoustic output from the receiver will be lacking in treble.

The values of C and R must be chosen so that in the maximum treble position, *i.e.*, when R_{13} is at a maximum, it does not appreciably shunt the anode load, and with R_{13} set to zero, produces a change of -15 db. at the treble end—which appears to cover normal requirements.

C_{13} has a reactance of 1,600 ohms at 5,000 c/s, this modifying the anode load of V_3 so as to produce a gain of 3 from this stage, which corresponds to -15 db. (at 5,000 c/s) on the normal gain of 20 at 400 c/s. It will be obvious that at the highest A.F. frequencies V_3 will work into an abnormally low anode load with the net result of amplitude distortion occurring due to the curvature of the anode characteristics. As shown earlier in connexion with the diode, a valve

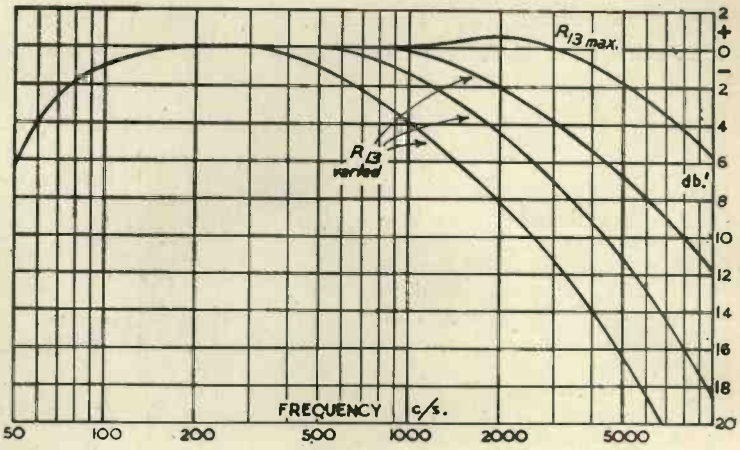


Fig. 35. Effect of tone control R_{13} C_{17} (Fig. 34) on audio frequency response.

should work into a resistance much greater than its internal resistance, R_p, to avoid this effect.

A small detail worth mentioning is the location of the tone control circuit. If connected in the output stage, V_4, where large peak voltages exist, it becomes necessary to employ a condenser with a fairly high working voltage, whereas connected as shown in Fig. 34, the voltage across the condenser is comparatively small and the possibility of breakdown is less likely.

A.F. Response

Fig. 35 illustrates the response characteristics obtained for different positions of R_{13}. The drop of -7 db. at 50 c/s. is due to the low primary inductance of the output transformer, but which will be hardly noticeable aurally as the speaker is almost certain to possess a bass resonance around 70-100 c/s. when fitted into the receiver cabinet.

The overall response can be obtained by referring back to Curve III given in Fig. 25 and adding the db. loss at each frequency to those given in Fig. 35, Curve I. This will then represent the "electrical" response to a broadcast signal applied to the A.E. terminals.

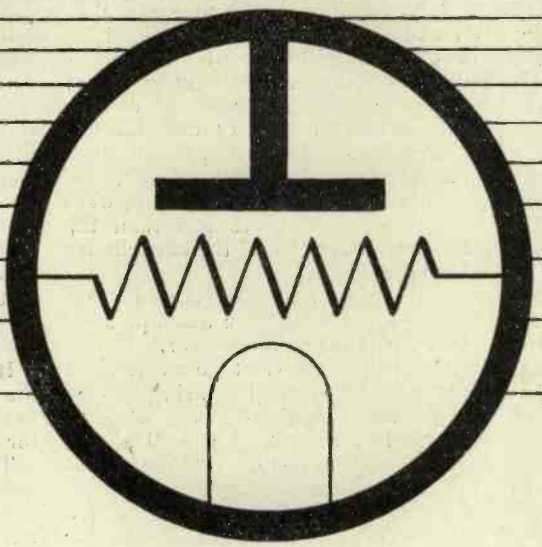
Variations in acoustic output from a loudspeaker are due to the resonances introduced by the moving parts, and consequently tend to emphasise certain frequencies; furthermore, the directional characteristics introduce distortion as the higher frequencies are mainly concentrated along the axis of projection, so that at positions to one side of the speaker the reproduction loses crispness and definition. It will be obvious that attention must be given to the selection of a loudspeaker in order to obtain the most pleasing response. This detail, however, will depend upon individual tastes and operating conditions.

The design of a modern receiver is very complex as many of the problems involved cannot be fully dealt with by paper work alone, and require considerable developmental work before satisfactory results are achieved. The proceeding parts have covered, stage by stage, only the basic features encountered in the design of what has been described as an average receiver.

Provided these points are appreciated by the less experienced reader, he is then in a position to consider the effect of circuit modification upon an existing design so that it meets with his particular requirements. There are many ways of achieving a certain result especially where the experimenter is concerned, as the question of cost is relatively unimportant and time usually unlimited.

A certain amount of test gear is necessary, and this is where the experimenter is sometimes at a disadvantage. It is difficult to state precisely what test gear will be required, for so many different kinds of measurements have to be made, but the possession of (1) valve voltmeter for R.F. measurements, (2) C.R. Oscillograph, (3) A.F. output meter, (4) L.F. oscillator, and (5) an R.F. signal generator is most essential as they are in constant use. Other equipment, such as microammeters, inductance/capacity bridges, condenser and resistance boxes, A.F. attenuator, harmonic analyser, and so on, add up to a very imposing list, but are only necessary for commercial development work. For a detailed description of recommended receiver measurements, readers are referred to previous issues of this journal.*

* *Electronic Engineering* for August, September October, 1941.



An Improved Inter-electrode Capacitance Meter

THE use of a fluorescent lamp as a voltage stabiliser in a photo-cell amplifier circuit was suggested in 1940 by M. H. Sweet,* who used it for regulating the heater voltage of 4 6.3v. valves consuming 1.2 amperes. A similar regulating circuit has been successfully adopted by A. H. Barco in an improved interelectrode capacitance measurer, described in *R.C.A. Review* for April, 1942†. The original model of this meter was battery operated and contained an oscillator which supplied radio frequency, at about 500 kc/s. with voltage of a few hundred volts. This potential was impressed upon a calibrated standard concentric-cylinder capacitor in series with a thermocouple. For measurements the standard was set at maximum capacitance and the thermocouple indicator reading noted. The unknown inter-electrode capacitance was then shunted across the standard and the standard decreased until the thermocouple indicator returned to the initial reading. The inter-electrode capacitance could be read directly on an inverted scale on the standard capacitor.

