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Electronic Engineering

INCORPORATING ELECTRONICS, TELEVISION AND SHORT WAVE WORLD

PRINCIPAL CONTENTS

2/- DEC., 1942

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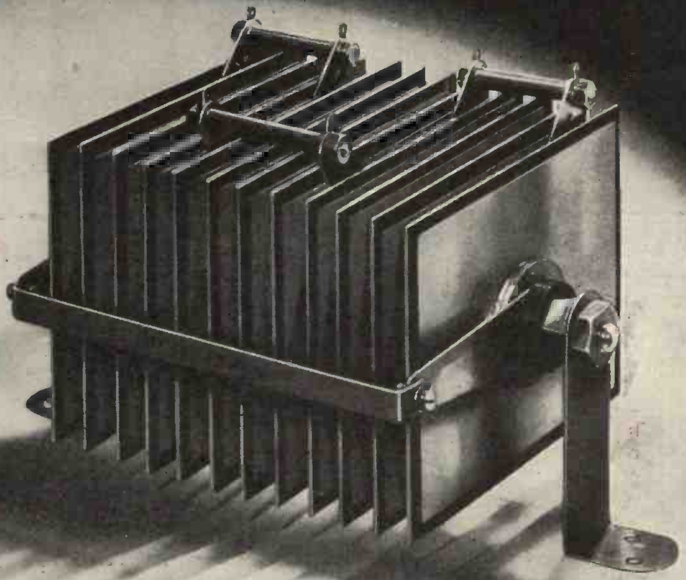

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Electronic Engineering

DECEMBER, 1942

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CONTENTS

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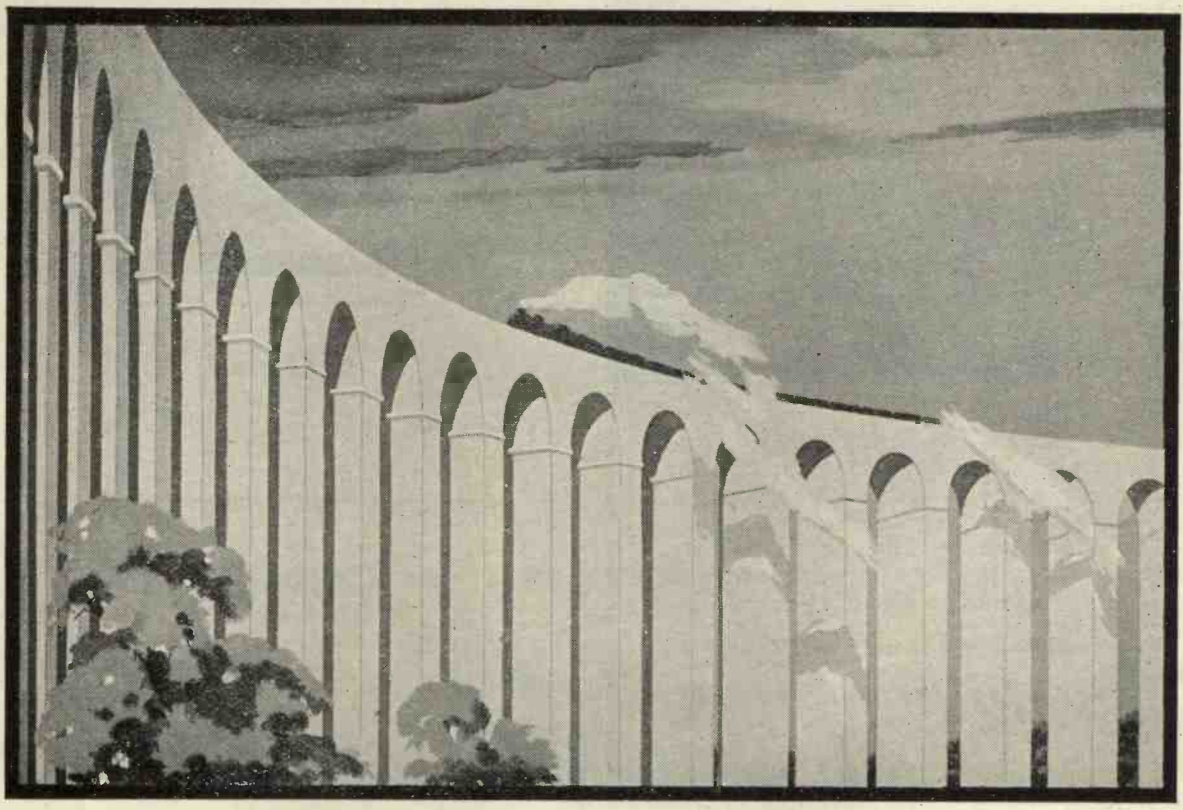

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TELEGRAMS :
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Alchemy

Robert Boyle, who has been called the father of modern chemistry, was not above making a bit of a mystery when it suited his purpose.

In one of his lesser known pamphlets, published anonymously in 1678 and re-published later under his own name, he describes an experiment in metallurgical chemistry which is interesting to recall in these days of new and wonderful alloys.

The pamphlet is written in the form of an account of a discourse given by Boyle to a company of fellow chemists (who frequently interrupted him) and is verbose and tedious in parts.

Briefly, it tells of his meeting with a mysterious Stranger from the East (they always are !) who gave him a screw of paper containing a rare Powder, together with Advice on how to experiment with it.

Boyle was " both surprized and troubled to find so very little Powder " on opening the paper, and did not dare to weigh it in case he lost it on the balance. However, having got the necessary witnesses—a wise precaution in those days of alchemy and fraud—he proceeded to melt 2 drachms of gold in a crucible and cast the Powder in, allowing it to " defuse " throughout the gold.

The results, when cool, astonished him as the gold became converted into a brittle alloy " looking like Brass or worse " with totally different properties from the original metal.

This is where our opinion of Boyle as an experimenter is lowered. In reply to a question from a sceptic in the audience he had to confess that he did not test the alloy with nitric acid as he had none in stock. His account of the subsequent tests is vague, although he did manage to find the density, which was approximately 15 (gold is 19).

After hinting at other still more wonderful properties of the Powder, which he was not at liberty to disclose, he concludes " This Experiment plainly shews that Gold, though confessedly the most homogeneous and least mutable of metals, may be chang'd both as to malleableness, colour, homogeneity and Specifick Gravity "—and by a " despicable quantity of Matter."

On another page in this issue, Dr. Hunt, with less mystery, describes how copper can be changed both as to malleableness, colour, and other characteristics usually associated with it, by an almost despicable quantity of beryllium.

Beryllium was first isolated in 1838, or was it ?

If the mysterious Stranger had only given Boyle a little more of the wonderful Powder . . .

At the conclusion of the meeting in 1678, one of the chemists uttered what might fairly be called a mouthful :

" We ought not to be so forward as many men of otherwise great parts are wont to be in prescribing narrow limits to the power of Nature or Art, and in deriding those who pretend to believe uncommon things in Chymistry as Cheats or Credulous."

ELECTRONIC ENGINEERING MONOGRAPHS.

The Publishers of this Journal are arranging to issue at intervals a series of Technical Monographs on specially interesting radio and electronic subjects.

They hope that the production of such monographs will enable radio engineers and students to build up at reasonable cost a library of specialised information which could only be otherwise obtained from a number of books and periodicals.

The monographs will be in the form of octavo size booklets of uniform style, bound in stiff red paper.

The first has already appeared under the title

FREQUENCY MODULATION
by K. R. Sturley, Ph.D., A.M.I.E.E.

based on the articles which have recently appeared in this Journal, revised by the author.

Copies can be obtained from technical booksellers, price 2/6, or direct from the Publishers at the above address, price 2/8 including postage.

" An account of a Degradation of Gold, made by an Anti-Elixir : A Strange Chemical Narrative." (Herringman, London.)

Acknowledgment is made to Cdr. R. T. Gould, whose book " *Enigmas* " furnished the reference.

Beryllium Copper

A High Strength Alloy for Electrical and Instrument Springs

By L. B. HUNT, Ph.D., A.R.C.S., M.Sc.*

THE development of electronic engineering has presented many problems to the metallurgist, who has been prevailed upon during the past twenty years or so to provide a number of alloys having special combinations of electrical and physical properties. The majority of these newer materials have been required for use inside the radio valve or the cathode-ray tube, but in other cases special properties have been found necessary or advantageous in the purely electrical parts of the circuit.

One of the principal requirements of designers in this industry is a reliable material for use as current-carrying springs, possessing the optimum combination of elastic strength and electrical conductivity. Many such applications arise in radio engineering, chiefly in components such as wave change switches, volume controls, appliance plugs, micro-switches, fuse clips, vibrators and measuring instruments. For moderately light mechanical duty the use of phosphor bronze or nickel silver for these parts has become standard practice, but in many cases springs in these materials are found to be deficient in either mechanical or electrical properties, or sometimes in both.

In this event, beryllium-copper is found to give the necessary combination of high elastic and fatigue strengths with high conductivity. As will be seen from the detailed properties of this alloy given in Table I, beryllium-copper offers an outstanding balance of electrical and mechanical properties which make it suitable for springs designed for very severe duty.

Other non-ferrous spring materials, such as phosphor-bronze and nickel-silver, acquire their strength and elastic properties by severe cold work, but the provision of spring properties in this way gives rise to difficulties when further working is necessary in forming the spring from strip or wire. Where the design of such parts is relatively simple, hard rolled strip can, of course, be used, but in many cases more or less severe drawing or forming operations are involved and only annealed or lightly cold rolled materials are permissible, with the obvious sacrifice in spring properties. Further, the electrical properties of these alloys are by no means high, normally being of the order of 10 to 15 per cent. that of copper. Beryllium-copper not only offers greatly superior elastic and fatigue strengths, but also improved conductivity and the great advantage of being readily fabricated in the soft state and then heat treated to give the required spring properties. Springs or other parts of complicated shape may thus be produced in beryllium-copper which would be incapable of production in a hard rolled material, and can then be given a tensile strength up to 90 or 95 tons per square inch, with a limit of proportionality of 50 tons per square inch.

Composition and Properties

Table I gives a comparison of the properties of beryllium-copper with the average properties of extra-hard phosphor bronze (6 per cent. tin, 0.2 per cent. phosphorus) and high carbon spring steel in strip form. It will be seen that the spring properties of beryllium-copper are very little inferior to those of high carbon steel and greatly superior to those of extra-hard phosphor bronze, a material which is not capable of any great degree of drawing or forming. (The Erichsen value of this temper of phosphor bronze strip will be approximately 3.0 mm.). Beryllium-copper is not, of course, magnetic, and has the further advantages of excellent resistance to corrosion and wear. The high limit of proportionality of beryllium-copper permits of very much higher safe stresses than may be employed when using phosphor bronze, while its

Fig. 1. Typical beryllium-copper components of electrical and pressure measuring instruments, snap action switches and radio apparatus.

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* Mallory Metallurgical Products, Ltd.

higher endurance or fatigue limit also enables considerably greater working stresses to be employed in springs subject to rapid and continuous deflection. Thus a beryllium-copper spring can be of smaller cross section than one of phosphor bronze, without loss of current-carrying capacity, and its length can be decreased to obtain the desired deflection. For these reasons this alloy is being employed to a greatly increased extent for parts of electrical equipment and instruments when no other material can provide the combination of workability with high mechanical properties in the finished product.

Beryllium-copper has been known to metallurgists and engineers for some years now, but has not always been as readily available as was desirable. Normally the alloy has consisted of 2.25 to 2.5 beryllium, with the balance copper, but more recently it has been found that the addition of a small percentage of cobalt gives somewhat improved and more uniform physical properties. Such an alloy has been made available in this country during the past two years and it is with this material, known as Mallory 73 beryllium-copper, that this article is concerned. The nominal composition of Mallory 73 is beryllium 2.0 per cent., cobalt 0.5 per cent. and copper the remainder. It is available in all those forms likely to be required by electrical, radio, and instrument manufacturers, including strip down to 0.002 inch thick, wire down to 0.001 inch in diameter, hair-spring strip down to 0.010 inch by 0.0005 inch, round and oval tubing, and contact bi-metal strip having a silver inlay or facing.

The use of this material for electrical and instrument springs has other subsidiary advantages besides those already outlined. For example, in the final heat treatment of the springs or other parts all internal stresses are relieved, with the result that the possibility of " season cracking " does not arise, while undesirable effects due to creep, drift or elastic hysteresis are also reduced to very small proportion. Again, a single spring in beryllium-copper can replace an assembly of steel backing spring and current-conducting lead. Where slightly elevated temperatures are involved in service, beryllium-copper has the further advantage that it may be used up to 200° C. without loss of strength. The modulus of elasticity of beryllium-copper, approximately 18.5×10^6 lb. per square inch, is slightly higher than that of phosphor bronze, but very much lower than the figure for steel. Springs in this alloy, therefore, have considerable amplitude of movement within the very great elastic range.

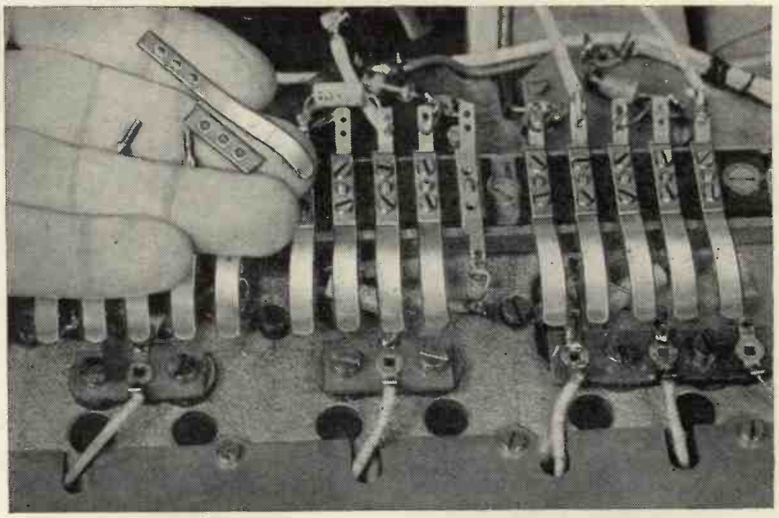

Fig. 3. An assembly of beryllium-copper contact blades, having an electrodeposited coating of silver and rhodium, employed in radio apparatus manufactured by Murphy Radio Limited.

Production of Springs

No difficulty is experienced in the manufacture of spring parts in beryllium-copper provided that due care is taken in the selection of the most appropriate temper of strip or wire. It should be borne in mind, however, that beryllium-copper work-hardens rather more rapidly than the usual brasses and bronzes. For this reason a softer temper should be used than would normally be chosen in the case of phosphor bronze, for example.

Mallory 73 Beryllium-Copper has been made available in four tempers of sheet or strip, the quarter-hard, half-hard and hard tempers being obtained by imparting increasing reductions in thickness, while wire is similarly available in three tempers. The properties associated with each of these tempers are set out in Tables II and III. The properties given here are, of course, those of the alloy before final heat treatment. For deep drawing purposes, for bending flat over very small radii and where no " spring-back " is allowable in forming, beryllium-copper strip should be employed in the soft or fully annealed state. The quarter-hard temper should in general be employed for light drawing operations and for work such as severe bending over a comparatively small radius. For more simple forming operations the half-hard temper should be used, while the full hard condition is preferable for the blanking of flat springs or parts involving only light bending or setting operations. In the case of Mallory 73 beryllium-copper in wire form similar conditions obtain. The annealed temper should be employed for very severe bending and similar operations, while the quarter hard and half hard tempers will be found more suitable for progressively less severe forming. For coil springs wire in the half hard condition is normally most satisfactory.

In any given case beryllium-copper should be used in the hardest temper which will satisfactorily form the parts, as the final strength and hardness obtained on heat treatment will be slightly higher for increased amounts of cold work, while the possibility of distortion occurring during heat treatment is greatly reduced with the harder tempers.

The manufacture of hair springs for instruments is carried out in exactly the same manner as for phosphor bronze. After winding in the usual barrel or " pill-box," springs are given their heat treatment as described below. This serves not only to set the spring, but also to develop the required spring properties. It will be seen, therefore, that while other materials normally suffer some reduction of elastic limit during the setting operation, beryllium-copper actually increases very markedly in strength.

No attempt should be made to carry out bending or setting operations of any kind on beryllium-copper after final heat treatment on account of its very high elastic limit and spring properties.

Heat Treatment

The exceptional physical properties of beryllium-copper are obtained, as already mentioned, by a simple low-temperature heat treatment. This process is a function of time and temperature, hardening occurring more

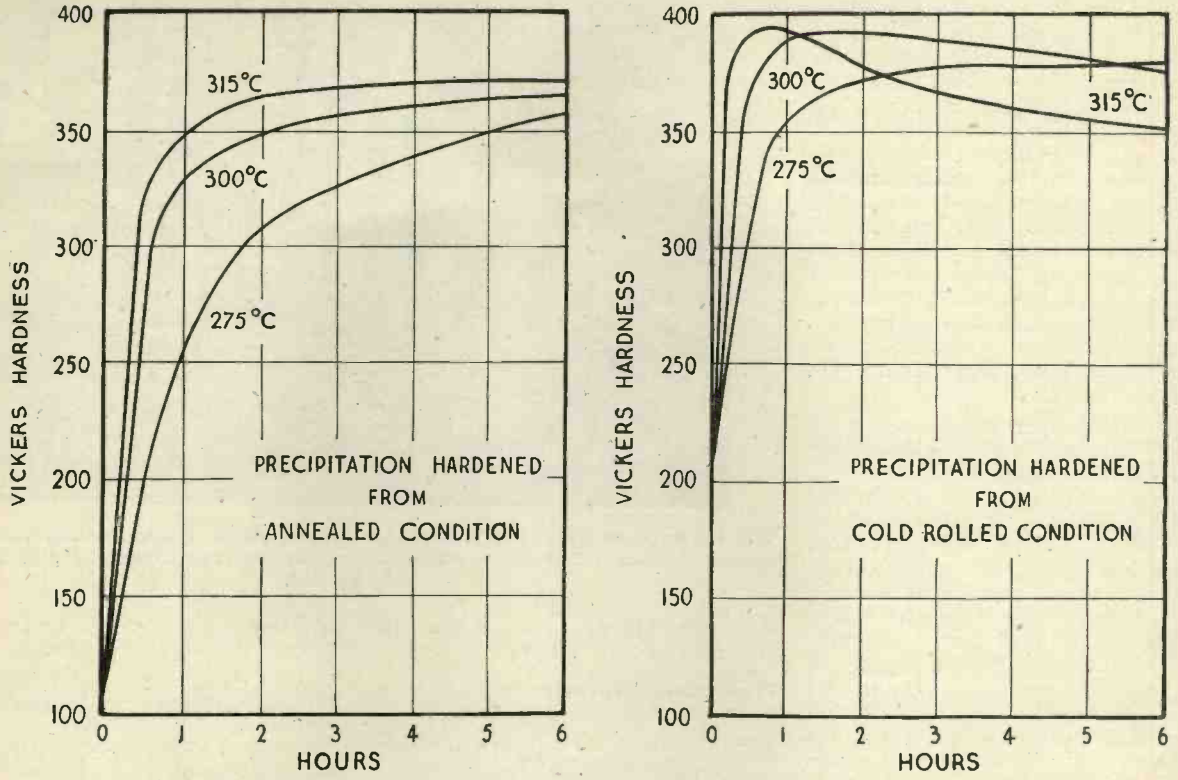

Fig. 2. The effect of time and temperature on the hardness of Mallory 73 beryllium-copper developed during heat treatment.

rapidly at higher temperatures. If heating is continued for too long, or if the temperature is too high, some re-softening is brought about, more particularly in hard rolled strip or hard drawn wire. Material in the annealed or lightly worked condition is less susceptible to this effect but, on the other hand, develops slightly lower properties and requires a longer time and a rather higher temperature for optimum hardening. These points are illustrated in the two sets of curves. In the left hand curves, for annealed material, it may be seen that at 275° C. hardening takes place very slowly. At 300° C. hardening occurs more rapidly, but considerable time is required to develop maximum properties. At a higher temperature, however, such as 315° C., hardening is more rapid and satisfactory results can be obtained at this temperature after a treatment lasting 2 hours.

The curves on the right hand side, for strip rolled half-hard, show first that appreciably higher hardness figures can be obtained by comparison with annealed material, and secondly that this is brought about at rather lower temperatures and with shorter times. At 275° C. hardening is still too slow to be convenient, but it is possible to hold the alloy at this temperature for a considerable time without re-softening. At 300° C. hardening occurs fairly rapidly and a reasonable latitude is possible in the time for which the work may be held at temperature. At 315° C., however, a very rapid hardening is followed by some re-softening, and the permissible margin in the time factor may be found to be too narrow for satisfactory commercial heat treatment. Best results are thus obtained with the following treatments:

From annealed condition, 2 hours at 310 to 320° C.

From quarter hard or half hard condition, $1\frac{1}{2}$ hours at 300 to 310° C.

From hard condition, 1 hour at 300 to 310° C.

This treatment may be carried out in an air furnace provided that the parts are suitable for subsequent cleaning by pickling. If this is not the case, as with very small or delicate springs for example, the use of a hydrogen or nitrogen atmosphere furnace is necessary. Care should be taken in heat treatment that the charge is actually at the temperature shown on the pyrometer, the hot junction of the thermo-couple being placed as close as possible to the work. The specified period for heat treatment should be taken only from the time when the charge reaches the proper temperature. Cooling may be carried out either by quenching or in air.

To avoid distortion during heat treatment, finished springs and similar parts should be adequately supported. Flat parts may be clamped between heavy steel plates, while formed parts should be clamped in appropriate jigs, or wired together in packs if their design makes this possible. Small parts may alternatively be laid in trays and surrounded with sand. The risk of distortion is much less in the harder tempers.

The times and temperatures given above represent good average practice for the development of maximum hardness and spring properties, but in certain cases or for special requirements it may be necessary to modify these treatments to some extent. Advice in this connexion will readily be given by the manufacturers of beryllium-copper, who will also undertake the heat treatment of finished parts for users who are not themselves equipped with suitable furnaces.

Soldering and Brazing

Beryllium-copper may be soft soldered without difficulty *after* final heat treatment. It may also be silver brazed successfully, but in this case the jointing operation must be carried out *before* heat treatment.

TABLE I
Properties of Beryllium-Copper compared with Phosphor Bronze and Spring Steel

	Beryllium-Copper	Phosphor Bronze	Spring Steel
Ultimate stress, tons/sq. in.	80—90	45—50	95—100
Limit of proportionality, tons/sq. in. ...	47—50	28—33	60—65
Brinell hardness	350—420	185—225	400—450
Fatigue limit, tons/sq. in.	± 20—22	± 10—12	± 35—40
Modulus of elasticity, lb./sq. in., x 10^6 ...	18—19	15—17	29—30
Electrical conductivity, per cent. I.A.C.S.	25—30	12—15	3—4

TABLE II
Tempers of Mallory 73 Beryllium-Copper Strip
(Properties before final Heat Treatment)

Temper	Vickers Hardness	Tensile Strength tons per sq. in.	Elongation per cent.	Erichsen value Millimetres
Annealed ...	100—120	30—35	45—60	8.0—9.0
Quarter Hard ...	160—185	35—40	25—35	5.5—6.5
Half Hard ...	200—225	40—50	10—20	4.0—5.0
Hard	230—250	50—55	5—10	2.5—3.0

TABLE III
Tempers of Mallory 73 Beryllium-Copper Wire
(Properties before final Heat Treatment)

Temper	Tensile Strength tons per sq. in.	Elongation per cent. on 2 in.
Annealed	35—40	25—40
Quarter Hard ...	40—50	5—10
Half Hard	50—60	2—5

When parts to be soft soldered can be washed after soldering, it is recommended that a zinc chloride flux be used, and in these cases no difficulty is found in making joints, using either a soldering bit or a torch. If the material has not been pickled, the surfaces to be joined should be cleaned with fine emery. After the joint is made the assembly must be thoroughly washed to remove all traces of flux and prevent subsequent corrosion. Many articles, including the majority of electrical components, cannot be washed after soldering, while it is frequently stipulated in electrical work that no flux other than resin may be used. In these cases good joints can be made with certainty, using resin as a flux, provided that sufficient care is taken to clean the surfaces immediately before soldering. When parts are in an inaccessible position for cleaning, the surfaces to be joined should be tinned before assembly. Resin cored solder may be used, in which case no additional flux need be applied. Successful results have also been obtained by painting with a solution of resin in methylated spirits. Once the surfaces have been tinned they may readily be joined on any later occasion, using a resin cored solder.

When high-strength joints are required beryllium-copper may be successfully silver brazed, but this must be done before final heat treatment as the temperatures involved are, of course, much higher than the hardening temperatures. A low melting point silver brazing alloy such as Easy-flo (melting point 630° C.) should be used. Mechanical cleaning before silver brazing is not essential, and equally good results may be obtained on oxidised as on freshly pickled material. The only degreasing treatment required is a preliminary warming with the blowpipe flame before applying the Easy-flo flux, which is easily able to cope with the oxide film formed on beryllium-copper. The alloy behaves very similarly to brass in brazing and the normal technique may be employed using gas-air or oxy-gas heating. The normal heat treatment should, of course, be carried out after brazing.

It will be seen that beryllium-copper again offers a marked advantage by comparison with other non-ferrous alloys in that strong and effective joints can be made, or electrical contacts fitted, by silver soldering, without the final mechanical properties being thereby affected. This procedure is not, of course, possible on any alloy relying upon cold work for its strength and hardness, as the temperature involved would be more than sufficient to anneal the material.

Advantages in Electronic Engineering

It will be appreciated from the foregoing that electrical, radio and instrument manufacturers now have available for special purposes a material possessing greater strength and elastic properties than any other non-ferrous alloy, a conductivity higher than those of the brasses and bronzes, and at the same time possessing excellent workability. As compared with the normal type of spring material which relies upon cold work for its elastic properties, in the case of beryllium-copper a simple heat treatment develops these properties, with a consequent simplification of manufacturing and assembly problems.

Again the possibility of making joints or attaching electrical contacts by silver brazing is of considerable importance in electrical engineering, while the availability of silver or silver alloy faced bi-metal strip opens up still further possibilities in designing contact arrangements for strength and simplicity.

For components required to carry radio-frequency currents, beryllium-copper may also be given an electro-deposit of silver, or of silver plus rhodium. In this way a heavy duty spring is provided, with the silver to carry the R.F. current and the rhodium to protect the silver from tarnish and to maintain a low, constant contact resistance. Rhodium and silver electrodeposits are also being employed on beryllium-copper to give a wear resisting contact surface in wave change switches subject to particularly severe duty.

These advantages of beryllium-copper have rapidly been recognised by radio engineers, and its use has become standardised for a variety of contact blades and clips, special relay springs, micro-switch blades, fuse clips and similar parts. Unfortunately, many of the interesting applications of this unique alloy cannot be described in detail until they are once more directed towards human comfort and enjoyment.

Wave Analysis

Part II — Analysis of Semi-Periodic Wave-Forms

By K. BOURNE[*]

IN a previous article (Part I),[†] wave-forms were classified under four broad headings. Those waves in Class D, *i.e.,* "transient," are normally treated only from a theoretical point of view, and there is little of practical value to be gained by attempting to analyse these electrically. The wave-forms in Class C, *i.e.,* continuous but non-recurrent, are of great importance in acoustics, including as they do the human voice and musical instrument tones, single or in combination. When the wave-form is continuously changing, the analysis must obviously be made as quickly as possible if it is to give a close approximation to the instantaneous spectrum of the wave; however, many musical tones and sustained vowel sounds are sufficiently steady in pitch and free of incidental noise that they may be considered as Class A or B waves. Even in this case, speed is very desirable; fortunately, a very moderate degree of accuracy is sufficient, since in most cases the tones under investigation cannot be exactly reproduced, and an approximate spectrum gives enough information regarding the character of the sound.

It is convenient to note here a fundamental division in the methods of analysis of these waves: between methods which utilise the tone at the time of its production, and those which record it by some graphical method, and analyse the record mechanically or optically. The first group has obvious advantages in disposing of the necessity for recording apparatus, and in giving a direct and rapid result; on the other hand, the graphical record of a wave-form (and in practice this means a record on sound-film, variable-area or variable density) gives a permanent result, correct at any and every instant to the limits of accuracy of the recording apparatus (and these limits can be made very close indeed). Thus the film recording can be made to yield precise information as to the value of the harmonic components of a wave-form, giving frequency, phase and amplitude: but the extraction of this information is not easy, and the film method totally fails when the spectrum is required to be instantly visible (as in following continuous changes of frequency or quality of a tone).

Methods using Film Recordings

Whilst various methods can be devised for the continuous graphical recording of electrical wave-forms, the commercial types of recordings on photographic film, or on transparent tape with an opaque coating (*e.g.,* the Philips-Miller recorder) have been brought to a very satisfactory degree of accuracy over the audio-frequency band, and are the only methods likely to be used at present. If it were desired to record high-frequency wave-forms for exact analysis a precise cathode-ray technique would appear the best method, or possibly the use of a Kerr cell.

There are three main methods of analysis of film recordings: optical diffraction methods, mechanical or graphical analysis, or electro-mechanical means.

In optics, the diffraction grating is, in its simplest form, a series of lines ruled very closely and regularly upon a transmitting or reflecting surface. When the ruled surface is illuminated by a beam of light, the spaces between the lines act each as a source of light, and the wave-fronts from each source combine beyond the grating to give alternative maxima or minima of light, according to the phases of the superimposed waves. The positions of the maxima are dependent upon the wavelength of the light, and thus the grating will give the spectrum of the incident light. The regularity of the spectrum given is dependent upon the accuracy of the grating, and much care is taken to ensure its absolute regularity. If, now, a monochromatic light source is used, but falls upon an irregular grating, the dispersion produced will be a function of the grating, and indicate the nature of the grating irregularity. It was stated by Brown[1] that the film record of a sound, when used as an optical grating, would give an immediate indication of the nature of the sound. The theory has been worked out in detail by Schouten,[2] who shows that for a film recorded on the variable intensity system, only the first order of the spectrum is present, and this spectrum consists of lines representing the tones present in the recording, spaced on a linear frequency scale, their intensities being proportional to the squares of the amplitudes of the individual tones. If the recording is variable amplitude, the diffraction pattern is very much more complicated (for a complete treatment, see Schouten, Physica, 7,2, February, 1940), but on the horizontal axis of the pattern, the spectrum of the sound is represented as with the variable intensity film.

This method gives a visible spectrum for any sound, even a single impulse, and is thus well adapted to indicate the nature of a waveform in a direct and rapid way: where considerable accuracy is required, there are evident optical limitations, more particularly in the measurement of the amplitudes of the individual components.

A more accurate analysis can be carried out by enlarging the film record (which must be amplitude recorded) to a suitable size, and analysing the curve obtained by one of the schedule methods for Fourier analysis[3] or by the use of a specially designed form of planimeter (the one usually employed is the Henrici Harmonic Analyser).[4] By either of these means, accurate information on the amplitude and phase of all harmonic components present can be obtained; but if many components are present, both methods are very laborious and slow.

There are two electro-mechanical methods which greatly decrease the work involved in the analysis of a curve. One fairly obvious method[5] is to take the strip of film on which the sound is recorded, or an enlarged copy, and fasten it around a circular wheel so that the ends abut with as little discontinuity as possible, and transform the curve back into electrical vibrations by an orthodox sound-head. This backward procedure will give a steady output of harmonics of the period of rotation of the circular film; if the length of film has been carefully chosen to correspond with the fundamental period of the recorded tone (or a multiple of this period), these harmonics will give the spectrum of the recorded tone. The electrical output can be analysed at leisure by any of the methods developed for the analysis of steady periodic wave-forms.

The most rapid method yet devised appears to be that of Montgomery,[6] who claims an analysing time of about 1½ minutes. This method assumes a periodic function, but since it employs only a single period, it gives a close approximation to the instantaneous frequency spectrum of a changing wave-form. The film recording (which

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[*] Post Office Research Station

[†] *Electronic Engineering*, Sept. 1942, p. 149.

may be variable density or variable amplitude) is combined in an optical system with a variable density film bearing the recording of a pure sine wave; by using a series of films bearing different frequencies, the successive harmonic components of the recorded sound can be obtained.

The recorded wave-form can be expressed as the following Fourier Series

$$f(x)=a_0+\sum_1^\infty a_n \cos nx+\sum_1^\infty b_n \sin nx$$

$$=c_0+\sum_1^\infty c_n \cos (nx-\theta_n)$$

where $a_n=c_n \cos \theta_n$

$b_n=c_n \sin \theta_n$

and to these series apply the well-known formulæ

$$a_n = \frac{1}{\pi}\int_0^{2\pi} f(x) \cos nx.\ dx.$$

$$b_n = \frac{1}{\pi}\int_0^{2\pi} f(x) \sin nx.\ dx.$$

$$c_n = \frac{1}{\pi}\int_0^{2\pi} f(x) \cos (nx - \theta_n) dx$$

$$\text{where} \int_0^{2\pi} f(x) \sin (nx - \theta_n)\ dx = 0$$

$$a_0 = c_0 = \frac{1}{2\pi}\int_0^{2\pi} f(x)\ dx$$

The optical transmission of the film recording of $f(x)$ is $A + f(x)$ (where A is large enough to make $A + f(x)$ always positive), and of the second film, $B_n[1 + M_n \cos (nx - \theta)]$ where M_n is somewhat less than 1, B_n is average transmission of screen, and θ gives position of second film relative to first (which is assumed to be a single period of $f(x)$ occupying a length of 2π units). When, by the optical system, the two films are superimposed the transmission is given by

$$T = \int_0^{2\pi} B_n[A+f(x)][1+M_n \cos (nx-\theta)]\ dx$$

Simplifying and substituting from equations above.

$T=2\pi B_n(A+c_0)+\pi B_n M_n c_n \cos (\theta - \theta_n)$

If now θ is varied from O to π, the difference in the values of T is $2\pi B_n M_n c_n \cos \theta_n = 2\pi B_n M_n a_n$; and similarly from $\theta = \pi/2$ to $\theta = 3\pi/2$, the difference in T is $2\pi B_n M_n c_n \sin \theta_n = 2\pi B_n M_n b_n$. Thus from a knowledge of the constants of the sine wave recording, the harmonic components of the recorded sound can be found, by using successive values of n. The amplitude of the nth harmonic is actually $c_n = \sqrt{a_n^2 + b_n^2}$; this can be found directly, for let T be a maximum at $\theta = \theta_n$ and a minimum at $\theta = \theta_n + \pi$; then the difference is $2\pi B_n M_n c_n$.

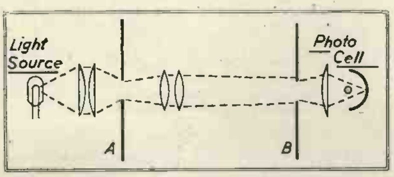

Fig. 1. Optical arrangement of Montgomery's method.

The optical system used to facilitate the superposition of the films is shown in Fig. 1. The film bearing the wave-form $f(x)$ which is to be analysed is placed at A, and an enlarged image at B is adjusted so that one period just fills a window of predetermined size. The successive films of cos nx are brought up and slid across behind this window: the variations of photo-cell current are amplified and recorded on a chart by means of a level recorder. The apparatus as described in Montgomery's article gives a fundamental frequency range of 65 c/s-310 c/s (using the customary recording speed for sound film), and deals with harmonics up to the 30th, thus the total range covered is extended to 9,300 c/s: the smallest harmonic amplitude measureable is about 2 per cent.

Methods not using recording

If useful results are to be obtained on short duration and changing wave-forms, speed of analysis is essential. There is an inevitable compromise involved here between speed and resolving power, since the more selective a frequency-discriminating device is made, the longer its build-up time. Those methods which use a single resonating element are obviously at a further disadvantage, since the frequency range must then be covered sufficiently slowly to allow the element to build up at any point where a component may be present. Thus the more rapid methods use a multiplicity of selective elements, the limit in this being reached when a diffracting grating is used. Other elements used are band-filters and vibrating reeds.

The band-filter method has been adopted by Freystedt,[7] who uses a series of 22 filters, 3 for each octave, to cover a range of 40 c/s to 5,500 c/s. The general arrangement is shown in Fig. 2; the outputs of the band-filters are connected successively to a cathode ray oscillograph so that the amplitude of the sound averaged over each band is shown as a line on the oscillograph. The time of build-up of a band-filter is approximately equal to $1/(f_2 - f_1)$ where f_2 and f_1 are the cut-off frequencies[8] so in this case for the lowest frequency filter the build-up time is of the order of 1/10 second. The time taken to cover the frequency range with the switching mechanism is thus conveniently made 1/10 second, as there is no point in making it faster. It is seen that any attempt to improve the resolution will result in proportionately slower response: in any case, there are practical limits to the number of band-filters which can be employed.

A very similar method[9] has been used to obtain statistical data on musical instruments and orchestras, finding the peak and average power in the bands covered: this is, however, hardly wave-form analysis in the strict use of the term.