The stability requirements of the anode supply indicated the need for some form of voltage-regulated power supply. Additional measures of stability and regulation were obtained by generous usage of gas-filled regulators (of the VR-150 and VR-105 types) at various critical points in the circuit. The final result was an excellent order of stability. Line voltage variations (anode power unit input only—not including heater supply) of ± 15 % were found to cause errors less than the limit of readability of the capacitance standard scales.

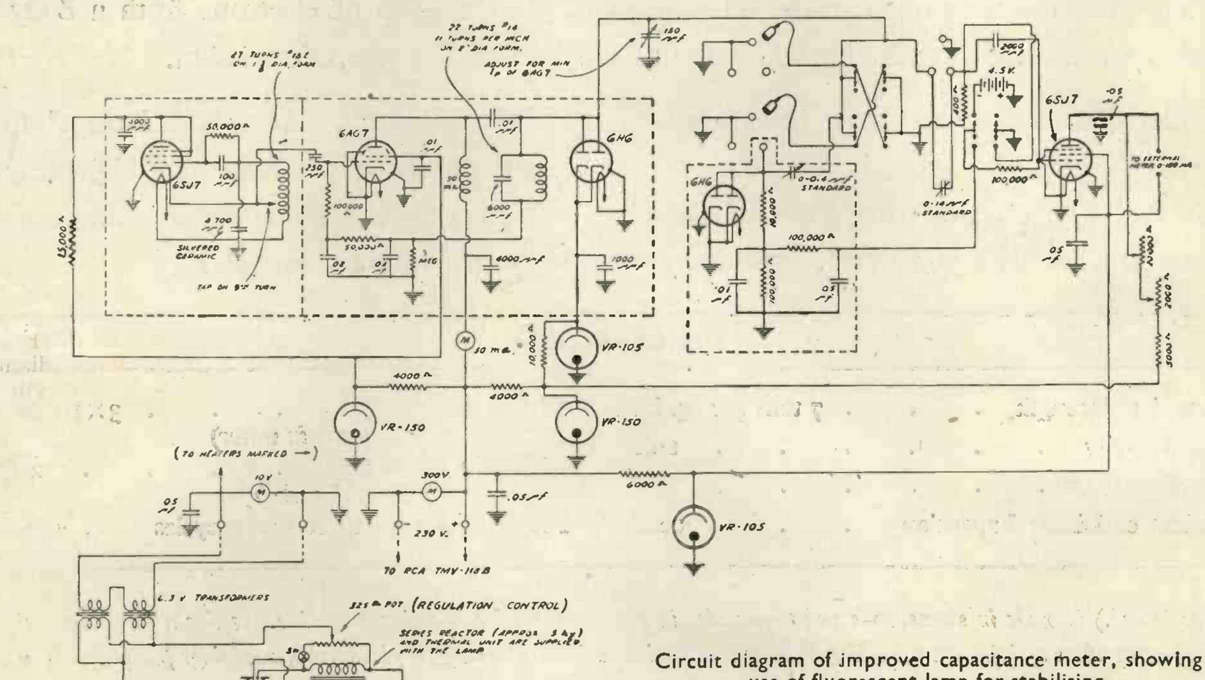
Preliminary tests revealed that heater-voltage variation was a possible source of considerable error. It was found that line-voltage variations caused proportional variations in capacitance readings, i.e., a 5 % change in line voltage caused a 5 % error. A satisfactory solution was finally evolved utilising the negative-resistance characteristic of a fluorescent lamp to compensate for line-voltage variations. The lamp is supplied with a series current-limiting reactance and operates with a tube drop of about 65 volts. This voltage decreases as the line voltage is increased and vice versa. Obviously, at some intermediate point on the series current-limiting impedance the net effect of line-voltage variations is zero. The regulated output is obtained by using a shunt voltage divided across a 0.5-henry series reactor. It must be noted that fluorescent lamps of the type used (GE 20W. 3,500° C. white) employ heated emitters for starting purposes. These heaters are automatically cut out of the circuit after a short time by a thermal relay actuated by the current through the limiting reactor. Hence, it is necessary to remove all external loads from the circuit for the starting of the fluorescent lamp and then subsequently to apply the load to be regulated.

It was found advantageous to mount the fluorescent lamp above the capacitance meter so that it might be used for illumination as well as regulation.

The older model of the capacitance meter contained two valve voltmeters, one for each of the two capacitance ranges. The one used on the low sensitivity range (0-14 $\mu\mu$F) was a anode circuit detector and the one on the high sensitivity side (0-0.4 $\mu\mu$F) consisted of a diode detector followed by a d-c amplifier.

The undesirable features of the self-excited oscillator were greatly minimised in the improved model of the

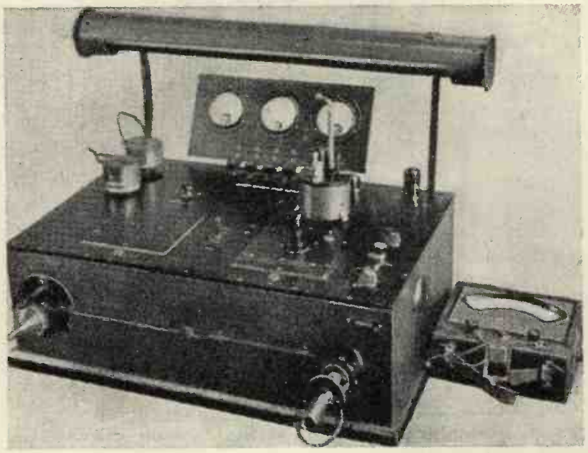
* *Electronics*, Aug. 1940.
† p. 434.



Circuit diagram of improved capacitance meter, showing use of fluorescent lamp for stabilising.

capacitance meter. This was done by using a highly stable oscillator to drive a buffer-amplifier stage. The amplifier stage output tank impedance was made as low as conveniently possible (1,000 ohms or so) without having to provide excessive anode-current swings in the amplifier stage. The high-capacitance tank in the output circuit served to swamp the detuning effect of shunt capacitances and the low-tuned impedance provided good regulation for conductive loads. Additional stability of the radio-frequency potential was achieved by the use of an automatic-gain control on the amplifier stage.

In the improved model, heater type tubes 6SJ7 and 6H6 are employed. The 6SJ7 operates as a conventional anode-circuit detector on the low-sensitivity range and as the d-c amplifier for the 6H6 detector output on the high-sensitivity range, the change being accomplished by a d-p-d-t switch.



View of meter with stabilising lamp mounted above panel for illumination.

Colour Television—*Continued from page* 200.

slotted rods which hold the filters follows the scanning lines.

Receiving Equipment

Practically any good monochromatic television receiver design may be made the basis of a colour television receiver, and a photograph of a 9 in. tube colour receiver is shown in Fig. 11. The additional equipment required is the colour disk with its driving and synchronising gear.

An effect which sometimes tends to destroy interlacing is due to electrostatic charges which accumulate on the moving colour disk, and variations in the charge produce movement of the scanning lines as the disk rotates. It is possible to remove this charge with a semi-conducting coating on the tube face or by other electrostatic screening means.

Colour disks have been made of metal or of Lucite, Plexiglass, etc., and Wratten filters can be obtained coated on acetate stock which can be riveted to the plastic or metal disk.

The generation of a properly shaped filter disk for the receiver is shown in Fig. 12. This type of shape is suitable for a receiving or transmitting tube with short decay or storage times. The curve which is obtained in Fig. 12 is an envelope of the position of a scanning line as traced on to the filter which is moving with the line. The required filter shape for a given mechanical arrangement is obtained by developing curves which make allowance for positive and negative tolerances to take care of fluctuation in the disk position, viewing angle and screen decay, Fig. 13. Generally, the minimum disk diameter is about twice the outside diameter of the tube plus 1 or 2 inches.

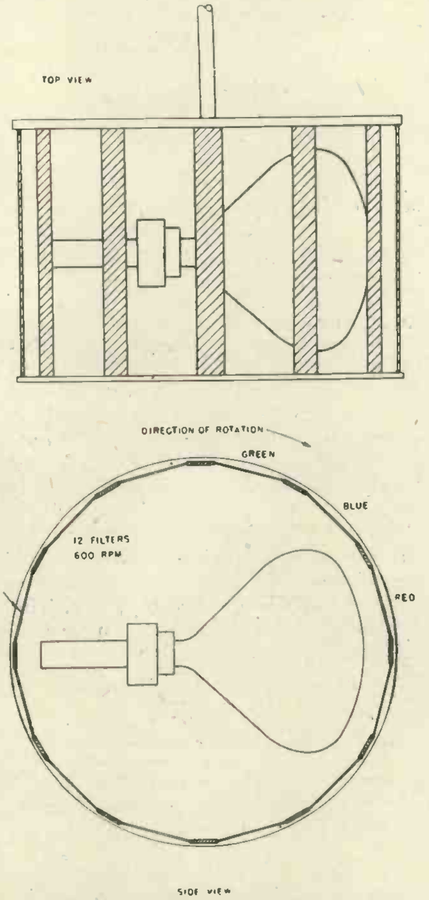


Fig. 14. Diagrammatic view of drum receiver for colour television.

The optimum location will be determined by such factors as the distance from the disk shaft to the picture frame, but can be determined for any particular arrangement.

Colour drums have been used at the receiver as well as at the transmitter instead of colour filters. A short cathode-ray tube can be placed within the drum (Fig. 14). The drum is designed for a lower speed of revolution than is usually possible for the disk. Successful drums have been built to operate at a speed of 600 revolutions per minute or one half the usual disk speed.

The small table model receiver utilises a cathode-ray tube developed especially for colour pictures. The screen of the tube is flat and has the exact shape of the final image. The tube produces a picture equivalent in size to that of a conventional 7 in. round tube. Thus the colour disk is 15 inches in diameter. A 10 in. lens with a focal length of 12 inches, which is built into the receiver, increases the image to correspond to that of a conventional 9-in. tube. Owing to low magnification, distortion and decrease of the viewing angle are appreciably reduced.

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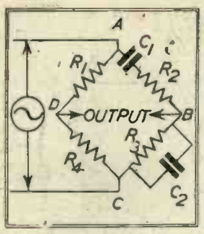
An extensive bibliography is appended to the original paper.



The Wien Bridge

and some of its Applications

By J. S. WORTHINGTON

THE Wien bridge is a simple and efficient frequency selective network of resistors and condensers, but until recently it was seldom met outside the laboratory. The fundamental circuit is shown in Fig 1 above, and with slight modifications it can be used for such purposes as oscillator frequency selection, frequency measurement, capacity and condenser power factor measurement and heterodyne reduction.

For use in audio-frequency measurements the unknown frequency is applied across A and C (preferably through an isolating transformer) and resistors R_2 and R_3 varied simultaneously until a null point is indicated (*e.g.*, by phones) across D and B. For this to occur the voltages at points B and D must obviously be equal and in phase with one another. The phase shift through AD and DC (*i.e.*, R_1 and R_4) is zero, so that the phase shifts through AB and BC must cancel out for this condition to appear :—

Phase shift through AB is $\dfrac{1}{R_2\omega C_1}$

Phase shift through BC is $R_3\omega C_2$

$$\therefore \frac{1}{R_2\omega C_1} = R_3\omega C_2$$

$$\therefore \omega^2 = \frac{1}{R_2R_3C_1C_2}$$

$$\text{and} \quad f = \frac{1}{2\pi\sqrt{R_2R_3C_1C_2}}$$

If R_2 is made equal to R_3, and C_1 equal to C_2 this expression reduces to :—

$$f = \frac{1}{2\pi RC} \qquad \dots \quad (1)$$

For balance of the bridge at frequency "f" :—

$$\frac{Z_{AD}}{Z_{DC}} = \frac{Z_{AC}}{Z_{BC}} \quad \text{i.e} \quad \frac{R_1}{R_4} = \frac{R - \dfrac{j}{\omega C}}{\dfrac{\dfrac{Rj}{\omega C}}{R - \dfrac{j}{\omega C}}}$$

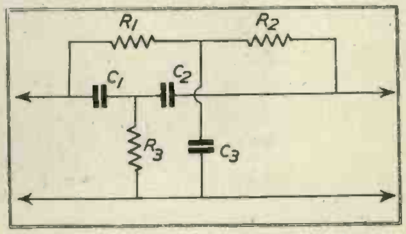


Fig. 2. Equivalent Parallel-T Network.