Greater resolution at the expense of build-up time is obtained in the assembly of reeds which has been called an Acoustic Spectrometer.[10] In this instrument, 48 reeds per octave are used. The frequencies of the reeds are chosen to increase in geometric progression: their response (which is ap-

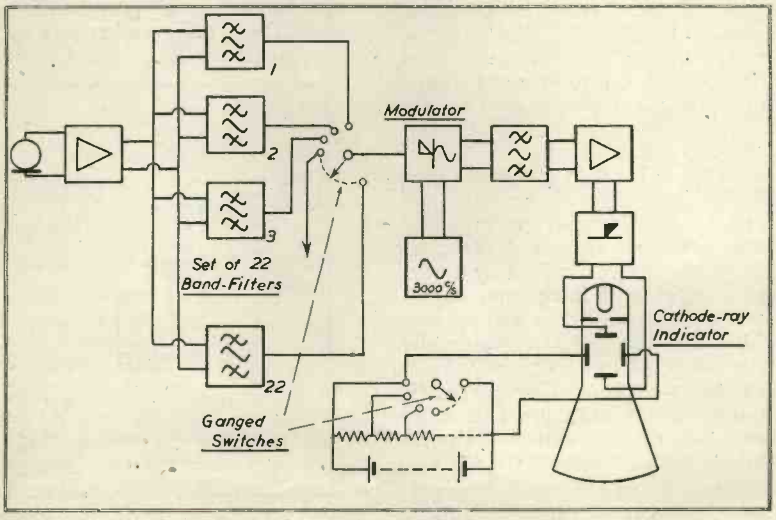

Fig. 2. Freystedt's band-pass filter method.

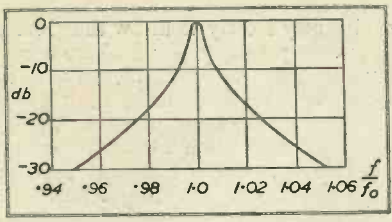

Fig. 3. Response of reeds in Hickman's apparatus.

proximately constant for all reeds) is shown in Fig. 3. This corresponds to a logarithmic decrement δ of .0072, giving a rate of decay of $f/8$ db. per second, where f is the resonant frequency of the reed concerned; thus a reed at 130 c/s (a low speech frequency) cannot follow changes quicker than some 16 db/sec.: this speed is sufficient for most speech and much music. Another limitation is that the ratio $\frac{\text{amplitude at resonance}}{\text{amplitude at zero frequency}}$ (which equals $\pi/2\delta$) is 46.7 db. and so the amplitude of a reed due to a component of its frequency cannot be greater than 46.7 db. above that of any lower frequency reed in response to the same tone. This low selectivity for frequencies some distance from the resonant point is, of course, inherent in any selector consisting of only one resonant element for each frequency: a tuned circuit before the reed would eliminate the trouble, which is, however, of little consequence in the normal uses of the instrument.

The wave-form to be analysed is normally a sound received by a microphone and amplified in the normal manner before being fed to the reeds. The reeds are driven so that their second mode of vibration (6.3 times the first) is eliminated (by driving on the node for this vibration); the third mode is 17 times the first, and those reeds with frequencies less than 1/17 of the top frequency to be analysed are specially shaped to remove this mode. The general arrangement of the reeds is shown in Fig. 4. The light falling upon the concave mirrors on the reed tips is focused on to a screen, and lines of length proportional to the reed amplitudes give a visible spectrum of the sound.

The extreme limit in speed has been reached by Meyer,[11] who performs the analysis by a diffraction grating. The general arrangement is shown in Fig. 5; the sound is received by a high quality condenser microphone, amplified and modulated up to a frequency at which a grating of reasonable size can be used. The frequency band chosen is 45-50 kc/s. (for an audio band of 0-5 kc/s) and cannot be much higher, since the absorption of the air is becoming quite high. There is also the problem of radiator and receiver; the sound source is a very thin duralumin ribbon, 5 μ thick by 6 mm. by 120 mm. in a field of 20,000 gauss, through which the output of the H.F. amplifier passes, whilst the receiver is a small condenser microphone (diaphragm 10 mm. diameter of 10 μ duralumin foil). The acoustic grating consists of 300 steel needles of 3.4 mm. diameter fastened between parallel iron plates 12 cm. apart, forming a concave grating 3 metres in length. The elementary theory of a diffraction grating is applied by Meyer to obtain the resolving power, which in terms of frequency discrimination is f/n where f is the frequency, n the number of grating elements. Thus in the case where f is about 45 Kc/s and $n = 300$ the resolving power is 150 c/s. It is shown by Meyer that a simple concave grating gives large aberrations, and a considerable amount of empirical modification to the grating curve is necessary to obtain good results. The final grating has a resolving power of 125 c/s. and a dispersion of 8 cm. per 1,000 c/s.

The output from the ribbon loudspeaker falling on the grating is reflected by the steel rods and the interference pattern thus produced is calculated so as to spread the component frequencies over a curve in front of the grating. The condenser microphone moves along this path, and the amplified output operates a recording oscillograph; the beam of light from the oscillograph is moved at right-angles to the recording direction in synchronism with the movement of the microphone. Thus a spectrum is obtained on a frequency base, and is recorded photographically; photographic considerations limit the speed to about 1/10 second. This is not the ultimate speed, which is given by the time required to build up the complete diffraction pattern, and is actually the difference in time taken by the extreme sound rays to reach the grating and reflect back to the microphone position. In accordance with the general rule that response time is inversely proportional to resolution, this time is somewhat under 1/100 second. Although giving sufficient resolution over most of the band, this method is not altogether satisfactory in that secondary maxima appear after each main component, the first of which has an amplitude 20 per cent. of the principal amplitude: it is evident this may be very confusing in a crowded spectrum.

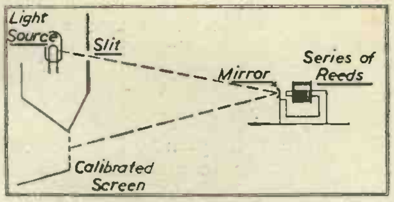

Fig. 4. Layout of Hickman's apparatus.

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3 See, for instance, Grover, *Bull. Bureau Standards* 9, 567 (1913) and Taylor, *Phys. Rev.* 6. 303, (1915).

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11 Meyer, *Jour. Acoust. Soc. America*, 7, 88 (1935). This is a translation by F. A. Firestone, of the original German.

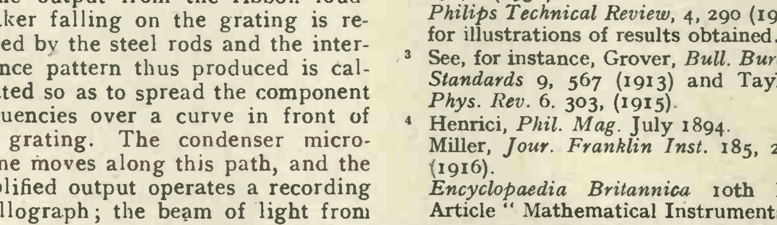


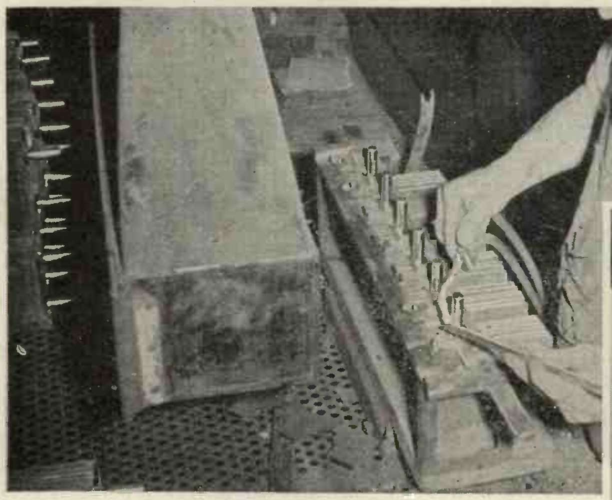
Fig. 5. Schematic diagram of Meyer's apparatus.

The Manufacture of Accumulators

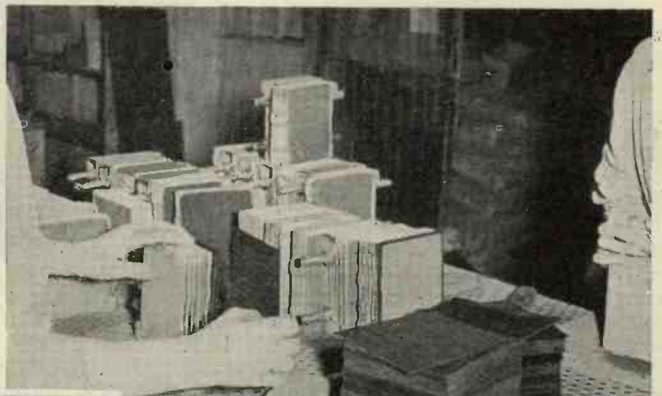
Stages in the manufacture of radio and car batteries illustrated by photographs taken at the works of the Edison Swan Electric Co., Ltd.

Casting lead grids in pairs for the battery plates.

Spreading paste of active material on the lead grids.



Joining batches of plates to form the battery and melting on the terminal post.



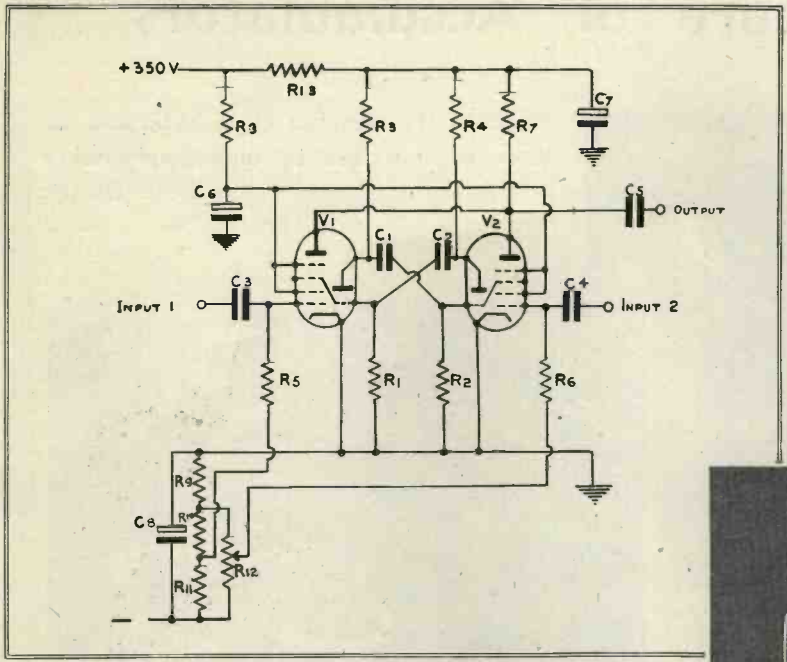
(*Above*) Fitting wood separators between plates. (*Left*) Sealing in tops of finished batteries with compound.

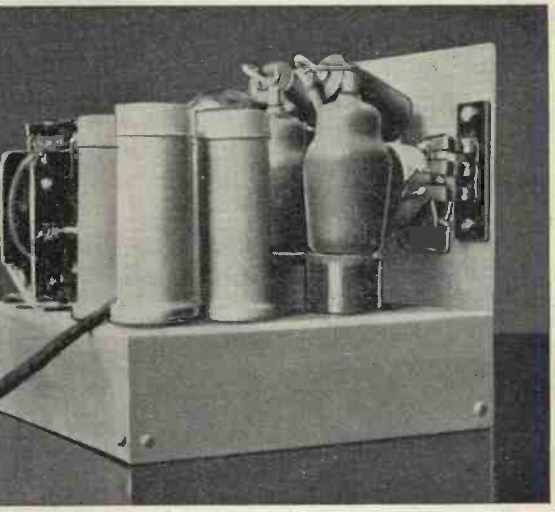
These photographs were taken from a short instructional film: "The Acid Test," distributed by Wallace Productions, Ltd. Copies of the film are available for Service instructional purposes and enquiries should be addressed to the Publicity Dept., The Edison Swan Electric Co., Ltd., 13a, King Street, Twickenham, Middlesex.

A Simple Electronic Switch

By A. W. RUSSELL

(Mullard Wireless Service Co.)





VALUES OF COMPONENTS

R_1 =	1 megohm	R_7 =	15,000 ohms	C_1 =	500 $\mu\mu$F
R_2 =	1 megohm	R_8 =	40,000 ohms	C_2 =	500 $\mu\mu$F
R_3 =	40,000 ohms	R_9 =	56 ohms	C_3 =	0.5 μF
R_4 =	40,000 ohms	R_{10} =	56 ohms	C_4 =	0.5 μF
R_5 =	1 megohm	R_{11} =	56 ohms	C_5 =	0.5 μF
R_6 =	1 megohm	R_{12} =	0.2 megohm	C_6 =	32 μF
		R_{13} =	1,500 ohms	C_7 =	16 μF
				C_8 =	50 μF

V_1 = V_2 = MULLARD ECH35

Fig. 5. Complete circuit diagram of the electronic switch using two triode hexodes. The photograph on the right shows the assembly on a chassis measuring 6½ inches square.

WITHIN the last few years, a number of electronic switch circuits have been devised with the object of extending the scope of the cathode-ray oscillograph, by making it possible to view simultaneously two or more phenomena when using a single beam tube. Each circuit had some particular merit, but in general the more successful required a considerable number of valves, while the more economical usually gave a bad "haze" between the two traces.

The principle underlying the electronic switch is very simple. The two voltages E_1 and E_2 to be studied are applied to the valves V_1 and V_2 which have a common anode load R_A. By some means, V_1 and V_2 are rendered alternately conducting at a frequency $2n$ cycles/sec. The output voltage E_o across R_A will then be represented by Fig. 2a, and when this is applied to the vertical deflector plates of a cathode-ray tube, the two voltages will appear as shown in Fig. 2b. If the steady anode currents I_{a1} and I_{a2} of the two valves had been the same, the two traces would have appeared on a common base line, so that variation of the steady anode currents provides a convenient way of shifting one trace relative to the other for identification, etc.

Another point is readily apparent from Fig. 2a. If the changeover from V_1 to V_2 is not instantaneous there will be a transition stage when both valves will draw current resulting in a sudden drop in output voltage E_o. These so called transition voltages are the cause of the background haze frequently associated with electronic switches.

It is most convenient to use a pair of multi-grid valves for V_1 and V_2, so that the switching may be effected by a suitable voltage applied to the auxiliary grid, thus ensuring that there is no interaction between signal and switching voltages. In order to reduce the transition time to a minimum, the auxiliary grid used for control should have a short base, and a sharp cut-off, while it is advantageous that it should lose control of the electron stream when it is more than a few volts positive with respect to cathode, as otherwise the waveform of the switching voltage will be superimposed on the output voltage. Consideration of these points leads to the choice of the hexode as the most suitable valve, since grid 3 is situated between two screening grids maintained at a positive potential of 100 volts, as shown in Fig. 3a. The I_a-V_{g3} characteristic of the hexode portion of a Mullard ECH35 triode-hexode is shown in Fig. 3b.

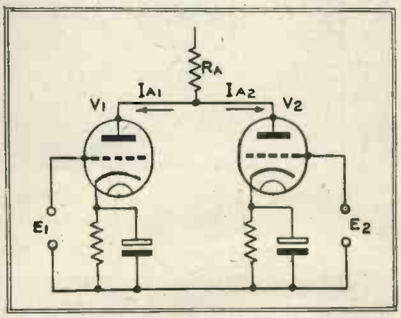


Fig. 1. Diagram illustrating the principle of the electronic switch.

The generation of a suitable switching voltage is the next consideration. It is fairly obvious from the foregoing that the ideal waveform would be square topped with "crossovers" as sharp as possible. Such a waveform, or a close approximation to it, is given by the well known multivibrator circuit of Fig. 4. The frequency of vibration n of such an oscillator is given by the formula $n=1/4RC$ where R is measured in megohms and C in microfarads. This circuit using triodes, following on the choice of a pair of hexodes for the switched valves, leads one irresistibly to attempt to satisfy all the conditions using a pair of triode-hexodes only. Since in the triode-hexode the grid of the triode portion is connected internally to grid 3 of the hexode, no external coupling is required and the whole circuit becomes very simple.

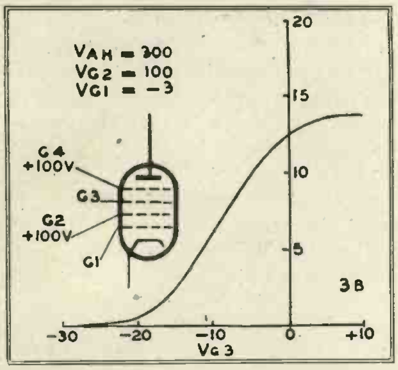


Fig. 3. Connexions and characteristics of ECH 35 triode hexode.

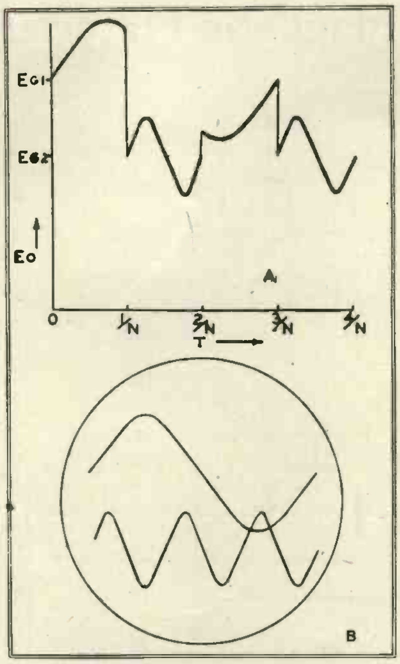


Fig. 2. (*a*) Wave form of output voltage across anode resistance of Fig. 1 and (*b*) the waveform on the screen of the tube.

Fig. 5 shows a practical circuit which has been evolved for use with the Mullard Oscillograph GM.3156. The frequency of oscillation of the multi-vibrator has been fixed at 500 cycles/sec. for this particular work, but can be varied by changing the values of C_1, C_2, R_1 and R_2. The choice of the switching frequency is governed by two considerations. When both input circuits are earthed, the output voltage will be a square wave and this must be handled without distortion by the oscillograph amplifier. If the frequency of the square wave be N cycles/sec., then in order to preserve the flatness of the tops of the waves, the amplifier must have a response flat to $N/10$ sinusoidal cycles/sec., while in order to handle the transient "crossovers," it must also be flat up to $10N$ sinusoidal cycles/sec. The amplifier of the GM.3156 is flat from 0.1 cycles/sec. to 10,000 cycles/sec. and can therefore handle square waves with frequencies from 1 cycle/sec. to 1,000 cycles/sec.

There is only one point in the final design which calls for special comment, and that is the method of obtaining bias for the hexodes and for shifting one trace relative to the other. This method was chosen for two reasons. In the first place it prevents cross-modulation of the two input voltages due to a common cathode impedance and in the second place it prevents unwanted coupling between triode and hexode portions of the valves.