$$\frac{R_1}{R_4} = -\left(R - \frac{j}{\omega C}\right)^2 \frac{\omega C}{Rj}$$

$$= \frac{-\left(R^2 - \dfrac{2Rj}{\omega C} - \dfrac{1}{\omega^2C^2}\right)\omega C}{Rj}$$

$$= \frac{-R^2\omega C + 2Rj + \dfrac{1}{\omega C}}{Rj}$$

substituting $\omega = \dfrac{1}{RC}$:—

$$\frac{R_1}{R_4} = \frac{2Rj}{Rj} = \frac{2}{1}$$

$$\therefore R_1 = 2R_4$$

An ingenious application of this bridge is due to Dr. Woodward[1] in which the sharp null point is used to eliminate an interfering heterodyne in the output of a receiver. The output of the receiver is fed across A and C, and the phones connected across B and D. The network is tuned to eliminate any frequency required by varying R_2 and R_3 simultaneously. In the original model these were ordinary volume controls ganged together. The values of the components can be calculated from the equation

$$f = \frac{1}{2\pi RC}$$

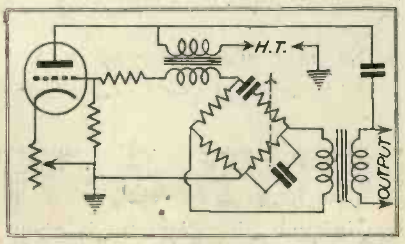


Fig. 3. Audio-frequency Oscillator using Wien Bridge.

and the values of R_1 and R_4 chosen to give a suitable impedance to match the external circuits, bearing in mind the fact that R_1 must always equal $2R_4$.

Audio oscillators using the Wien bridge, or the equivalent parallel-T network have been introduced commercially.[2] The equivalent parallel-T network is shown in Fig. 2, its advantage being that it has a common ground for input and output circuits. It can be tuned by varying three ganged resistors, R_1, R_2, and R_3, or three ganged condensers, C_1, C_2, and C_3. If in this circuit $C_1 = C_2 = \frac{1}{2}C_3$ and $R_1 = R_2 = 2R_3$, the frequency is given by

$$f = \frac{1}{2\pi R_1C_1} \qquad \dots \quad (2)$$

An audio oscillator using these circuits is essentially an amplifier with both positive and negative feedback, the positive feedback occurring at all

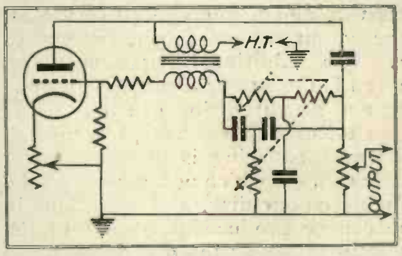


Fig. 4. Audio-frequency Oscillator using parallel-T network.

frequencies, and the negative feedback just cancelling it at all frequencies except that for which the bridge is balanced. By varying this null point the frequency of oscillation is varied.[3]

Simple circuits of oscillators using the Wien bridge or its equivalent parallel-T network for negative feedback are shown in Figs. 3 and 4 respectively. Such an oscillator has many advantages over conventional heterodyne and LC oscillators. The heterodyne oscillator is comparatively complicated and requires several valves and tuned circuits, and an LC oscillator requires an iron-cored coil and large condensers to tune to audio frequencies, with the result that the frequency is usually varied in steps rather than continuously, and the output wave-form is distorted. The Wien bridge oscillator can be made very

Sorry U.S.A.—but

ERSIN MULTICORE

THE SOLDER WIRE WITH 3 CORES
OF NON-CORROSIVE ERSIN FLUX

will not be with you again until after Victory

It sounds like "coals to Newcastle" but U.S.A., the home of cored solder, found it worth while to import British made Ersin Multicore—the A.D. approved solder wire with three cores of non-corrosive Ersin flux, although Ersin Multicore, after paying freight and duty, cost American manufacturers fifty per cent more than American made solders.

Why? With the correct, rapid soldering technique, H.R. or "dry" joints are impossible with Ersin Multicore, and production is speeded up. Present conditions do not permit the export of Multicore to U.S.A., but after Victory, U.S.A. manufacturers will, once again, be able to use "the finest cored solder in the world."

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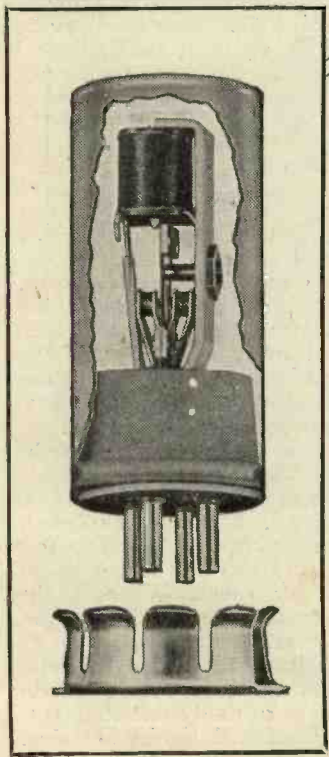

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simply, is tuned continuously by resistors, and gives an output practically free from harmonics owing to the sharp characteristic curve of the degenerative network.

An audio oscillator using a circuit related to the Wien bridge was described in *Electronic Engineering*, June, 1941. This oscillator consists of an amplifier with negligible internal phase shift, the output of which is fed back to the input through the

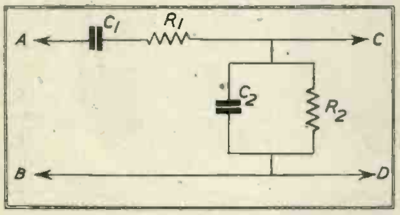

Fig. 5. Phase-shifting Network.

phase shifting network shown in Fig. 5.[4] With an alternating voltage applied across A and B, at one frequency, and one frequency only, is the voltage appearing across C and D in phase with the input voltage. If the gain of the amplifier exceeds the attenuation of the network, oscillations are generated, the frequency being controlled by varying R_1 and R_2 simultaneously. In the instrument in question these resistances are controlled by decade switches, arranged to read as conductances, making the oscillator direct reading in frequency.

In Fig. 6, let $R_1 = R_2$, and $C_1 = C_2$, A voltage, E_1, is applied across the network.

Let the resultant current through the network be I.