Fig. 6a shows the wave form of the switching voltage generated by the multi-vibrator, while Figs. 6b and 6c show the electronic switch in operation: The two traces show the amplitude and phase relationships between the grid and anode voltages of a back-coupled 1,000 c/s. oscillator. Fig. 6b was obtained with a low value of grid leak only just sufficient to maintain oscillation, while in Fig. 6c it was increased to a much higher value, giving rise to distortion of the waveform.

The experimental model described, together with a power pack to permit operation direct from the 50 cycle mains, can be assembled on a box chassis measuring only 6½ in. by 6½ in.

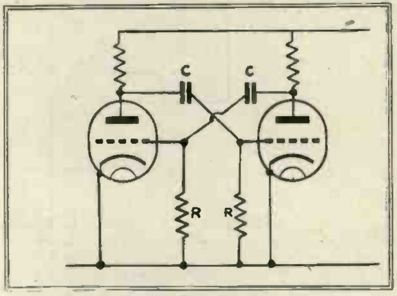
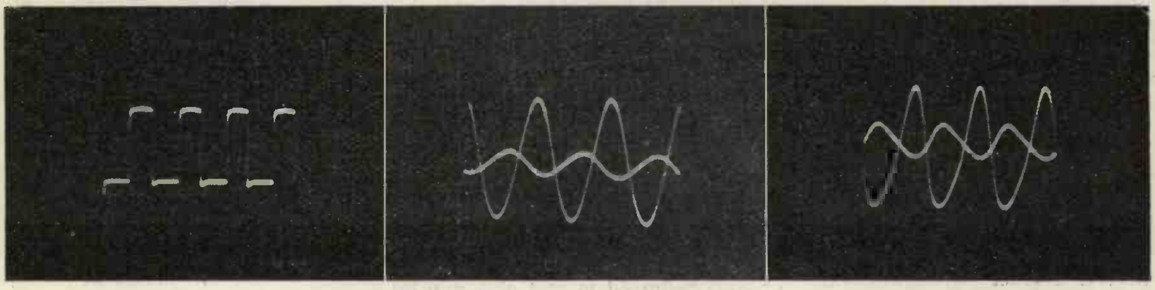


Fig. 4. Basic multivibrator circuit.

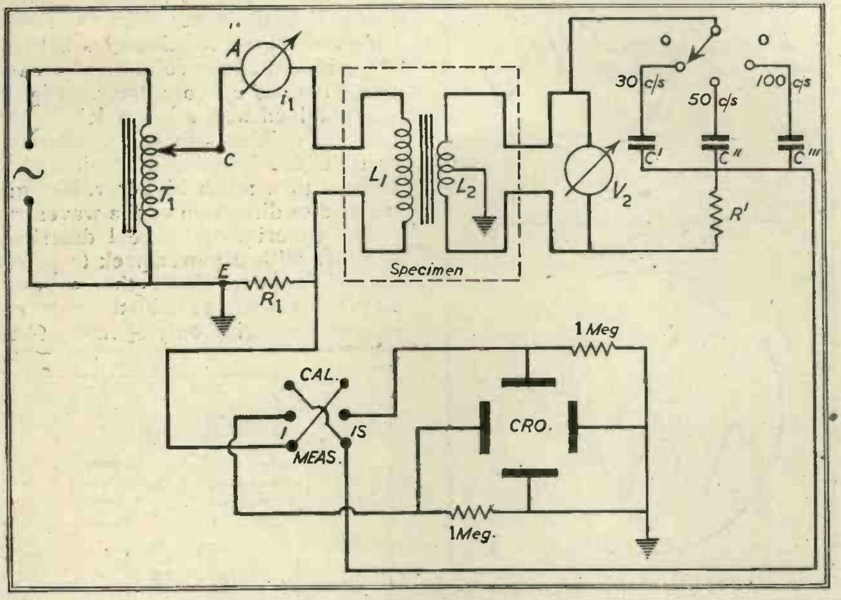
In conclusion, it may be helpful briefly to consider the best way of using the electronic switch. The oscillograms Figs. 6a and 6b were obtained when the electronic switch was followed by an amplifier having a gain of approximately 30, giving a sensitivity of approximately 300 mV. rms/cm deflection. Taking into account the amplication of the electronic switch, this gives an overall sensitivity of approximately 20 mV. rms/cm. The limitation on the amplification which can usefully follow the switch is set by the magnitude of the switching voltages and the squareness of the multivibrator waveform. With the switch described, quite good results can still be obtained with a gain of 100, but as a general rule, if the input voltages to the switch are less than 100 mV. it is best to use a stage of amplification in front of the hexodes rather than increase the gain following the switch.



a b c

Fig. 6. (*a*) Waveform of switching voltage (*b*) and (*c*) typical waveforms obtained from the switching circuit.

The Testing of Magnetic Materials by the C.R.O.



Complete testing circuit for magnetic materials.

IN the September issue of the *Journal of Scientific Instruments*,* K. Kreielsheimer, of the New Zealand Dept. of Scientific and Industrial Research describes a method of obtaining the B-H curves of magnetic materials using an electrostatic deflection C.R. tube. If a voltage $e_1 = E_m \sin \omega t$ is supplied to the primary of a transformer having no losses and no secondary load, the alternating current traversing the primary coil gives rise to an alternating field and flux practically in phase with this current. According to Lenz's law an e.m.f. will be induced by this changing flux in the primary and secondary winding in equal phase, but the primary e.m.f. is of opposite sign to the supply voltage. Hence, the primary e.m.f.

$E_1 = -E_m \sin \omega t = -10^{-8} N_1 q dB/dt$, where N_1 = number of primary turns and q the cross-section of the core. It follows:

$$B = -(E_m/N_1 q) \cos \omega t \times 10^8 \text{ and } E_2 = -E_m(N_2/N_1) \sin \omega t.$$

This simple analysis shows clearly that all voltages available on the transformer terminals are sine functions, whilst the induction B is a cosine function, *i.e.*, 90° out of phase with either primary or secondary e.m.f., and that the voltage amplitudes are proportional to B_{max}. It is then obvious that a correct hysteresis loop utilising the voltage drop across a small resistor in the primary circuit as proportional to H and the primary or secondary voltage as proportional to B can only be obtained if the voltage which is to represent the B-deflection has undergone a phase shift of $\frac{1}{2}\pi$.

We know from consideration of the B-H loop that B_{max} and H_{max} are corresponding points. Hence, the maximal deflection in the horizontal should coincide with the greatest vertical deflection. This co-phase condition, however, no longer holds near the centre of the loop where remanence and coercive force indicate a phase displacement between H and B. As B follows a cosine law, H or the magnetising current i_m in the primary must show a considerable distortion from the cosine shape. It is, therefore, not permissible to adjust the range of flux densities over which the specimen is to be investigated by means of a variable resistor in the primary. Because of the distorted form of the primary current the voltage drop across this resistor would cause the input voltage to the transformer to be no longer a sine curve. It follows that the resistance inserted in the primary circuit to create a voltage drop proportional to H must be small compared with the input impedance of the specimen transformer. Any variation of B_{max} required is best obtained by a continuously variable supply transformer (*e.g.*, a Variac).

The importance of the $\frac{1}{2}\pi$ phase shift of the output voltage has been recognised by Johnson[1] and Cosens[2] who load the secondary coil of a transformer with a condenser and a resistance in series. By making $R \gtrless 1/\omega C$ utilising the voltage across C for vertical deflection, a very close approach to a 90° phase shift is obtainable, but only by sacrificing most of the available output voltage. It might be pointed out that the above considerations are not influenced by the actual presence of losses. The primary current i_1 is composed of the wattless component, i_m, the magnetising current, and the loss component, *i.e.*, which determines the area of the hysteresis loop. With a negligible secondary load, the terminal voltage of the secondary coil equals E_2 which when advanced 90° will coincide with the phase of i_m.

This necessary rotation of the voltage vector by $\frac{1}{2}\pi$ can be achieved simply and accurately by using a phase-adjusting device, previously described.[3] Although in this arrangement the secondary is loaded by a resistor and condenser in series the fundamental difference from Johnson and from Cosens lies in the fact that R is adjusted to be equal to $1/\omega C$. As has been shown in the publication referred to, the voltage between the midpoint or artificial midpoint, of the secondary winding and the common point of R and C is then 90° out of phase to E_2 and of half its value. Fig. 1 gives the complete circuit diagram proposed for the determination of the iron losses. As indicated, various R-C combinations can easily be provided to enable the investigation to be carried out at various frequencies f, for which $R = 1/\omega C$. When selecting the values of the R-C combination it should be kept in mind that the load created should be negligible.

The method used was found to be very sensitive and results obtained appear to be accurate within 2 per cent. Routine tests could easily be arranged as a comparison method with a standard specimen excited by the same magnetising current, whilst the hysteresis loops could be superimposed on the oscillograph screen by means of an electronic switch.

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* p. 137. Vol. 19. 1942.

[1] JOHNSON, J. B. *Bell System Tech. J.* 8, p. 286 (1929).

[2] COSENS, C. R. *Wireless Eng.* 12, p. 190 (1935).

[3] KREIELSHEIMER, K. *Wireless Eng.* 17, p. 439 (1940).

The Cathode Follower

By C. E. LOCKHART

THE "cathode follower" type of circuit has the distinction of being one of the most versatile circuits used by the Radio Engineer. The circuit in its most elementary form is shown in Fig. 1, and derives its name from the fact that an input signal that makes the grid go positive with respect to H.T. negative will also make the cathode go positive with respect to H.T. negative. In addition if the resistance R_c is made sufficiently high the output voltage is practically equal to the input voltage, and the cathode may therefore be said to follow the grid as far as voltage fluctuations are concerned.

The main advantages of the "Cathode Follower" type of circuit may be summarised as :—

1. High input impedance.
2. Low output impedance.
3. High anode impedance.
4. Freedom from distortion.

The circuit shown in Fig. 1 representing the most elementary form of negative feed-back, has been treated on a number of occasions,[1] and [2] but as far as the author knows a generalised treatment of the "Cathode Follower" circuit has not been published before.

Owing to the complexity of the generalised treatment it is proposed to separate the problem into two stages; the first part limits the problem to low and medium frequencies where inter-electrode capacity displacement currents are negligible compared with the electron currents and electron inertia effects can be neglected. The second part treats the problem in a more generalised manner essential at the higher frequencies.

General Relations

If we assume linear valve characteristics we can express the anode current of a triode by

$$I_a = \frac{V_a + \mu V_g}{R_a} \quad \ldots \quad (1)$$

where V_a and V_g are the static potentials applied between the anode and cathode and grid and cathode respectively. I_a is the steady anode current, and R_a is (dI_a/dV_a) or the Anode A.C. resistance of the triode. (For a list of symbols *see* Data Sheet).

For an input signal of "E_1" volts (Fig. 2a) the signal "E_g" applied between grid and cathode is $(E_1 - i_aZ)$ and the change in anode-cathode volts is $(-i_aZ)$ where i_a is the resulting change in anode current and Z is the impedance in the cathode circuit.

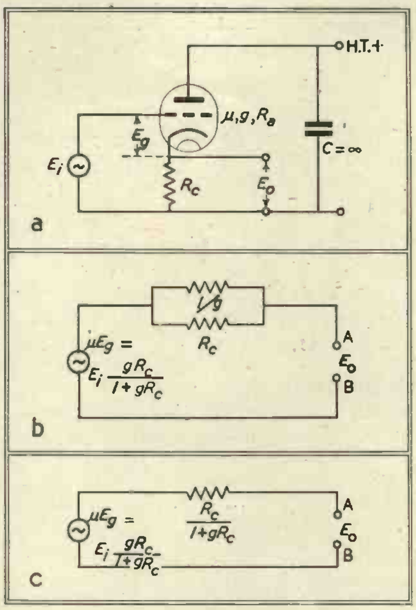


Fig. 1. (a) Simplest form of Cathode Follower circuit (b) & (c) Its equivalent circuits (for $\mu >> 1$) from which the voltages E_0 across any load connected between terminals A & B can be calculated. The capacity C is assumed to be an infinite by-pass condenser.

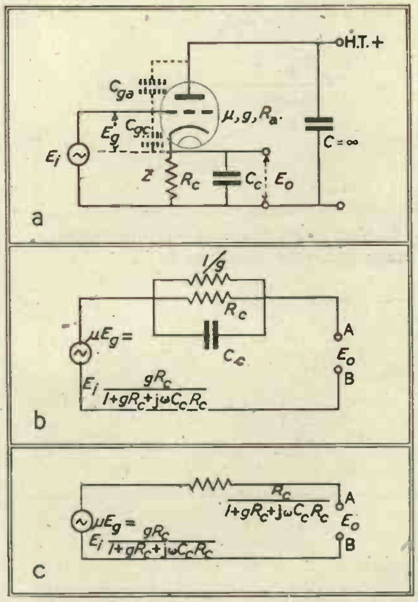


Fig. 2. (a) Cathode Follower circuit with stray capacity C_0 across the cathode resistor R_C. (b) & (c) Equivalent circuit when $\mu >> 1$ and when inter-electrode capacities are neglected.

Therefore

$$I_a + i_a = \frac{(V_a - i_aZ) + \mu(V_g + E_1 - i_aZ)}{R_a} \quad (2)$$

and subtracting (1) from (2)

$$i_a = \frac{\mu E_1}{R_a + Z(1 + \mu)}$$

$$= \frac{gE_1}{1 + gZ(1 + 1/\mu} \quad (3)$$

The output voltage E_o is (i_aZ) and the stage gain E_o/E_1 is given by

$$\text{Voltage Gain} = \frac{gZ}{1 + gZ(1 + 1/\mu)} \quad (4)$$

The grid-cathode voltage $E_g = E_1 - E_o$ or from (1) and (2)

$$i_a = \frac{-i_aZ + \mu E_g}{R_a} = \frac{\mu E_g}{R_a + Z} \quad (5)$$

therefore

$$E_g = E_1 \frac{1 + {}_gZ/\mu}{1 - {}_gZ(1 + 1/\mu)} \quad (6)$$

The output impedance of the device is obtained by calculating the current taken from a voltage source E_o (Fig. 2a) when E_1 is zero. This current "i_o" is

$$i_o = \frac{E_o}{Z} + \frac{E_o + \mu E_o}{R_a}$$

Therefore the admittance $A_o = i_o/E_o$ is

$$A_o = \frac{1}{Z} + g\left(\frac{1 + \mu}{\mu}\right) \quad (7)$$

and the output impedance is approximately equal to the impedance Z in parallel with a resistance $1/g$ when $\mu \gg 1$. The equivalent circuits are shown in Figs. 1b, 2b, and c.

Due to the negative feed-back produced by the cathode load the anode A.C. resistance for voltage fluctuations between anode and H.T. negative is increased to R_a'. From (2)

$$i_a = \frac{v_a - i_aZ(1 + \mu)}{R_a}$$

therefore

$$R_a' = \frac{v_a}{i_a} = R_a[1 + gZ(1 + 1/\mu)] \quad (8)$$

Cathode Load

The most usual load employed in the cathode is a resistance "R_c" (Fig. 2a) and at low frequencies Z in the above expressions may be replaced by "R_c" (*see* Fig. 1). It is, however, impossible to eliminate a capacity shunt across the cathode load, due to say the anode-cathode capacity of the valve, stray wiring capacities and the capacity of any following stage. As the frequency of the input signal is increased so the effect of "C_c" becomes increasingly important. The impedance of R_c and C_c in parallel is given by

$$Z = \frac{1}{j\omega C_c}\left(\frac{R_c}{R_c + 1/j\omega C_c}\right) = \frac{R_c}{1 + j\omega R_c C_c} = \frac{R_c}{1 + jp_o} \qquad (9)$$

if we let $\omega_o = 1/R_c C_c$ then $\omega R_c C_c = f/f_o = p_o$.

The value of Z given by (9) can now be inserted in the expression previously obtained.

Anode-Current

Inserting the value of Z from (9) into (3) we get

$$i_a = E_1 \frac{g(1 + jp_o)}{(1 + gR_c + R_c/R_a + jp_o)}$$

rationalising

$$i_a = E_1.g \frac{(1 + gR_c + R_c/R_a + p_o^2) + jp_o(gR_c + R_c/R_a)}{(1 + gR_c + R_c/R_a) + p_o^2} \quad \ldots \quad \ldots \quad \ldots \quad (10)$$

if we now express i_a in the form $E_1|gK| \underline{/\psi}$

$$i_a = E_1.g \frac{\sqrt{(1 + gR_c + R_c/R_a + p_o^2) + p_o^2(gR_c + R_c/R_a)^2}}{(1 + gR_c + R_c/R_a)^2 + p_o^2} \underline{/\psi} \quad \ldots \quad \ldots \quad \ldots \quad (11)$$

where

$$\psi - \tan^{-1}\left[\frac{p_o(gR_c + R_c/R_a)}{1 + gR_c + R_c/R_a + p_o^2}\right] \qquad (12)$$

It should be noted that the term

$$(1 + gR_c + R_c/R_a) = 1 + gR_c(1 + 1/\mu)$$

so that the relation $R_c/R_a \ll gR_c$ is synonymous with $\mu \gg 1$.

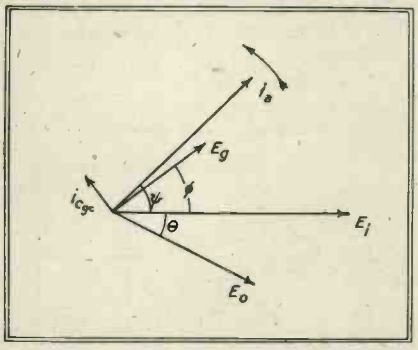


Fig. 4. Vector relations of cathode follower stage at higher values of $2\pi f C_c R_c = p_o$

Grid-Cathode Voltage

From (6) and (9)

$$E_g = E_1 \frac{(1 + R_c/R_a) + jp_o}{(1 + R_c/R_a + gR_c) + jp_o} = \qquad (13)$$

$$= E_1 \frac{\sqrt{[1 + R_c R_a)(1 + R_c/R_a + gR_c) + p_o^2]^2 + (p_o g R_c)^2}}{(1 + R_c/R_a + gR_c)^2 + p_o^2} \underline{/\phi} \quad \ldots \quad \ldots \quad \ldots \quad (14)$$

$$\text{where } \phi = \tan^{-1}\left[\frac{p_o g R_c}{(1 + R_c/R_a)(1 + R_c/R_a + gR_c) + p_o^2}\right] \quad \ldots \quad \ldots \quad \ldots \quad (15)$$

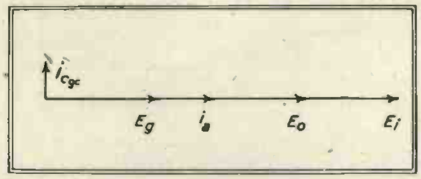


Fig. 3. Vector diagram of cathode follower stage for $2\pi f C_c R_c = p_o \rightarrow 0$

Output Voltage

$$E_o = E_1 \frac{gR_c}{(1 + R_c/R_a + gR_c) + jp_o} \qquad (16)$$

rationalising

$$E_o = E_1 gR_c \frac{[1 + R_c/R_a + gR_c) - jp_o]}{(1 + R_c/R_a + gR_c)^2 + p_o^2} \qquad (17)$$

$$= E_1 \frac{gR_c}{\sqrt{(1 + R_c/R_a + gR_c)^2 + p_o^2}} \underline{/\theta} \qquad (18)$$

where

$$\theta = \tan^{-1} - \frac{p_o}{(1 + R_c/R_a + gR_c)} \qquad (19)$$

when $\mu \gg 1$

$$E_o = E_1 \frac{gR_c}{1 + gR_c} \sqrt{\frac{1}{1 + [(p_o/(1 + gR_c)]^2}} \underline{/\theta} \qquad (20)$$

$$\text{where} \quad \theta = \tan^{-1} - \left(\frac{p_o}{1 + gR_c}\right) \qquad (21)$$

The phase angle θ is plotted against p_o for different values of gR_c on Data Sheet No. 41.

Voltage Gain

As the voltage gain is given by the ratio E_o/E_1 all the expressions for E_c in the above paragraph become expressions of voltage gain when divided by E_1.

Relative Gain

The term "relative gain" M is used to express the ratio of the gain at a frequency f to the gain at a very low frequency, *i.e.*, $p_o \rightarrow O$.

The gain at very low frequencies is given from (15) by

$$\text{gain at } p_c = 0 = \frac{gR_c}{1 + R_c/R_a + gR_c} \underline{/0^\circ} \qquad (22)$$

so that the relative gain M in db. is :

$$M_{\text{in db}} = 20 \log \sqrt{\frac{1}{1 + \left(\frac{p_o}{1 + R/R_a + gR}\right)^2}} \qquad (23)$$

or when $R/R_a \ll 1$ or $\mu \gg 1$

$$M_{\text{in db}} = 20 \log_{10} \sqrt{\frac{1}{1 + [p_o/(1 + gR_c)]^2}} \qquad (24)$$

This expression has been plotted on Data Sheet No. 39 for values of p_o up to 10 and gR_c products up to 20.

Time Delay

When considering video amplifiers in television we are concerned not only in the gain response characteristic, but also in the linearity of the phase angle θ versus frequency in order that signals of different frequencies shall take an equal time to pass through the amplifier (*see* Data Sheets I-III[3] and XXIII-XXV[4]).

The time difference between the output voltage vector and input signal voltage vector reaching their maximum amplitudes for a signal of frequency f is given by

$$t = \frac{\theta}{\omega} = \frac{\theta}{2\pi f} \qquad (25)$$

In the case of the cathode follower circuit Fig. 2. θ is always negative and therefore t is negative. The output signal is thus time delayed with respect to the input signal. Also as $\theta = 0$ when $f = 0$ the value of t from (25) will give the time delay of the output voltage vector of frequency f relative to that of the output voltage vector of very low frequency when the input signals of these two frequencies start simultaneously.

Now from (19) and (25)

$$t_1 = - \frac{\tan^{-1}\left(\frac{p_o}{1 + R_c/R_a + gR_c}\right)}{2\pi f} \text{ secs.} \qquad (26)$$

In order to enable a generalised graphical representation to be made the expression (27) must be rewritten in the form

$$f_o t_1 = -\frac{\tan^{-1}\left(\dfrac{p_o}{1+R_c/R_a+gR_c}\right)}{2\pi p_o} \text{ secs.} \quad (28)$$

which simplifies to

$$f_o t_1 = -\frac{\tan^{-1}\left(\dfrac{p_o}{1+gR_c}\right)}{2\pi p_o} \text{ secs.} \quad (29)$$

when $\mu \gg 1$. Expression (29) has been plotted on Data Sheet No. 40.

Vector Relations

The problem is better visualised by a graphical illustration of the vector relations expressing the equations derived above. Taking first the case of very low frequencies or $p_o = \omega CR \approx 0$ the vector relations for Fig. 2a are illustrated in Fig. 3.

Here we have for the reference vector the input signal E_1; the anode current i_a and the output voltage E_o are in phase with E_1 as both ψ and θ are zero. The angle ϕ is also zero so that $E_g = (E_1 - E_o)$ is in phase with E_1.

If we now take the case of a considerably higher frequency input signal, while still maintaining E_1 as the reference vector (*see* Fig. 4), then the anode current vector will now lead the input signal E_1 by the angle ψ given by equation (12). The output voltage vector E_o will, however, lag behind E_1 by an angle θ as Z is capacitive. The vector for E_g, being the difference between the vectors E_1 and E_o will lead E_1 by an angle ϕ.

A very significant feature which becomes at once apparent, is the possible rapid rise in the grid-cathode voltage E_g when E_o is no longer in phase with E_1. Equation (14) which simplifies to:

$$E_g = E_1 \frac{\sqrt{[(1+gR_c)+p_o^2]^2+(p_o gR_c)^2}}{(1+gR_c)^2+p_o^2} \quad (30)$$

when $R_c/R_a \ll 1$ and $\mu \gg 1$ must therefore be used to check that the signal E_g does not drive the valve into grid current or off its characteristic at the highest frequency to be employed.

Input Admittance

In the relations so far derived the capacity current through the cathode impedance Z due to the grid to cathode capacity C_{gc} (*see* Fig. 2) has been assumed negligible compared with the anode current. We can proceed to determine the input admittance of the stage with the above assumptions and provided the electron inertia effects are small enough to ensure negligible grid loading.

The Input Admittance A_1 is given by the ratio of the current taken from the source of the signal E_1 to the voltage E_1. As the grid loading is assumed negligible this current is equal to the current through the capacities C_{ga} and C_{gc}, therefore

$$A = \frac{i_{Cgc} + i_{Cga}}{E_1} \quad (31)$$

and

$$i_{ga} = E_1 (j\omega C_{ga})$$

$$i_{Cgc} = E_g (j\omega C_{gc}) \quad \ldots \quad (32)$$

From equations (31) (32) and (13) we have

$$A_1 = j\omega C_{gc} \frac{(1+R_c/R_a+jp_o)}{(1+gR_c+R_c/R_a)+jp_o} + j\omega C_{ga} \quad (33)$$

Rationalising we have

$$A_1 = j\omega C_{ga} + \omega C_{gc} \frac{-p_o gR_c + j[(1+R_c/R_a)(1+gR_c+R_c/R_a)+p_o^2]}{(1+gR_c+R_c/R_a)^2+p_o^2} \quad (34)$$

From which the input resistance R_1 is

$$R_1 = -\frac{(1+gR_c+R_c/R_a)^2+p_o^2}{\omega . C_{gc} . p_o . gR_c} \quad (35)$$

$$\approx -\frac{(1+gR_c)^2+p_o^2}{\omega . C_{gc} . p_o . gR_c} \text{ when } \mu \gg 1 \quad (36)$$

and the input capacity is

$$C_1 = C_{ga} + C_{gc} \frac{(1+R_c/R_a)(1+gR_c+R_c/R_a)+p_o^2}{(1+gR_c+R_c/R_a)^2+p_o^2} \quad (37)$$

$$\approx C_{ga} - C_{gc} \frac{1+gR_c+p_o^2}{(1+gR_c)^2+p_o^2} \quad (38)$$

when $R_c/R_a \ll 1$ and $\mu \gg 1$

$$\text{and} \quad C_1 \approx C_{ga} + \frac{C_{gc}}{1+gR_c} \quad (39)$$

when in addition p_o is small compared to unity.[5]

From the complete solution given in Part II it will be seen that the above equations are accurate at low values of p_o provided $gR_c \gg C_{gc}/C_c\ (1+R_c/R_a)$ and at higher values of p_o if in addition $C_{gc}/C_c \ll 1$.

The interesting facts derived from the above expressions are that the grid-cathode capacity C_{gc} is reduced by the factor $(1 + gR_c)$ and that the input resistance of a cathode-follower circuit as shown in Fig. 2 is negative. The latter fact could have been predicted from the vector diagram Fig. 4, as the current i_{Cgc} must be represented by a vector leading the vector E_g by 90°. If this current vector is now resolved into two components one in phase-quadrature (imaginary component) and the other in phase (real component) with the vector E_1, it will be seen that the real component is negative in sign.

If we consider the case of a normal amplifier with a resistive load R_c in the anode circuit, then in the case of a valve with an anode A.C. resistance R_a large compared with R_c, the voltage amplification is given by:

$$\text{Gain} = gR_c$$

Using this gain as a reference figure, we can consider that the effect of connecting the resistance in the cathode circuit (Fig. 1) is to reduce the gain by $(1 + gR_c)$ and to reduce the grid-cathode capacity by the same ratio.

Output Impedance

From Eq. (7) we have:

$$Z_o = \frac{1}{A_o} = \frac{Z}{1+gZ(1+1/\mu)}$$

$$= \frac{R_c}{1+gR_c(1+1/\mu)+jp_o} \quad (40)$$

$$= \frac{R_c}{\sqrt{[1+gR_c(1+1/\mu)]^2+p_o^2}} \underline{/\gamma} \quad (41)$$

$$\text{where } \gamma = \tan^{1}\left[-\frac{p_o}{1+gR_c(1+1/\mu)}\right] \quad (42)$$

With high slope valves gR_c may be made large compared with unity, and then at low frequencies expression (41) reduces to the output impedance, being $1/g$. Thus with a mutual conductance of 10 mA/V. an output resistance of only 100 ohms is obtained.

Anode A.C. Resistance

The anode A.C. resistance R_a' of the cathode follower is by (8):

$$R_a' = R_a\left[1 + \frac{gR_c}{1+jp_o}(1+1/\mu)\right] \quad (43)$$

For any voltage fluctuation v_a between anode and H.T. negative, the fraction of v_a developed across R_c is given by:

$$\frac{\text{A.C. volts across output}}{\text{A.C. volts across H.T. line}}$$

$$= v_a R_c/R_a \frac{1}{1+R_c/R_a+gR_c(1+1/\mu)+jp_o} \quad (44)$$

with an absolute value of

$$v_a \frac{R_c}{R_a} \frac{1}{\sqrt{[1+R_c/R_a+gR_c(1+1/\mu)]^2+p_o^2}} \quad (45)$$

When p_o is very small, such as at low frequencies, but with $\mu \gg 1$ and $gR_c \gg 1$, expression (44) simplifies to:

$$\frac{\text{A.C. volts across output}}{\text{A.C. volts across H.T. line}} = \frac{v_a}{\mu}$$

The complete solution of the cathode follower circuit together with general circuit considerations, will be given in the next series of Data Sheets.

DATA SHEETS XXXIX, XL, AND XLI

The Performance of the Cathode Follower Circuit

LIST OF SYMBOLS

A_i = Input Admittance of cathode follower.
A_o = Output Admittance of cathode follower.
C_{ac}, C_{ga}, C_{gc} = Anode-cathode, grid-anode, and grid-cathode inter-electrode capacities, respectively.
C_c = Total effective capacity across cathode load resistance R_c.
C_i = Input Capacity.
E_i = Input signal voltage between grid and negative H.T.
E_o = Output signal voltage.
E_g = Signal voltage developed between grid and cathode.

The above voltages are assumed to be of sinusoidal form and must all be expressed on the same basis, i.e., in R.M.S. or Peak volts. The instantaneous value of the voltage is of the general form $e = E \sin \omega t = E \sin 2\pi ft$.

f = $\omega/2\pi$ = Frequency of applied signal.
$f_o = \dfrac{1}{2\pi\ C_c R_c}$, i.e., frequency in c/s at which $\omega_0\ C_c\ R_c = 2\pi f_0\ C_c\ R_c = 1$.
g = Mutual conductance of a triode = di_a/dv_g.
i_a = Signal component of the anode current.
M = Relative amplification in db. = $20 \log_{10} \dfrac{E_o}{E_i}$.
p_o = $\omega\ C_c\ R_c = f/f_o$.
R_a = Anode A.C. resistance of a triode = di_a/dv_a.
R'_a = Anode A.C. resistance of cathode follower.
R_c = Resistance of cathode load in ohms.
R_i = Input resistance of cathode follower.
t = Time in seconds.
$-t_1$ = Time delay of frequency f relative to zero frequency.
$= \dfrac{\text{Phase angle in radians}}{2\,\pi f} = \dfrac{\text{Phase angle in degrees}}{360.f}$
v_a = A.C. voltage applied between anode and H.T. negative.
Z = Impedance of cathode load circuit.
Z_i = Input impedance of cathode follower.
Z_o = Output impedance of cathode follower.
μ = Amplification factor = gR_a.
ψ = Phase angle of i_a relative to E_i.
ϕ = Phase angle of E_g relative to E_i.
θ = Phase angle of E_o relative to E_i.
γ = Phase angle of output impedance of cathode follower.

THE use of these Data Sheets is best illustrated by working an example. Suppose it is desired to investigate the performance as a cathode follower at video frequencies of a triode-connected pentode having a mutual conductance of $g = 5$ mA/V, $\mu = 80$, and a total stray capacity across the load of 70 $\mu\mu$F.

Let us first calculate the performance with a cathode load of 2,000 ohms.

(1) The value of $C_c R_c = 0.14 \times 10^{-6}$, and therefore

$$f_o = \frac{10^6}{2\pi \times 0.14} = 1.14 \text{ Mc/s.}$$

Also we have: $gR_c = 10$, and $R_a = \mu/g = 16{,}000$ ohms.
Therefore $R_c/R_a = 1/8$.
The gain for $p_o = 0$ is $10/(1+10) = 0.91$.

(2) *Relative Gain.*

With the above constants and the $gR_c = 10$ curve of Data Sheet 39, we can now plot the curve of relative gain versus frequency. For this purpose we first express the applied frequency in terms of p_o $(=f/f_o)$. In the attached table the values of f have been tabulated in the first column, while $p_o(=f/f_o)$ is given in the second column.

From the values of p_o it is now possible to read the relative gain M from Data Sheet 39. The values so obtained are given in the third column of the table.

(3) *Phase Angle.*

In a similar manner the phase angle of the output voltage E_o relative to the input voltage E_i may be read off Data Sheet 41.

The values so obtained are given in column 4 of the table.

(4) *Values of Time Delay.*

To obtain the values of the Time Delay we can either use the expression:

$$t_1 = \frac{\theta^\circ}{360^\circ.f} \text{ secs.}$$

where θ is given its appropriate negative sign, or read off $(f_o t_1)$ directly from Data Sheet 40 and then obtain t_1 by dividing by $f_o = 140{,}000$ c/s. The values of $f_o t_1$ have been tabulated in the fifth column and the time delay $(-t_1)$ expressed in microseconds in the last column. For very low frequencies (*i.e.*, $p_o \to 0$).

$$f_o t_1 = \frac{1}{2\pi(1+gR_c)} \text{ secs.}$$

and
$$t_1 = -\frac{C_c R_c}{1+gR_c} \text{ secs.}$$

The calculation of time delay for the case of a modulated carrier will be dealt with in part II.

TABLE

Applied Frequency f in Mc/s.	$p^\circ = f/f_o$	Relative Gain M in db.	Phase Angle θ°	$f_o\ t_1$	Time Delay in μS $(-t_1)$.
0	0	0	0	0.0145	0.0127
1.0	0.88	−0.06	4.5	0.0144	0.0126
2.0	1.76	−0.11	9.0	0.0143	0.0125
3.0	2.64	−0.25	13.4	0.0141	0.01235
4.0	3.52	−0.42	17.7	0.014	0.01225
5.0	4.4	−0.65	21.8	0.0138	0.0121
6.0	5.28	−0.9	25.5	0.0136	0.0119
7.0	6.15	−1.18	29.2	0.0135	0.0118

RELATIVE GAIN OF CATHODE FOLLOWER STAGE

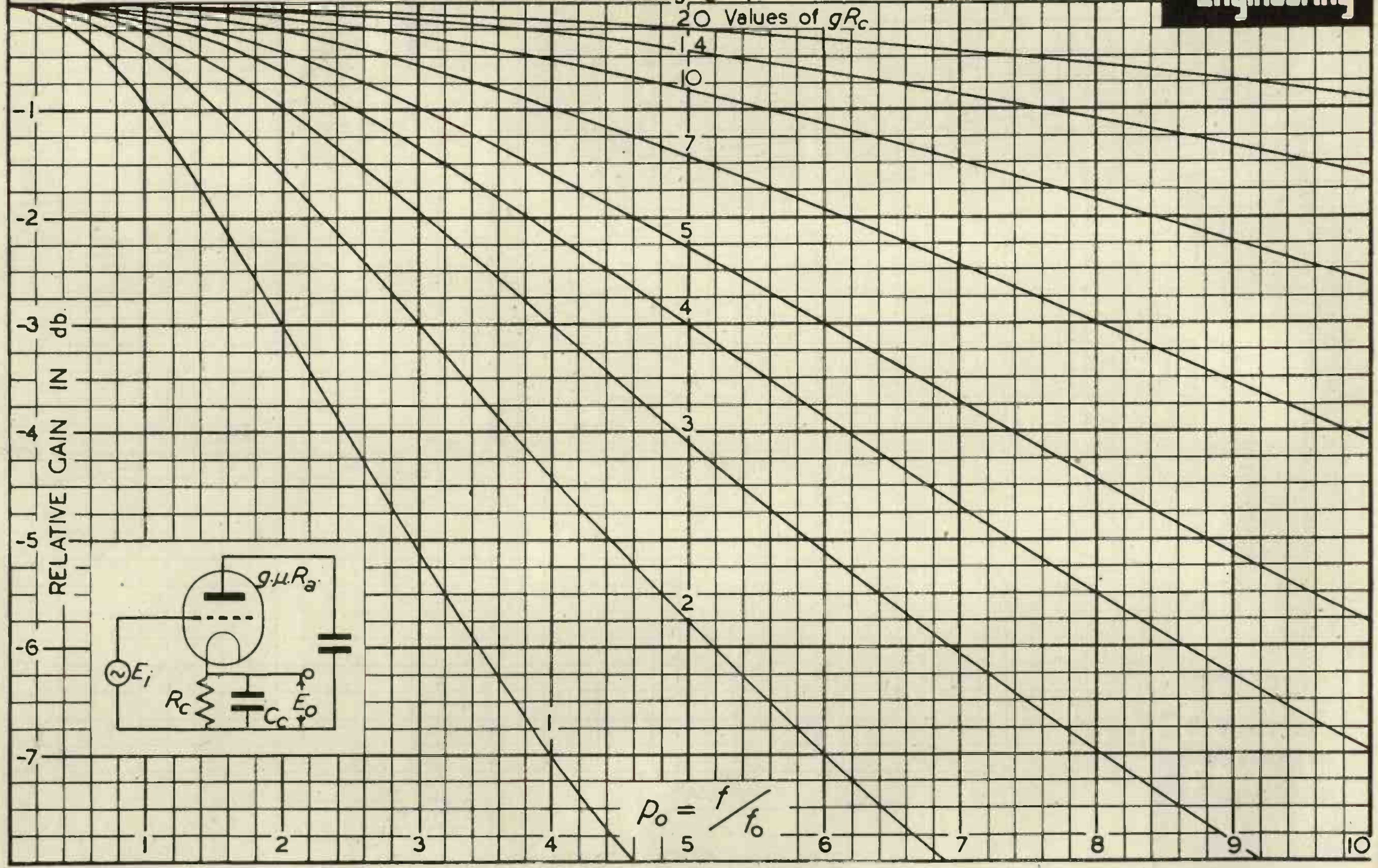
Relative Gain in db. $= 20 \log_{10} \frac{|E_o|}{|E_i|} = 20 \log_{10} \sqrt{\frac{1}{1+\left(\frac{p_o}{1+gR_c}\right)^2}}$