Then, for AB :—I leads by $\tan^{-1}\dfrac{1}{R\omega C}$

This current flowing through BC gives rise to a voltage, E_o, across BC, lagging on I by $\tan^{-1} R\omega C$.

Therefore, when

$$\frac{1}{R\omega C} = R\omega C$$

then E_1 is in phase with E_o,

$$\therefore \omega = \frac{1}{RC} \text{ and } f = \frac{1}{2\pi RC}$$

A fault of most communications receivers is the inability to use A.G.C. when the B.F.O. is switched on. The output from this local oscillator acts on the A.G.C. rectifier exactly as would a received signal, producing a control voltage which greatly reduces receiver gain. This situation can be avoided by injecting the B.F.O. voltage in such a way that it does not affect the A.G.C. circuit. This is accomplished by using the normal second detector as one arm of a Wien

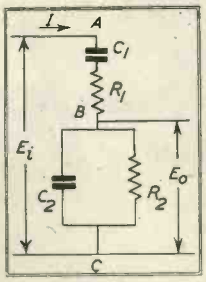

Fig. 6

bridge and applying the I.F. output voltage and the B.F.O. voltage across opposite corners of the bridge.

The basic circuit is shown in Fig. 7. The second detector is represented by the parallel impedance Z_1 and is balanced by the series impedance Z_2, at the intermediate frequency. R and R represent the halves of a centre-tapped I.F. output transformer, and the circuit forms a Wien bridge, balanced for the intermediate frequency. The second detector, being one arm of the bridge, receives the outputs of both the I.F. amplifier and the B.F.O. Any audible difference in their frequencies will therefore appear in the output of the second detector. The A.G.C. rectifier, on the other hand, receives only the output of the I.F. amplifier and so does not respond to the B.F.O. voltage.

The essential condition is that $Z_1 = Z_2$ at the intermediate frequency. Z_1 consists of the inter-electrode capacity of the signal diode, C_1, shunted by its input resistance, R_1. When the load resistance of a diode is high, the input resistance is nearly equal to half the value of the load resistance. Z_2 consists of C_2 and R_2 in series, and the values can be calculated from the equation :—

$$R_2 - \frac{j}{\omega C_2} = \frac{\dfrac{R_1}{\omega C_1}}{R_1 - \dfrac{j}{\omega C_1}}$$

To avoid rapid changes in receiver sensitivity as the incoming signal is keyed it is necessary to use a large time constant in the A.G.C. circuit, or else arrange for an adjustable A.G.C. delay voltage equal to the peak value of the signal applied to the A.G.C. rectifier. If this condition is satisfied smooth reception of C.W. signals will be assured and the output level will be held constant by the A.G.C.

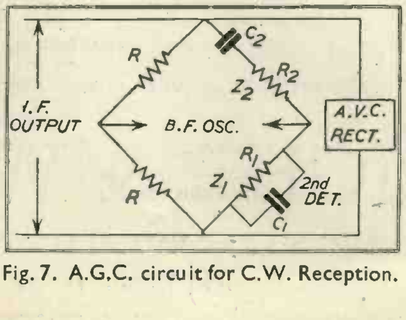

Fig. 7. A.G.C. circuit for C.W. Reception.

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1 "Hetrofil — an Aid to Selectivity." *QST* September, 1939.

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3 R. W. Gaywood, *QST*, January, 1941.

4 "A New Frequency Generator," *Electronic Engineering*, June, 1941.

5 "A.V.C. for C.W. Reception," *QST*, January, 1941.

Conducting Silver Films on Refractory Materials

IN the manufacture of several classes of small electrical components it is necessary to produce conducting films on refractory materials, as for example, on mica, in the manufacture of small condensers. Such films may be produced by applying a paint composed of silver oxide suspended in a drying oil, such as linseed oil, and subsequently reducing the oxide by heating, to produce an adherent silver film.

Drying oil residues in the presence of easily reducible oxides are, however, very inflammable. In a factory process the powder which collects in spray booths, trunking, etc., presents a serious fire risk. This risk may be eliminated by replacing the drying oil with a non-inflammable binder. Such a binder must distil completely below the decomposition temperature of the silver oxide, be chemically stable and, either alone or with solvent, have a viscosity suitable for the preparation of a paint. The chlorinated diphenyl resins are suitable materials for the purpose, and a viscous liquid type distilling below 400° C. should be chosen.

The paint is prepared by grinding silver oxide in the medium and then thinning with turpentine or petroleum solvent to a suitable consistency. The silver oxide may be prepared by precipitating from a nitrate solution.

Methods of application may be chosen to suit the particular component. The refractory surface may be either sprayed, brushed, or dipped, and after allowing to airdry for a few minutes, transferred to an oven at 150-160° C. for 1 hour. When dry the parts are placed in a furnace at 500° C. for 10 minutes.

The resultant film will be fully adherent and will be sufficiently coherent to take a highly reflecting surface if polished with a mild abrasive.

This process has been suggested by Mr. R. King of the Gramophone Co., Ltd.

CORRESPONDENCE

Laboratory Slang

DEAR SIR,—After reading the article on the Stroboscopic Use of the C.R. Tube in your issue of August, I feel moved to make the following observations:

(a) The use, in scientific publications, of "laboratory slang"—by which I mean language which, while common knowledge in one's laboratory, is unknown outside its walls—should be deprecated.

(b) The author has used the word "scalariform," which presumably means "shaped like a scalar." The noun "scalar" is unknown to me. Would not "stepped" or "step-like" have been clearer?—Yours

Southsea. DR. IPFEED.

DEAR SIR,—In the August issue of *Electronic Engineering* you published a very interesting article on the stroboscopic use of the cathode-ray tube by G. Bocking. I feel impelled to write to protest against the use of incorrect terminology and of slang terms which are not well known. My chief objections in the case of Mr. Bocking's paper are to the use of the following terms.

(1) Long-tailed pair.
(2) Scalariform.
(3) Risers.

With regard to (1), the term is sometimes used to indicate two valves having a common cathode coupling. This term is applied because the characteristic curve of the combination is as shown in the accompanying figure and this characteristic is somewhat similar to that of a valve in which complete cut-off cannot be obtained. The characteristic of such a valve is generally described as having a long-tail. The term "long-tailed pair" is, however, very new and, to a reader who is not familiar with the special attributes of a pair of cathode coupled valves, the term would be unlikely to have any meaning. There is, in my opinion, no objection whatever to the use of the term provided an explanation as to its meaning is given at the beginning of the paper in which it is employed.