$f_o = \frac{1}{2\pi C_c R_c}$

$\mu >> 1$

DATA SHEET
No. 39

Electronic Engineering



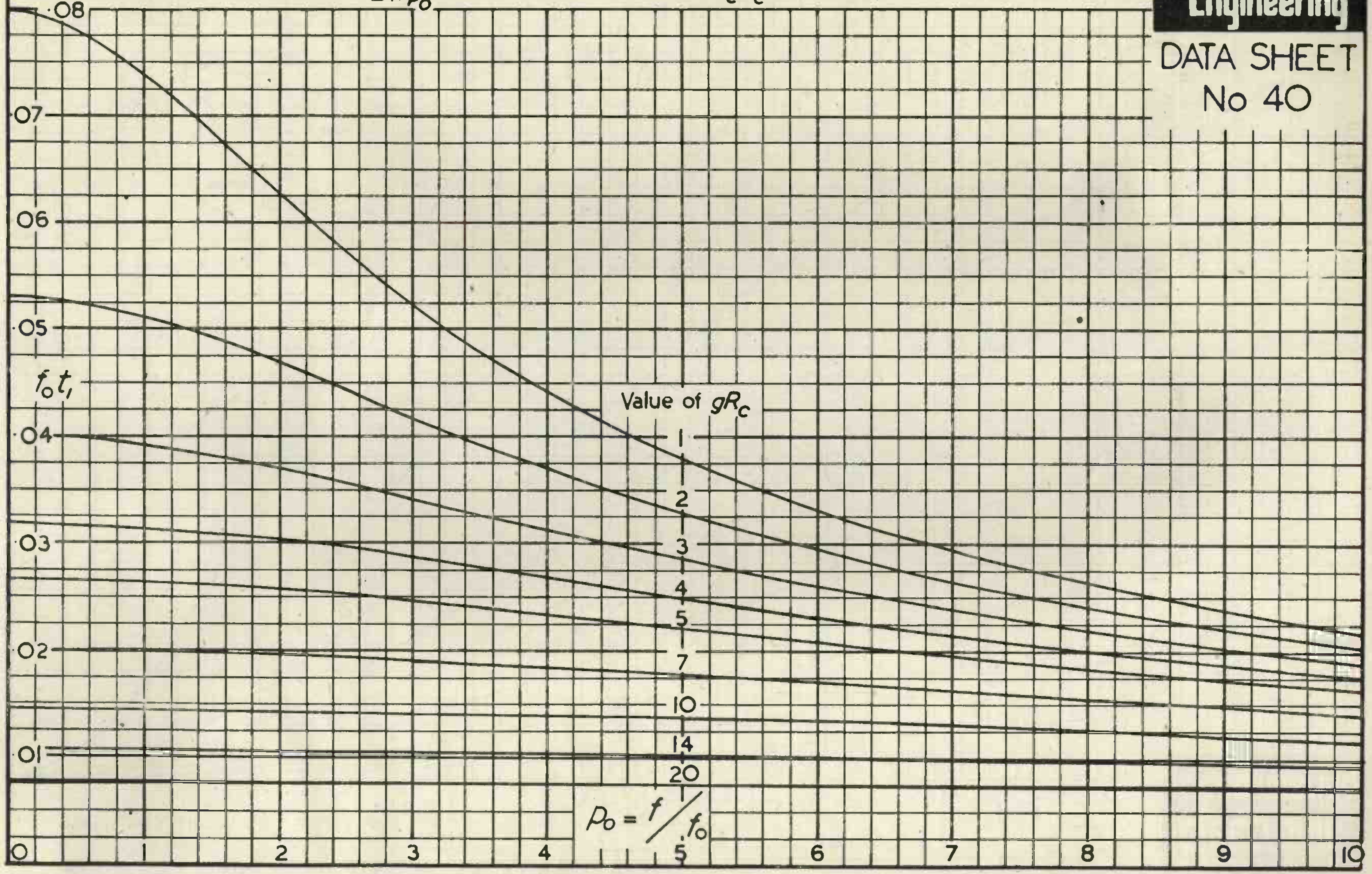
The Scale divisions on these Data Sheets are 0.45 cm.

Electronic Engineering

DATA SHEET
No 40

TIME DELAY OF CATHODE FOLLOWER STAGE

$$f_o t_1 = -\frac{\tan^{-1}\left(\frac{p_o}{1+gR_c}\right)}{2\pi p_o} \qquad f_o = \frac{1}{2\pi C_c R_c} \qquad \mu \gg 1$$



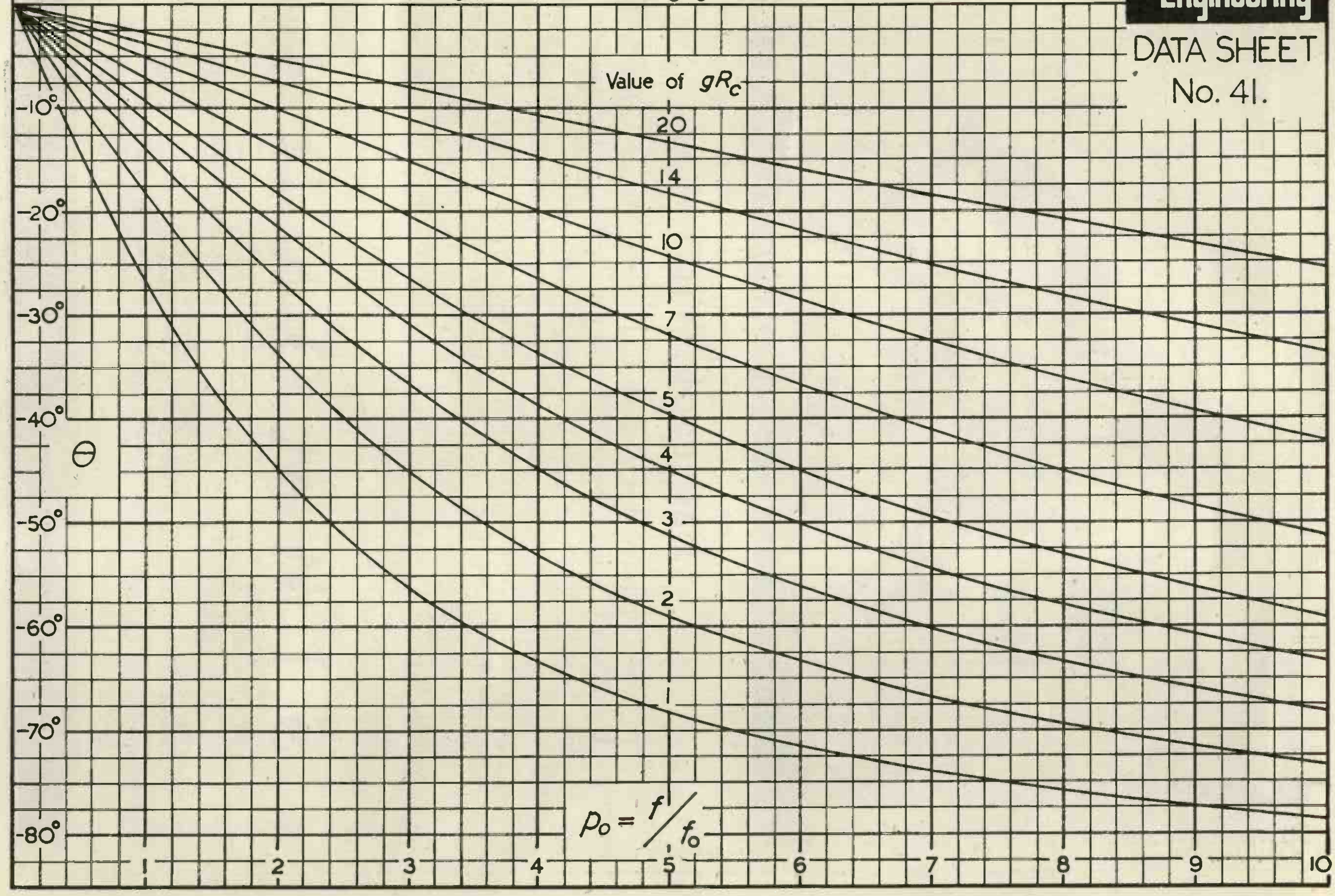
The Scale divisions on these Data Sheets are 0.45 cm.

Electronic Engineering

DATA SHEET
No. 41.

PHASE ANGLE OF OUTPUT VOLTAGE OF CATHODE FOLLOWER STAGE

$\theta = -\tan^{-1}\left(\frac{p_0}{1+gR_c}\right)$ $\quad f_0 = \frac{1}{2\pi C_c R_c}$ $\quad \mu >> 1$



Electron Optics

A lecture delivered before the Electronics Group and the Midlands Branch of the Institute of Physics on the 31st of October, 1942

By D. GABOR, Dr.Ing.*

INTRODUCTION. Electron optics was born fifteen years ago, when H. Busch proved in an important theoretical paper that the magnetic field of a cylindrical coil acts as an electron lens. Such coils, known as "concentrating coils" had already been used by a full generation of physicists, without anybody noticing their lens effect. Busch's paper was more than an eye opener; it was almost like a spark in an explosive mixture. In 1927 the situation in physics was such, that nothing more than the words "electron lens" were needed to start a real burst of creative activity. On the experimental side the work of Richardson, Wehnelt, Langmuir and others had provided easily controllable sources of slow electrons, and the production of high vacua was a matter of everyday laboratory routine. On the theoretical side, two years before Busch's paper Schrödinger discovered wave mechanics, by combining certain ideas of de Broglie with the half forgotten analogy of mechanics and optics which Sir William Rowan Hamilton had worked out almost exactly a hundred years earlier. By the time Busch had established the analogy of axially symmetrical electro-magnetic fields with lenses by somewhat laborious calculation of trajectories, the Hamiltonian analogy was again so familiar to theoretical physicists, that their immediate reaction was :—"Of course, we could have told you so! We could have told you even more. There are not only electron lenses, there must be a whole *optics* of electrons. You have only to look it up in Hamilton's writings and translate it into your experimental language!"

But the experimental physicists did not bother much about the Hamiltonian methods. The words "electron lens" were enough for them to ask themselves :—"What can we do with electron lenses? Can we make electron microscopes, electron telescopes, etc.? Within a few years of Busch's paper Knoll and Ruska had made the first electron microscope and Zworykin created the first practical high vacuum cathode-ray tube. Fifteen years of vigorous development followed. Looking back on these fifteen years, I can scarcely remember a paper dealing with any *practical* aspect of electron optics in which Hamiltonian methods had been used. Merely the terminology was borrowed from optics, but the authors, almost without exception built up their results from the consideration of electron trajectories. This situation is well reflected in the numerous excellent treatises on electron optics. The Hamiltonian analogy adorns a few pages, for the rest it is just dynamics of electrons in electric and magnetic fields, translated finally—when possible—into optical terms. This is not surprising, seeing that the computers of optical instruments do not use Hamiltonian methods either, but develop their lenses by "ray tracing," which is closely analogous to the computation of electron trajectories.

Until three years ago I took precisely the same view of the practical value of the Hamiltonian methods as most other workers in the field. I looked often with respect and admiration at their beautiful exposition in Born's "Optik," or their adaptation by Walter Glaser to electromagnetic fields, but it never occurred to me to use the methods. For the last three years I was not an active worker in electron optics, and dismissed the subject entirely from my mind. But when I had the pleasure of accepting the invitation to give this lecture, I found that I must have done some unconscious mental digestion in the meantime Suddenly I discovered, that certain problems on which I had laboured in vain with trajectories, could be solved almost effortlessly with Hamiltonian methods.† As I want my experience to benefit others, I intend in this lecture to adhere more closely to Hamiltonian ideas than usual.

I see also another justification for this. In the last six or seven years electron optics has definitely outgrown the analogy with optical instruments. Whereas in the older type of electron optical instruments the main problem was to bring electron trajectories to a focus *in space*, new devices have now come into existence, in which electrons are brought together *in space and time*. This new sort of focusing has been termed "phase focusing" by Brüche and Recknagel. I prefer "space-time focusing," as I think this makes it at once clear what we are talking about, and establishes a natural connexion with the older part of electron optics. These new dynamical devices, such as the velocity modulation tube, the Klystron, or the Betatron have no counterpart in light optics, but they are fully covered by the Hamiltonian Analogy, if it is suitably extended to four instead of three dimensions.

The Scope of Electron Optics.—It appears now necessary to re-define the scope of electron optics. A limit must be drawn somewhere, or else the subject would cover the whole theory of electronic devices. It is not sufficient either to say that electron optics extends as far as the Hamiltonian Analogy, as some of the "transit time" devices which we want to include in the subject are not covered by the analogy in the restricted sense in which it is often used. But if we adopted this narrower interpretation, we should have to exclude also space charges.‡ If, however, we allow space charges, *i.e.*, the action of electrons on one another inside the vacuum device, it might be logically claimed that the outer circuits should be also included. Indeed, the mathematician could treat these by the same multidimensional Hamiltonian methods. As an entirely logical division appears impossible, I propose with the inevitable measure of arbitrariness, to consider electron optics as *co-extensive with electron dynamics, including the theory of space charges and space currents, but excluding their reaction on the outer circuits.*

The Ballistic Problems as Illustration of the Hamiltonian Analogy.

Instead of dealing in generalities, I prefer to explain the optical treatment of mechanical problems in the simple example of the ballistic problem. The Newtonian treatment of this problem is well known to everybody. Given the initial data, *i.e.*, the position and bearing of the gun and its "charge" which determines the initial velocity, from these data the whole trajectory can be calculated, among other things also the final point at which it strikes the ground. The corresponding Hamiltonian problem, however, is as follows :—We consider as given the initial point (the position of the gun), the final point (the target), and the total energy of the projectile (which is

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† Strictly speaking, Hamiltonian ideas, not methods, as for the purpose of this lecture it will be sufficient to explain the refractive index in more detail than usual, but it will not be necessary to introduce Hamiltonian devices such as the "Eikonal."

‡ These are strictly speaking always "many electron problems" and could not be treated by the usual "one electron" methods.

given by the charge). The question is now :—What will be the trajectory or trajectories of the projectile between the gun and the target for a given charge? The solution will give among other things the elevation of the gun, which in the Newtonian treatment was considered as part of the initial data.

The answer is given by the Principle of Least Action, which is a special case of Hamilton's Principle. Among all the curves which connect the gun and the target the projectile will choose the one along which the " action " is minimum. The "action" is the integral of the kinetic energy over time

$$\tfrac{1}{2}\int mv^2.dt$$

and as the velocity $v=ds/dt$, this can be immediately transformed into an integral over the path ds.

$$\tfrac{1}{2}m\int v.ds.$$

The velocity of a projectile in the gravitational field depends only on the initial velocity (*i.e.*, on the charge), and on the height above the initial point. As we have considered the charge as given, we can say that v is a *function of the position of the projectile only*. Therefore the " action " integral can be calculated at once for all imaginable paths connecting the gun and the target.

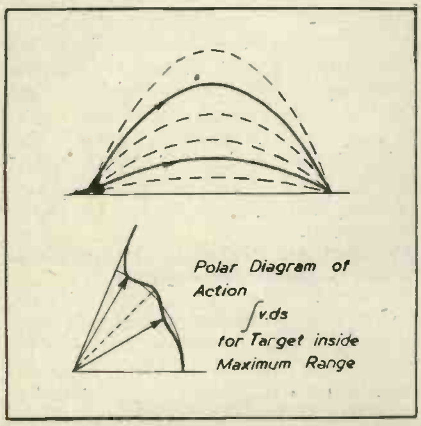

Fig. 1. The ballistic problem as illustration of Hamilton's Principle. Among all imaginable trajectories only those can be realised for which the " action " is a minimum. The two dynamically possible trajectories are shown in continuous lines.

To illustrate this, in Fig. 1 the family of parabolas is drawn which connect the two points. Of course we do not know *a priori* that the trajectory will be a parabola, but I have assumed this to simplify the illustration. We can now calculate the " action " for every one of these parabolas, and representing this in a polar diagram as a function of the elevation of the gun we obtain a curve with two maxima and two minima. If the projectile were strung on a wire connecting gun and target, it could travel along any of these parabolas. (Up to one which reaches a maximum height, beyond which the projectile would glide back into the gun. The diagram stops at this point; beyond it the action would become complex). But as the projectile is free, not strung on a wire, it can travel only along the two (continuously drawn) parabolas for which the action is minimum. As is well known, these two dynamically possible trajectories start with elevations symmetrical to 45°.

Sir William Rowan Hamilton gave a beautiful optical illustration to this rather dry mathematical theory. (Curiously, though the dry mathematical skeleton was remembered, the optical illustration was forgotten for almost a hundred years, until Schrödinger revived it). Hamilton noticed the analogy of the Principle of Least Action with Fermat's principle, according to which the light ray between two points travels in the shortest or the longest time. Let u be the velocity of light in a medium with refractive index n, and c the velocity in vacuo, $n = c/u$. Fermat's principle states that

$$\int dt = \int ds/u = \frac{1}{c}\int n.ds$$

must be a maximum or a minimum. Comparing this with the Principle of Least Action we can say at once, that the projectile will travel on a trajectory which *could* be chosen by a light ray in a suitably graded atmosphere, with an optical density gradually decreasing with height. The gradation must be such that at every point the refractive index n is proportional to the velocity of the projectile, v. (At a given " charge.")

One important difference must be noticed however. We see from the polar diagram in Fig. 1 that if we realised such a stratified atmosphere, an observer would see *four* mirages of the gun, corresponding to the two minima and the two maxima. (More exactly : The observer would see the gun once horizontally, in its real position, and in addition he would see three mirages). But only two of these paths, the ones corresponding to minima can be realised in the case of the mechanical problem.

We can now see at once why the Hamiltonian viewpoint is so eminently suited for the kind of problem likely to be encountered in electron optics. First of all the optical analogy provides us with a *complete terminology*. If it had done nothing else, this alone would be an outstanding achievement. Think of it what it means if we can do our thinking in terms of " lenses " instead of for instance " axially symmetrical bilinear transformations " ! Mathematicians may be able to do their thinking in such terms, ordinary physicists and engineers certainly can not. But apart from this, the Hamiltonian method provides also the most suitable mathematical system to formulate many problems (particularly focusing problems) even if the solutions are often reached in a more pedestrian way.

We can illustrate this immediately on the ballistic example if we ask the question :—" Can the gun focus its trajectories?" Yes it can, *if the target is at the maximum for a given charge*. (Fig. 2). If the elevation of the gun is approximately 45°, small deviations from this angle will cause only negligible deviations at the target. This is a well known result, only we are not used to thinking of it as " focusing effect of the gravitational field." In the polar diagram of the action a very flat minimum appears, in which one maximum and two minima merge into one. We can immediately generalise this :—*Focusing will occur if the action has a minimum of a higher order.*

The Electron-Optical Refractive Index

The methods applied in the ballistic example can be transferred without any essential change to the case of an electron (or any charged particle), in an electrostatic field. The reason is that both the gravitational and the electrostatic field are (scalar) potential fields. Consequently the " refractive index " of an electrostatic field will be proportional to the velocity v of the electron, and independent

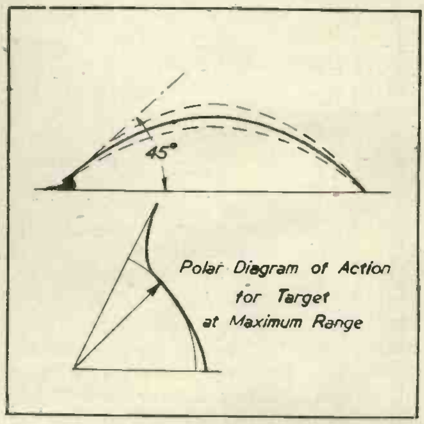

Fig. 2. Focusing effect of the gravitational field. If the gun elevation is 45° the range is maximum, and small departures from the elevation will not affect it. The focusing effect reveals itself in the action diagram as a very flat minimum.

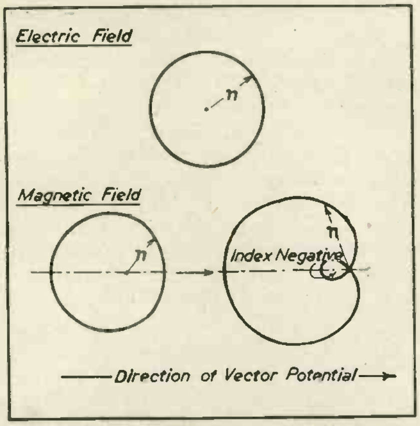

Fig. 3. Electron-optical refractive indices. Top figure :—Electric field. The refractive index is independent of direction, the n-surface is a sphere. Bottom figure :—Magnetic field. The n-surface is still a surface of revolution, but no more a sphere. It becomes increasingly asymmetrical with increasing magnetic field. In strong fields the index becomes negative in a certain angular range.

of its direction. It can be represented in a polar diagram as a sphere. (Fig. 3).

We have always qualified the definition of the refractive index by the clause " at a given charge." But this restriction is so natural in most electrostatic problems, that it can be almost forgotten. In practically all electron optical devices the electron source is a low temperature hot cathode, which emits electrons with velocities of the order of a tenth of a volt. As we have usually accelerating potentials of hundreds or thousands of volts, we can often neglect the small initial velocity and consider the cathode as emitting the electrons with zero ballistic charge. We must not, however, forget the initial velocity altogether, it has some very important consequences.

The electrostatic field behaves therefore " optically " like an isotropic medium with varying density, *e.g.*, like a gas or liquid. It is very different with the magnetic field. I can not dwell here on the derivation of the refractive index in this case. (It is based not on the Principle of Least Action, but on the more general Hamiltonian Principle, with the Lagrangian function instead of the kinetic energy). I can state only without proof that it is

$$n = v - \frac{e}{m.c} A.\cos(\mathbf{A.v.})$$

where A is the absolute value of the vector potential of the magnetic field and **(A,v)** is the angle between the vectors **A** and **v**. Representing this again as a surface, with radii proportional to n, we obtain a very queer shape. The refractive index in magnetic fields is often dismissed in treatises on electron optics as "dependent on direction, as in crystal optics," but this is rather covering up the difficulties of the question. In crystal optics the n-surface is never worse than an ellipsoid! The most singular feature of the magnetic refractive index is, that it is *not the same in opposite directions.* This has no counterpart in ordinary optics. Light can always travel back on the same path on which it came, but not an electron in a magnetic field! It is well known that the magnetic force on an electron is proportional to the velocity vector and at right angles to it. If therefore an electron travelling one way in a magnetic field is deflected clockwise, it will be deflected counterclockwise if it is coming the opposite way. The assymmetry of the refractive index expresses this fundamental (mis-)behaviour of the magnetic force. Moreover, the refractive index can become negative! Looking at Fig. 3 it can not surprise us that hardly anybody has ever treated a magnetic problem with optical methods.[*]

Electrostatic Lenses—The Hamiltonian Analogy immediately explains the existence and properties of electrostatic lenses, of which an example is shown in Fig. 4. As we have seen, the refractive index is proportional to the electron velocity, and this again is proportional to the square root of the potential[†] measured from the cathode as zero level. All we have to do is to imagine the field replaced by a medium of continuously increasing density. This is shown in Fig. 4 by increasing density of the shading. We can therefore consider the field as a succession of infinitesimally thin lenses, each bounded by a couple of equipotential surfaces. As a combination of lenses is again a lens, we have a full qualitative explanation of the electrostatic lens.

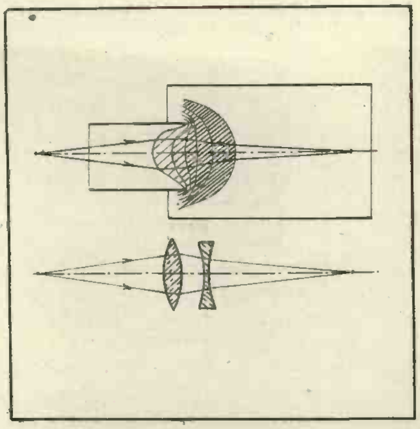

Fig. 4. Electrostatic electron lens. The potential increases from left to right. An exact optical replica could be constructed by filling the space between equipotential surfaces with materials of optical density proportional to the square root of the voltage. A condensing and a dispersing lens can be used as rough illustration, as shown in the bottom figure. The condensing lens is always stronger than the dispersing lens.

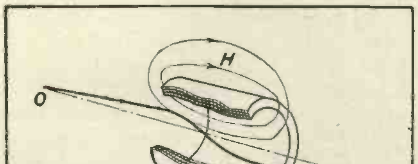

Fig. 5. Magnetic electron lens. Proof that a trajectory starting from a point of the axis will again intersect the axis.

Magnetic Lenses—The lens effect is far less evident in the case of a " magnetic lens," such as can be produced by the field of any axial coil. (Fig. 5). The magnetic field exerts a force on the electron at right angles to the field lines and to the movement of the electron. It is evident that the trajectory will circle round the axis in a sort of spiral. The question immediately arises :—" If an electron starts from a point of the axis, will the trajectory return somewhere to the axis, or will it miss it?"

It would not be profitable to apply simple optical considerations to this circling movement of the electron as in every plane at right angles to the axis the refractive index has the queer shape illustrated in Fig. 3. For once we borrow, without proof, a result from dynamics, which fully describes the movement of the electron at right angles to the axis. This is as follows :—

Let us rotate the trajectory round the axis. We obtain a surface of revolution. Between two points on its trajectory an electron acquires an angular momentum round the axis which is proportional to the total magnetic flux through the part of the surface of revolution which is defined by the two points.[*]

If an electron starts anywhere far outside the field and leaves it again, the resulting flux through the en-

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[*] I say " optical " not " Hamiltonian." The general Hamiltonian theory of magnetic fields has been developed by W. Glaser.

[†] This is true only for moderate voltages. At near-light speeds the refractive index increases faster than the square root of the voltage.

[*] This theorem was derived by Busch, but first expressed in these words by Bouwers.

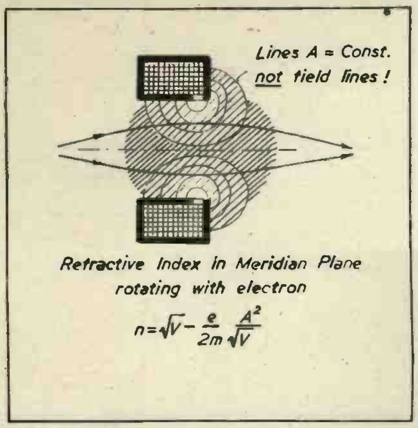

Fig. 6. Magnetic electron lens. Abstracting from the rotation of the trajectory around the axis the lens effect can be expressed by an isotropic refractive index, as in the case of the electrostatic lens, but with a different formula. The optical replica would have a density distribution as indicated by varying degrees of shading.

velope of its trajectory is zero, its total gain of angular momentum is therefore *nil*. If we consider an electron which has started from the axis, *i.e.*, with an angular momentum zero, it will have again zero momentum when leaving the field. This means that after leaving the field the movement of the electron will be strictly in a meridian plane, and either the path itself, or the tangent to it will intersect the axis.

This theorem, which we had to borrow from dynamics enables us now to return to the optical analogy. We consider the movement of the electron in a meridian plane which *turns with the trajectory round the axis*. In this rotating meridian plane it is quite profitable to apply optical considerations. We have seen in Fig. 3 that though the surface representing the refractive index in a magnetic field has a very strange shape in any plane containing the vector potential, *every section at right angles to the vector potential is a circle*. But the vector potential of an axial coil runs in circles round the axis, therefore we can interpret the movement of an electron in the (rotating) meridian plane as caused by *a refractive index of the familiar isotropic type*. (Fig. 6). Taking the rotation of the trajectory into account one obtains for this an expression.

$$n_{\mathrm{m}} \sim v - \frac{e^2}{m^2}\,\frac{A^2}{v}$$

or, expressing the electron velocity v by the electrostatic potential V, and leaving away an irrelevant constant factor

$$n_{\mathrm{m}} \sim \sqrt{V} - \frac{e}{2m}\,\frac{A^2}{\sqrt{V}}$$

In the purely magnetic lens V is constant, therefore the lines of constant refractive index will be the lines A = const. These lines are similar to the magnetic field lines, but not identical with them. On the axis A is always zero, therefore the magnetic refractive index will always *decrease* with increasing distance from the axis. In Fig. 6 this is again indicated by shading with varying density. This result may be of some practical importance, as it allows the transformation of a magnetic field into an *equivalent electric field* $V \approx n_{\mathrm{m}}^2$, and makes it possible to trace trajectories by well known methods. It must be noted, however, that the formula for n_{m} is valid only for electrons which have crossed the axis, or which have started from a cathode *outside* the magnetic field.

This figure shows clearly, and probably somewhat surprisingly, that the optical analogy of a magnetic lens is very unlike anything which we are used to call a " lens " in optical practice. There is nothing of the lentil shape about it, from which the word " lens " is derived. On the contrary, the field is built up of thin hyperboloidal shells, with refractive indices gradually decreasing. Near the axis the decrease is approximately proportional to the square of the axial distance. Though this is a very unusual type of lens, its effect is the same as of the familiar type. The electron will run along the axis on a potential crest, which acts on a negative charge like a trough for a heavy body. If it departs from the axis, the slopes at both sides of the trough will drive it back towards it. We understand at once why the magnetic lens acts always as a converging, never as a diverging lens. (The corresponding theorem for the electrostatic lens is not so easily proved).

It may be added, that the optical analogon could be made perfect if we made up a model out of transparent hyperboloids and rotated them at extreme speed. In such a rotating lens light would describe just the same sort of spiral path as an electron in a magnetic lens. There is, however, little hope of realising such a model experimentally as the convection of light by moving media (Fizeau effect, really a relativistic effect) becomes noticeable only at speeds at which any material would burst by centrifugal force.

Lens Errors.—As soon as Busch had proved that axially symmetrical fields, electric or magnetic, behave in the first approximation as lenses, a question very naturally suggested itself: " Can we make electron lenses as perfect as optical lenses?"

The question could not be answered off hand; many years of work were needed before it was fully elucidated. As compared with light optics electron optics started with a great advantage, and with a great disadvantage. The great advantage was that in electron optics we can realise refractive indices far larger than in light optics. In optical instrument the ratio of the largest and smallest refractive index is never more than 2 : 1, in electron optical devices we can easily realise ratios of 1,000 : 1.*

On the other hand electron optics started with a great disadvantage. Modern applied optics has a great variety of optical glasses (and nowadays also plastics), at its disposal, with a considerable variety of chromatic dispersion coefficients. Electron optics has only two "media," the electric and the magnetic field, both with fixed dispersion properties. Nor can we shape the electron optical lenses as freely as the optician shapes his lenses. We are limited by the laws of zero divergence for the electric field, zero curl for the magnetic field. (Freedom from space charges and space currents).

It will be useful for the understanding of the following explanations if we anticipate the results, and sum up the essence of 15 years of development in a few words:—

It was *not* possible for the science of electron optics to design optically perfect or at least highly corrected lenses, not because of lack of ability of the theoreticians, but because of the laws of the electro-magnetic field. The disadvantage which we mentioned above was too great. On the other hand electron optical practice could overcome all practical difficulties by availing itself of the great in-

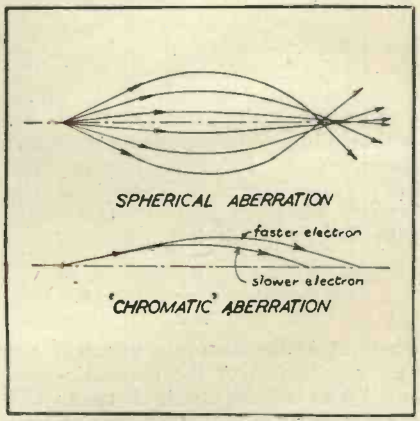

Fig. 7. Lens errors for axial object points.

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* As the initial energy of the electrons emitted by a barium cathode is of the order 0.1 volts, this can be realised with about 100,000 volts accelerating voltage.

herent advantage of electron optics, by the use of *high voltages*. The best theory could do was to point out the essential limitations, and to save the practical workers in the field from wasting time and energy in trying to reach unattainable ideals. In the short survey which follows, I want therefore to lay particular emphasis on these theoretical limitations.

The first question which arises is: " Can an electron lens system image a point of the axis in a point?" In this case there are two defects possible, spherical aberration and chromatic aberration. Both are illustrated in Fig. 7. Scherzer has shown in an important paper[1] that neither can be eliminated Trajectories starting at a larger off-axis angle will always cross the axis *before* trajectories closer to the axis. On the other hand faster electrons starting at a given angle will always cross the axis *beyond* slower electrons with the same initial tangent. The deeper reason for both defects is that it is impossible to realise electromagnetic *diverging* lenses.

Electron Mirrors

A certain possibility of overcoming these defects emerged in 1936 when Hottenroth and Recknagel[2] (AEG Laboratory) discovered that *electron mirrors* may have aberrations opposite to those of lenses. Such mirrors can be realised as shown in Fig. 8, but also with electrodes of the same shape as in lens arrangements, simply by giving the second electrode a potential sufficiently negative to repel electrons, and to return them towards the same side from which they came. Fig. 9 illustrates that spherical and chromatic aberration in the case of mirrors, and shows that they may have signs opposite to those of lenses. By combining lenses and mirrors it may be therefore possible to produce aberration free systems.

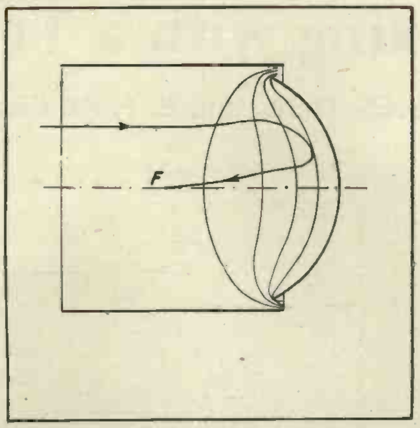

Fig. 8. Electron mirror. Similar effects can be obtained also with electrode arrangements as in Fig. 4, if the second electrode is strongly negative against the cathode.

But this was a forlorn hope, for several reasons. In the case of mirrors, the image appears at the same side as the electron beam arrives. There are two possibilities to overcome this. One is to use two mirrors, as in the Cassegrain systems of telescopes. But this has two difficulties: *First*, the second mirror creates a very bad disturbance in the arriving beam, by its edge field and by the struts which are necessary to hold it. (There is no electron optical equivalent to " painting the struts black.") *Second*, this second mirror, or image screen, cuts out the paraxial rays, and forces the designer to use trajectories so far from the axis, that it is not sufficient to correct the system to the third order. Even in light optics such mirror systems can be profitably realised only with non-spherical surfaces. There is also another possibility, viz., to incline the mirror. But skew (non-centred) systems are generally so difficult that even in light optics advances in this direction have been made only recently.[3]

Apart from these theoretical difficulties, electron mirrors offend badly against the fundamental *practical* principle of electron optics (of which more later):—" *Do not slow down your electrons if you can help it!*" No wonder that images obtained with electron mirrors are mostly of extremely bad quality, and the hope of obtaining fully corrected systems with electron mirrors must be classed as an illusion.

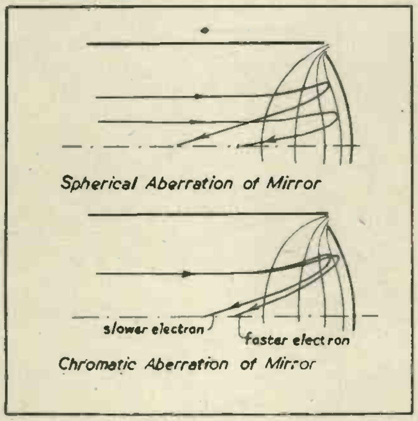

Fig. 9. Axial aberrations of electron mirrors.

There is also another possibility to eliminate the spherical aberration, by *space charges*. There seems to be a fairly widespread belief, that the space charge of an electron beam will always prevent its focusing in a point. This is true, of course, for a beam which would be stigmatic without a space charge. But generally the space charge produces an error of opposite sign to the spherical aberration, and it is possible that under certain conditions the two just cancel one another* It can be easily shown that this is not impossible, as the energy stored in a perfectly stigmatic beam carrying a finite current is finite. (Perhaps it may be worth while to refute the fallacious belief that if an electron beam passes through a point, the electrons much touch each other. They will pass in single file!) But this remedy has no practical importance. It can not be applied in electron microscopes, where the beam has an incalculable shape and density distribution, depending on the object, and in cathode-ray tubes, for reasons to be discussed later the spherical aberration presents no serious problem.