With regard to (2), the word "scalariform" as defined by the dictionary means "shaped like a ladder, as of the cells and vessels of plants, veins of insect wings, etc." It will thus be seen that the word "scalariform" is intended to convey an idea which is almost in one dimension and is not intended to convey the idea of a staircase which is the construction put upon it by the author.

With regard to (3), the word "risers" is defined in the dictionary as "the vertical members of a staircase" and in view of the author's use of the word "scalariform," it is presumed that the word "riser" is intended to mean that part of the waveform at which the rate of charge of the condenser C is increased above the normal by the application of an increase of potential to the grid of the valve VT in Fig. 6.

To sum up, the use of written or spoken laboratory slang is desirable provided the meaning is clear and accurate and that the words save time and are euphonious. Where a relatively new slang term is being introduced, it is absolutely essential that the precise meaning attached to the word shall be explained.

I trust that the author will appreciate that I am attacking his language and not his information which is, as stated above, of considerable interest. I would like to venture the opinion that the finest English is obtained by the use of those words which are the most simple and well-known and which are carefully chosen in order to express the precise intention of the author beyond possibility of doubt.

While discussing these aspects of the method of writing an article, I would like also to draw attention to the fact that the drawings are based on principles which have been out of date for 30 years. This unfortunately applies to the drawings in practically every journal in the country. This, however, is another matter and will no doubt bring about endless discussion.—Yours faithfully,

Edgware. O. S. PUCKLE.

EDITORIAL NOTE:

Mr. Bocking is unfortunately in hospital, so is unable to reply to his critics in this issue.

While we agree that "long-tailed" pair is an obscure, if picturesque, phrase, we cannot see the objection to taking a technical term out of Zoology or Biology if it suits the subject under discussion.

Mr. Puckle's criticism of the drawings presumably refers to the "cross-over loops" on the wiring, as the symbols are in the majority of cases B.S.I. 530 recommendations. It is our opinion that the cross-over loops render the wiring diagram completely unambiguous, which the "straight lines and dots" do not, and we are retaining them for this reason.

Any comments or suggestions from readers on improving or clarifying diagrams would be welcomed.

Resonance Indicator

DEAR SIR,—In connexion with the resonance indicator described by Mr. J. M. A. Lenihan in your current issue, I should like to point out so far as I am aware the use of the Magic Eye as an indicator with cathode resistor giving increased sensitivity due to positive feed-back was described for the first time in my article in *Wireless World* of March, 1942.

Although the circuit shown by Mr. Lenihan is effective at medium and high radio frequencies, it is not applicable to frequencies below about 50 kc/s due to the fact that there is insufficient capacitance in the triode anode circuit to act as a reservoir of energy. The electron beam, having negligible inertia, follows the rapid variations in the triode anode current instead of taking up a mean position corresponding to increased anode current. The result is blurring of the shadow and indefinite indications. At higher frequencies the shunt capacitance of the valve electrodes is sufficient. It appears, therefore, that the circuit should include a by-pass condenser across the anode load in order to obtain true anode bend detection at all frequencies.

Secondly, I think that the author is unwise to assume that because the feed-back is positive for D.C. and low frequencies, that it will be necessarily so at high R.F. In an actual case investigated in our laboratory the cathode voltage was exactly *in phase* with the grid voltage, thus giving *negative* feed-back. This is probably the cause of the greatly reduced sensitivity mentioned in the last paragraph of Mr. Lenihan's article, as otherwise this is difficult to explain. The addition of a cathode by-pass condenser at this point gave a 100 per cent. increase in sensitivity instead of the decrease we are led to expect. The anomalous phase relations at high R.F. are probably due to the stray capacitances in the valve. I feel that the addition of a cathode by-pass condenser would be advisable to remove any A.C. feed-back, whether positive or negative, as most of the increase in sensitivity is due to the presence of DC feed-back, which is unaffected thereby.—Yours faithfully, GEORGE A. HAY.

Newcastle-on-Tyne.

Upside Down

MR. A. W. RUSSELL (Mullard W.S. Co.) points out that the inversion of the block of Fig. 2b in his note on p. 173, last month's issue, gives the impression that the mean grid potential is positive, whereas it should be negative as stated in the text.

ABSTRACTS OF ELECTRONIC LITERATURE

THERMIONIC DEVICES

Voltage-Controlled Electron Multipliers

(B. J. Thompson)

The application of secondary emission multiplication to conventional grid-controlled amplifier tubes is discussed from the viewpoints of practical voltage gain per stage of amplification, signal to noise ratio and ultra high frequency applications. It is pointed out that the gain per stage is limited by the practical output current and the quotient of transconductance by current (N) and that electron multiplication increases the gain only as it permits the attaining of higher values of N. If the output current is assumed to be 20 mA and N is taken as one milliamperes per volt—little, if any, better than could be achieved without multiplication. If N is assumed to be 11.6 (the theoretical maximum for conventional grid control with a cathode temperature of 1,000 degrees Kelvin) the output transconductance could be greater than 200 mA per volt per milliampere.

Higher values of N might be obtained by some other method of control. In this case, the ultimate limits of transconductance would be set by the difficulty in stabilising the effective control-electrode bias voltage.

The signal-to-noise ratio of the voltage-controlled multiplier is determined chiefly by the input system of the multiplier the multiplier being a relatively noiseless amplifier following this input system. The noise level of the input system is determined by the input transconductance. If the use of a multiplier leads to reduced input transconductance the noise level will be increased as compared with conventional tubes.

The principle advantages to be attained from the use of the multiplier are found in applications where input loading and input capacity are serious. The reduction in transconductance of the input system for a given over-all gain which is permissible, leads to a corresponding reduction of input conductance (whether arising from electron-transit time or lead effects) and input capacity.

—*Proc. I.R.E.*, Vol. 29, No. 11 (1941), page 583.

Time-Delay Relays

(Seymour)

The possible uses of an automatic time delay switch are reviewed. It is explained that a short time delay action may be obtained by shunting a high capacity condenser across the terminals of a magnetic relay. A schematic diagram of a single-valve relay unit is shown and its mode of action explained. Three methods of varying the time delay are given and it is pointed out that this relay arrangement has the advantage of economy.

—*Electrician*, August 7, 1942, page 142.*

A New Instrument for Recording Transient Phenomena

(S. J. Begun)

In many cases it is of great importance to study phenomena which do not occur periodically. Such phenomena, called transients, arise for example, at the time a short circuit appears in a power line, during the discharge of the condenser of a spot welding machine, during the starting of electric machinery and on very many other occasions.

Most of the instruments available thus far for investigation of transients employ the method of film recording which has the disadvantage of requiring a development process.

A new transient recorder has been developed, employing magnetic tape recording as a means of preserving the record of a transient and steadily repeating this record on the screen of an oscilloscope. This method has the advantage that it requires no processing and that the same magnetic carrier can be used continuously without loss of material.

—*Trans. Am. I.E.E.*, Vol. 61 (1942), page 175.

THEORY

Theory of the Magnetron

(L. Brillouin)

This study of the magnetron, in its static condition, has been pursued for the first time with strict regard to the influence of the space charge.

The speed of rotation of the electrons around the filament results from the magnetic field; hence, an apparent radial potential $P(r)$ may be defined, governing the radial movements of the electrons. This potential $P(r)$, at distance (r), is equal to the real potential $V(r)$ less the critical potential $V_0(r)$ which would just prevent the electrons from reaching an anode placed at r. The apparent potential $P(r)$ may be studied methodically; in the vicinity of the filament, it approaches Langmuir's result; as the distance becomes of the order of

$$r = L = \sqrt{-\frac{eI}{m\omega_H{}^3}}$$

the potential P departs from Langmuir's curve and then corresponds with a function $P_\omega(r)$ typical of the magnetron. Under certain conditions, in the region of correspondence $(r = L)$, a maximum potential may be obtained such that the electronic energy $eP = -|e|P$ is a minimum. The disposition of the apparent potential P in a magnetron under these conditions then becomes analogous to that found in a Barkhausen valve with a strongly positive grid and a moderate anode potential. It is thus possible to foresee that, under prescribed conditions, oscillations could be maintained in the magnetron.

—*Electrical Communications*, Vol. 20, No. 2, page 112.

Propagation Constant and Characteristic Impedance

(K. Spangenberg)

A reference sheet which deals with the graphical and analytical methods for determining the characteristic impedance and propagation constant of transmission lines having high losses resulting from series resistance.

—*Electronics*, Vol. 15, No. 8 (1942), page 57.

The Inclined Rhombic Aerial

(S. W. Harrison)

In this paper the use of an inclined rhombic aerial as a means of reducing the effect of fading is discussed. Equations are given for determining the angle at which a rhombic aerial should be inclined to obtain a desired response pattern at various elevation angles in the vertical plane containing the major axis.

—*Proc., I.R.E.*, Vol. 30, No. 5 (1942), page 241.

INDUSTRY

An Improved X-Ray Tube for Diffraction Analysis

(R. R. Machlett)

A new X-ray tube for diffractive analysis has been produced which affords much greater X-ray beam intensity than has been possible with previous designs, using thin sheets of malleable beryllium produced by addition of minute amounts of titanium, for windows in the metal tube wall. The low inherent absorption of such windows and their electrical and mechanical sturdiness are factors in the improvement of beam intensity. A shockproof form of the new diffraction analysis tube is also described.

—*Jour. App. Phys.*, Vol. 13, No. 6 (1942), page 398.

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

A Universal Test Meter
The Taylor "Model 90"

THE improved Model 90 is a universal test meter for general radio and electrical work giving a wide range of voltage and current measurements on A.C. and D.C. A resistance scale is also provided and an output scale calibrated in decibels, making a total of 38 ranges.

The sensitivity is 1,000 ohms per volt and the indicating meter is of the well-known Taylor square type (Model 425) with ample scale length and a knife-edge pointer.

The complete instrument measures 8 in. by 5 in. by 4 in. deep and weighs under $4\frac{1}{2}$ lb.

Ranges:

The instrument will read A.C. and D.C. voltages from 2.5 to 1,000, and in addition there is a 0.25 v. D.C. range corresponding to a full-scale deflection of 2.5 mA.

D.C. milliamperes can be read from 1.0 to 2,500 and A.C. milliamperes over the same range with the exception of the lowest.

Resistance measurements can be made from 0-10,000-100,000 – 1 megohm by setting the appropriate scale factor on the range switch.

By inserting a 54,000 ohm resistance in series with the unknown and a 60 v. battery the resistance range can be extended to 100 times the full scale reading, or 10 megohms.

The layout of the panel is seen from the photograph, the centre switch being the range selector and the left hand switch the A.C.-D.C. selector. On the right is the zero adjuster for the resistance measurement. A particularly useful addition to the model is the on-off switch for the meter located between the input sockets.

The interior of the meter is robust. The resistances are firmly anchored and the wiring well carried out. A condensed table of instructions is fitted on the underside of the case, and with the booklet provided there is no difficulty in understanding the connexions for the various ranges.

The meter has a 9 v. bias battery fitted internally for use in measuring resistance, which is replaced by removing the metal cover underneath the case.

Altogether this is a well-made instrument which is particularly suitable for hard wear in a workshop or laboratory and the price (£11 11s. 0d.) is very reasonable for the present time.

(Messrs. Taylor Instruments, Ltd., Slough, Bucks.)

PATENTS RECORD

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TELEVISION

Colour Television

In a colour television system where a revolving colour disk is used, the disk cannot be placed close to the photoelectric surface of the transmitting tube owing to the vacuum chamber. Neither can it be placed close to the fluorescent screen in cathode ray tubes (types in which an opaque screen is used) as the screen is some distance within the vacuum vessel.

According to the invention, an image of the photo electric mosaic surface is projected on to the colour filter and the image to be transmitted is projected on to the projected image of the surface. The result obtained is the same as if the surface were transposed to the plane of the colour filter and the image to be transmitted were projected on to the surface through a colour filter rotating in the same plane so that the dividing line between two adjacent colours will be sharply focused on the image.—*J. L. Baird, Patent No.* 546,470.

THERMIONIC DEVICES

Half Wave Modulation Electron Discharge Tube

An ultra high frequency discharge tube of the velocity modulation type including a toroidal chamber having inner and outer electrically condensating walls in the form of coaxial cylinders. These have openings arranged in a line transverse to the common axis. Arrangements are made for projecting a stream of electrons through the openings and also for varying the electrical resonance frequency.—*Standard Telephones and Cables, Ltd. (Assignees of C. V. Litton). Patent No.* 546,774.

RADIO

RECEPTION

Automatic Control System for a Carrier-wave Signal Receiver

The invention relates to an automatic control system for a carrier-wave signal receiver in which a sharp discriminator network is used. In order that signals may be automatically tuned in, a voltage is added to the grid bias in addition to any unidirectional-bias potential developed by the discriminator.

Thus, for off-tune signals that would otherwise be ineffective to develop a discriminator bias, the system " pulls in " to them because the control valve is sweeping the signal input to the network through a relatively wide frequency range. If the off-tune signal is in this swept frequency range, discriminator voltages are developed each time the oscillator traverses the frequency that causes the correct intermediate frequency to be developed.—*Hazeltine Corporation (Assignees of R. L. Freeman), Patent No.* 541,393.

Frequency Modulation Detector

The frequency modulated i.f. signals are applied to the primary of a transformer having two secondary windings, to which are connected the grid-cathode circuits of two valves V_1 and V_2. The windings are so connected that the input voltage develops opposite voltages on the two valves. Between the anode and cathode of V_1 is the primary of the a.f. transformer and a condenser C_1 in series, the anode and cathode of V_2 being connected directly across the condenser.

When an input is applied to the i.f. transformer each valve is conducting in turn and the condenser C_1 will be alternately charged through V_1 and discharged through V_2. Since the action of the circuit depends only on the reversal of polarity of the input wave and is independent of the amplitude, the current in the a.f. transformer primary will be proportional to the frequency of the input, provided that the period during which the valve V_1 is conducting is long enough to allow the condenser C_1 to charge to the H.T. supply voltage and the conducting period of V_2 is similarly long enough for a complete discharge.

The number of charging current pulses in a given time is proportional to the frequency of the input signal, and hence the mean current is a direct equivalent of the frequency modulated input.

—*Standard Telephones and Cables, Ltd., Patent No.* 546,721.

CIRCUITS

Constant Mean Frequency in Transmission by F. M. Carrier Wave

Arrangements for controlling the frequency of carrier wave generators in frequency modulation transmitters.

The master oscillation and the subharmonic of the carrier are made to beat together to produce two-phase beat currents. These are applied to drive a motor which effects a mechanical adjustment of a frequency controlling element in the carrier wave generator. So long as the two waves remain in synchronism the armature of the motor will not rotate. If the mean value of the carrier frequency should drift away from synchronism, the motor armature will rotate in a direction depending on the sense of the drift and by an amount proportional to the cumulative phase shift produced.

The rotation adjusts the carrier generator control element in such a direction as to compensate the drift and to bring about a correction of the mean frequency. Suppression of effects of rapid fluctuations of the phase is effected by mechanically filtering the rotational motion of the motor so that only slow variations representative of changes in the mean carrier frequency are transmitted to the frequency controlling element.—*Electrical Research Products. Patent No.* 546,677.

Improvements in H.F. Oscillators

An h.f. oscillator arranged so that the anode and control grid of the electron discharge device used, form opposite corners of a balanced bridge circuit, to the remaining corners of which the cathode, power supply and earth are connected.

In such a circuit high frequency currents transmitted through the apparatus and appearing across the first two corners of the bridge can produce no voltage between the remaining two corners of the bridge.

—*British Thomson-Houston Co., Ltd., Patent No.* 540,400.

Variable Reactance Devices

To control the resonant frequency of an oscillatory circuit, the effect is produced by current in the controlled circuit by means of the reactor valve.

This invention relates to variable reactance circuits and its object is to reduce the resistive loading effect by phasing the grid voltage so that it has a component in phase opposition to the alternating anode voltage.

Two reactor valves are connected in parallel and their mutual conductance varied in opposite direction, so that the two valves produce a cumulative net result. Another arrangement is to operate the two reactor valves as class B amplifier valves so that the anode current of one valve only is functioning. In this manner one valve provides an effective shunt inductive reactance to the oscillatory circuit while the other provides an effective shunt capacitive reactance. — *Marconi's Wireless Telegraphy Co., Ltd. (Assignees of L. C. Smith). Patent No.* 545,879.

Classified Announcements

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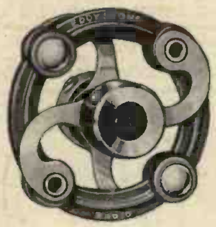
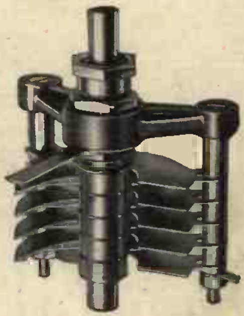


FLEXIBLE COUPLER

Cat. No. 1009. Flexible but free from backlash. Diameter 1½in. for ¼in. spindle. The insulation gap is sufficient for voltages in the region of 2,000, making it valuable for transmitting equipment.



MICRODENSERS

Cat. Nos. 1093/94 and 1129/30/31. Available for a wide range of capacities. Brass construction with all vanes soldered to give low series resistance at high frequency. Design ensures constantly maintained capacity.

LEAD THROUGH INSULATORS

For carrying high frequency leads through metal chassis and screens with the least loss. Cones of glazed Frequentite flanged to self centre into metal at each side. Cat. No. 1018.

SHORT WAVE H.F. CHOKES

Several types available, including Cat. No. 1010. Resistance 20 ohms. Inductance 1.25 millihenrys. 5-180 metres. Windings are arranged to give minimum self capacity and small external field.

STAND-OFF INSULATORS

Cat. No. 916

Ideal for mounting transmitting inductances, meters, spacing inside aerial feeders and for any insulating where high voltages are carried. Made in vitreous porcelain of special quality. Shaped to give very long leakage path and high mechanical strength.

6-PIN COIL FORMERS

Cat. Nos. 1002 and 1003. A "D.L.9" eight-ribbed moulding is used of 1⅜in. diam. giving 2⅛in. winding space. Particularly useful for the construction of efficient low-loss inductances for receivers and low-powered transmitters.

To obtain consistently good results on short-wave work, well-designed components are essential. Eddystone have specialised on precision components for H.F. and U.H.F. work and have developed a range second to none for efficiency. A wide selection is now obtainable from stock at Webb's Radio and further items are being rapidly added.

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The home of the Short-Wave Enthusiast

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