We see that the few remedies so far discussed are useless. Yet, practice has overcome the lens defects, even in the case of the only device in which the spherical and the chromatic aberration threatened to be dangerous limitations, in the electron microscope. I have mentioned already the simple panacea of practice: *high voltages*. In an electron microscope objects are made visible by electron absorption and scattering. If the beam voltage is raised, absorption becomes negligible and the energy losses of the electrons become less and less, not only relatively, but even absolutely. The scattered beam becomes increasingly homogeneous, and the chromatic effect becomes negligible. As regards the spherical aberration and also other lens effects, they are also reduced because the scattered beam is confined to a rather narrow angle. (Further practical remedies will be discussed in connexion with the electron microscope).

Theory could give only one helpful hint, and this came from R. Rebsch,[4] a pupil of Prof. Scherzer. Rebsch showed, that although the spherical aberration could not be eliminated, it could be reduced in electron microscopes to any insignificant value, by allowing the field of the objective lens to extend to the object, in a way which is equivalent to pushing a very small, very weak lens very close to it. This artifice is useful also in the case of electrostatic microscopes imaging the cathode, where high voltages would not bring any improvement.

(*To be continued*).

[1] O. Scherzer, Zeitschr. f. Physik 101, 593, 1936.

[2] G. Hottenroth Zeitschr. f. Phys, 103, 460, 1936. A. Recknagel, ZS. f. techn. Physik, 17, 643, 1936.

[3] Mainly through the work of Dr. C. R. Burch and Dr. E. H. Linfoot of Bristol.

[4] R. Rebsch, Ann. d. Phys, 5, 31, 551, 1938.

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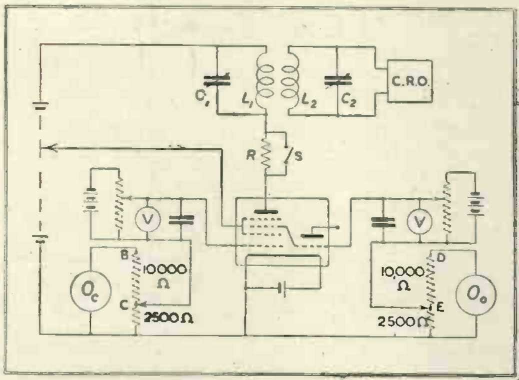
* For other possible beneficial effects of the space charge cf. O. Klemperer, Brit. Pat. 534,215.

Frequency Mixing with a Triode-Hexode

By E. HUGHES, D.Sc., and E. F. PIPER, A.M.I.E.E.*



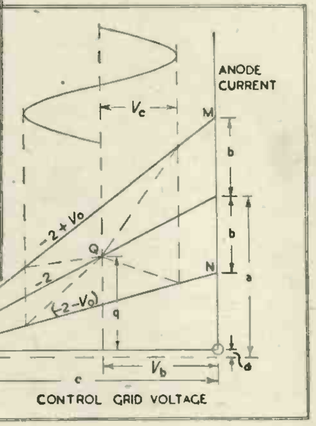


Fig. 1. (right) Characteristics of the triode-hexode.

Fig. 3. (above) Circuit used for experimental verification of expression 2.

WHEN two sinusoidal voltages of frequency f_1 and f_2 are added together, the amplitude of the resultant wave varies between a maximum and a minimum at the beat frequency of $(f_1 - f_2)$; but a beat-frequency *component* is not obtained unless this resultant wave is either partially or completely rectified. It is of interest to examine how such partial rectification is obtained with the triode-hexode, and to deduce the factors that govern the amplitude of the beat-frequency component.

Graphs showing the variation of anode current with control-grid voltage for a triode-hexode are reproduced in Fig. 1. Each graph corresponds to a given oscillator-grid voltage. It will be observed (*a*) that the graphs are linear over a considerable portion of their lengths, (*b*) that for a given control-grid voltage, the vertical distances between the graphs are practically proportional to the variation of oscillator-grid voltage.

Suppose the control grid to have a negative bias, V_b, and the oscillator grid a negative bias of, say, 2 volts. Consequently, the valve is being operated about point Q in Fig. 1, and it is evident that an alternating voltage of limited amplitude applied to *either* the control or the oscillator grid causes a corresponding undistorted variation of the anode current. But when sinusoidal voltages of slightly different frequencies are applied to the two grids *simultaneously*, the anode current is distorted as shown in Fig. 2. This waveform is the oscillogram of p.d. across a resistance *R* (Fig. 3) of 1,000 ohms inserted in the anode circuit when voltages of about 100 and 124 kc/s are applied to the control and oscillator grids respectively. It will be seen that the mean value of the anode current varies at the beat frequency of about 24 kc/s.

Let us now consider how this waveform can be derived from the characteristics shown in Fig. 1. It is found that the straight portions of the graphs, when produced, converge to a point P on or near the X axis.

Suppose the valve to be operated about point Q and an alternating e.m.f. of amplitude V_c and frequency f_c to be applied to the control grid only; then, using the symbols shown on Fig. 1,

instantaneous anode current $= i$

$$= \left\{ g(c - V_b + V_c \sin 2\pi f_c t) - d \right\} \quad (1)$$

where g = slope of graph PQ = a/c.

Next, suppose an alternating voltage of amplitude V_o and frequency f_o to be applied to the oscillator grid simultaneously with the above-mentioned voltage to the control grid. Then the slope will vary between a maximum corresponding to graph PM and a minimum corresponding to PN;

i.e., maximum slope $= \dfrac{a + b}{c}$, and

minimum slope $= \dfrac{a - b}{c}$.

Since the intercept b is practically proportional to the value of V_o, $b = kV_o$, where k is a constant for a given valve operating under given conditions.

Hence, *instantaneous* slope

$$= \frac{a + kV_o \sin 2\pi f_o t}{c},$$

and the general expression for the instantaneous value of the anode current is obtained by substituting for "g" in (1), thus:

$$i = \left\{ \frac{a + kV_o \sin 2\pi f_o t}{c} \left[c - V_b + V_c \sin 2\pi f_c t \right] \right\} - d$$

$$= \left\{ \frac{a}{c}(c - V_b) - d \right\} + \frac{k(c - V_b)V_o}{c} \sin 2\pi f_o t + \frac{a}{c} V_c \sin 2\pi f_c t$$

$$+ \frac{k}{c} V_o V_c \sin 2\pi f_o t . \sin 2\pi f_c$$

$$= q + \frac{k(c - V_b)V_o}{c} \sin 2\pi f_o t + gV_c \sin 2\pi f_c t$$

$$+ \frac{k}{2c} V_o V_c \left\{ \cos 2\pi (f_o - f_c) t - \cos 2\pi (f_o + f_c) t \right\} \quad .. \quad .. \quad (2)$$

where $q = g(c - V_b) - d$
= anode current with no alternating voltages applied to the control and oscillator grids.

From expression (2), it follows that:

(1) The anode current has components of frequencies f_o, f_c, $(f_o - f_c)$ and $(f_o + f_c)$.

(2) The amplitude of the component having frequency $(f_o - f_c)$ is proportional to the *product* of the two applied voltages and independent of the biases applied to the control and oscillator grids, so long as the valve is being operated within the linear limits.

(3) The conversion conductance, namely, the change in anode current at beat (or intermediate) frequency per volt applied to the control grid, is $\dfrac{kV_o}{2c}$, i.e., $\dfrac{b}{2c}$ in Fig. 1. Since the maximum

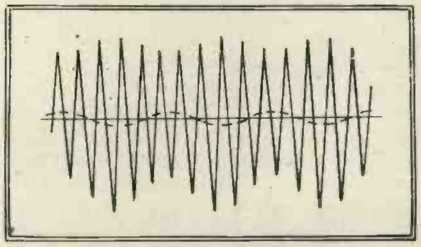


Fig. 2. Waveform of anode current with voltages of different frequency on control and oscillator grids.

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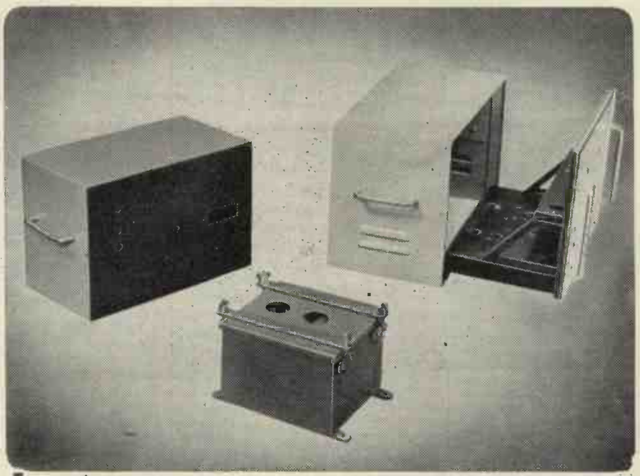

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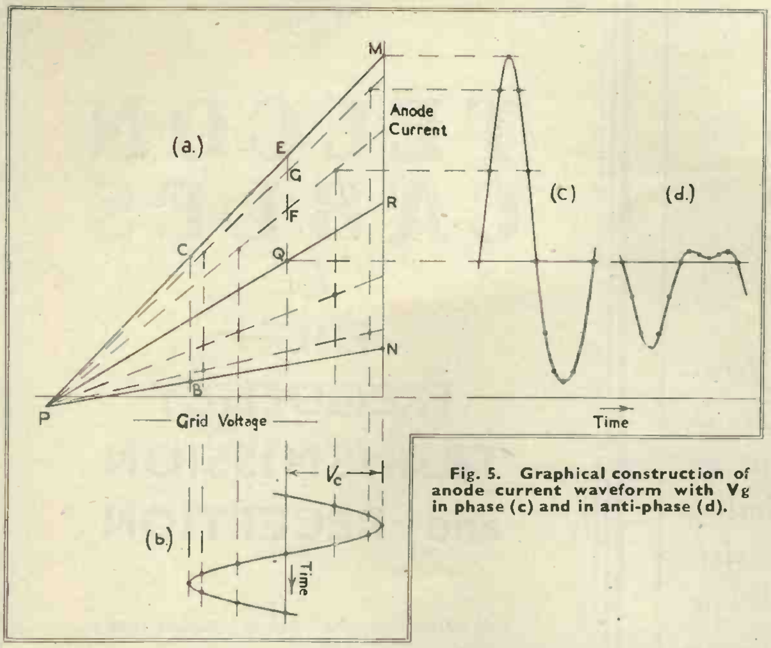


Fig. 5. Graphical construction of anode current waveform with Vg in phase (c) and in anti-phase (d).

value of b cannot exceed a, it follows that the conversion conductance cannot exceed half the mutual conductance, which is a/c.

Conclusions (1) and (2) can be confirmed experimentally with the aid of the circuit shown in Fig. 3. O_c and O_0 are two Avo-oscillators; L_1 and L_2 are suitable inductances, very loosely coupled; and R is a relatively low resistance of, say, 1,000 ohms, short-circuited by switch S except when it is used to give the waveform of the anode current. Non-inductive resistances of 2,500 and 10,000 ohms are connected in series across each oscillator output to enable the voltage applied to the respective grids to be each increased five-fold.

Suppose the frequency of O_c to be 110 and that of O_0 to be 250 kc/s. We should be able to detect in the anode current the presence of these two frequencies, together with frequencies of 140 and 360 kc/s. The simplest procedure is to adjust O_c (with O_0 off) to each of these frequencies in turn; and for each frequency, to note the scale readings on C_1 and C_2 for the maximum output on the cathode-ray oscillograph, CRO, connected across L_2C_2.

The oscillators O_c and O_0 are then set for 110 and 250 kc/s respectively, and it is found that each of the four frequencies can be tuned in on L_1C_1 and L_2C_2, thereby confirming the first conclusion.

The multiplicative action was checked by tuning L_1C_1 and L_2C_2 to the beat frequency of 140 kc/s, varying the amplitude of the voltages applied to the control and oscillator grids and noting the perpendicular distance between the positive and negative peaks on the oscillograph screen. The results obtained with an X24 triode-hexode are given in the table in the next column :—

Control-grid tapping	Oscillator-grid tapping	Product of relative amplitudes	Vertical distance between peaks
C (1)	E (1)	1	0.15 inch
B (5)	E (1)	5	0.7 inch
C (1)	D (5)	5	0.8 inch
B (5)	D (5)	25	3.5 inch

The figures in brackets represent the corresponding relative amplitudes of the applied voltages. It is seen that the output beat-frequency voltage is practically proportional to the product of the input voltages, and the multiplicative action is therefore confirmed.

When performing this test, it is important to check that the resistances used as potentiometers operate satisfactorily at the frequencies employed. This can easily be done by setting each oscillator in turn to 140 kc/s (the other being off), and noting the output voltages on the CRO when the contacts are on B and C or on D and E.

With reference to the second part of conclusion (2), it is not obvious from the valve characteristics that the amplitude of the beat-frequency component should be independent of the grid biases—especially the bias on the control grid. With the potentiometers shown in Fig. 3, it was possible to vary these biases independently, the *LC* circuits being tuned to the beat frequency of 140 kc/s. The variations of the output voltage are shown in Fig. 4, graph A being for the control grid and B for the oscillator grid. It should be pointed out that for the anode and screen voltages used in this test, the graphs of Fig. 1 began to depart from linearity when the control-grid bias exceeded about −2 volts, while the vertical spacing per volt between the graphs increased slightly as the oscillator-grid bias was increased from 0 to −4 volts. Hence, it follows that within the limits of linearity, the whole of conclusion (2) can be confirmed experimentally.

It is realised that the above mathematical proof neglects the effect of load impedance. The dynamic impedance of L_1 C_1 (Fig. 3) is negligible compared with the anode a.c. resistance of the mixer at all frequencies except when L_1 C_1 is in resonance. Even for this condition, the dynamic impedance is usually relatively low and therefore does not greatly affect the anode current. This can easily be demonstrated by connecting the CRO across R (with S open) and noting the waveform as C_1 is varied through resonance.

It is of interest to apply expression (2) to the case where the two frequencies are equal and where the voltages applied to the control and oscillator grids are (*a*) in phase, and (*b*) in anti-phase.

When the control and oscillator grid voltages are in phase, it follows from (2) that—

$$i = q + \left\{\frac{b\,(c - V_b)}{c} + gV_c\right\}\sin 2\pi ft + \frac{bV_c}{2c}\left\{1 - \cos 2\pi\,(2f)\,t\right\} \qquad \ldots \quad (3)$$

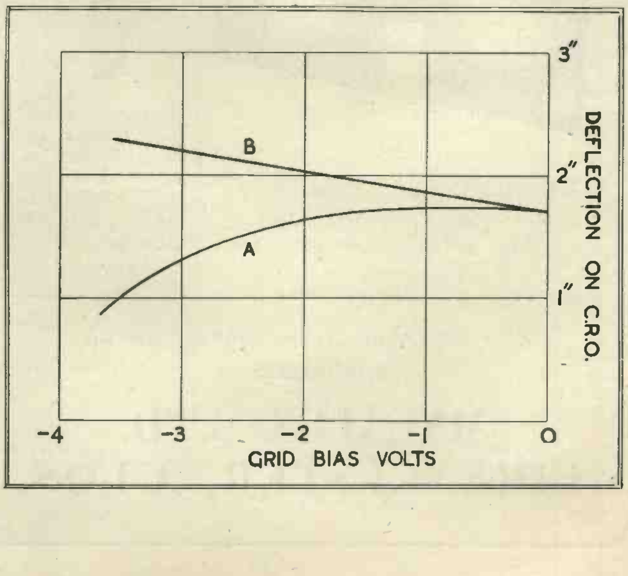


Fig. 4. Variation of output voltage with variation of control grid bias (curve A) and oscillator grid bias (curve B)

Hence the effect of the applied alternating voltages is to increase the mean value of the anode current by

$$\frac{\mathrm{b}V_\mathrm{c}}{2\mathrm{c}}.$$

The effect of reversing, say, the oscillator-grid voltage relatively to the control-grid voltage is to reverse the sign in front of " b " in expression (3). Hence, with the voltages in anti-phase,

$$i = \mathrm{q} + \left\{gV_\mathrm{c} - \frac{\mathrm{b}(\mathrm{c} - V_\mathrm{b})}{\mathrm{c}}\right\} \sin 2\pi ft - \frac{\mathrm{b}V_\mathrm{c}}{2\mathrm{c}}\left\{1 - \cos 2\pi\,(2f)\,t\right\} \qquad (4)$$

Hence the effect of applying the voltages in anti-phase is to reduce the mean anode current by $\mathrm{b}V_\mathrm{c}/2\mathrm{c}$ compared with the initial value q. The conversion conductance, b/2c, is therefore given by the change of the mean anode current when a p.d. of 1 volt peak value is applied to the control grid, or by half the change of mean anode current when that p.d. is *reversed* relatively to the oscillator-grid voltage (*see* Rapson's *Experimental Radio Engineering*, p. 34).

Expressions (3) and (4) indicate the introduction of second harmonics into the waveform of anode current. This can be confirmed experimentally by noting the waveform of p.d. across a low resistance in the anode circuit or by the graphical construction shown in Fig. 5. The valve is assumed to be biased so as to operate about point Q; and the oscillator grid voltage is assumed to be such that when it is at its maximum positive value, the anode current lies on graph CM, and when at its maximum negative value, it lies on BN. By drawing intermediate graphs such that QF = QE sin 30° = 0.5 QE, and QG = QE sin 60° = 0.866 QE, etc., it is possible, by projecting upwards from the curve of control-grid voltage and then horizontally, to deduce curve (*c*) of anode current when the control and oscillator grid voltages are in phase and curve (*d*) when they are in anti-phase. Oscillograms of p.d. across a low resistance inserted in the anode circuit of a triode-hexode, the oscillator and control grids of which were connected to separate secondaries of a low-frequency transformer, were in close agreement with the curves deduced graphically as in Fig. 5.

It was found that when two separate oscillators were used, it was much more difficult to get oscillograms of anode current showing waveform (*d*) of Fig. 5. This particular shape was very sensitive to change of oscillator frequency and to adjustment of the time and synchronising controls on the CRO.

All experimental evidence confirmed that expression (2) does represent correctly the factors that govern the operation of a triode-hexode when used as a frequency-mixer.

MODERN DESIGN

By WHITNEY S. GARDNER

You have removed the chassis from your receiver to replace a resistor. That job done, it simply becomes necessary to slip the chassis inside the cabinet and attach a few knobs. A cinch.

You place the chassis squarely on the table and, holding the cabinet firmly between the left and right hands, slowly lower it. The front bottom edge strikes the condenser shaft, so you ease the cabinet forward. Again you lower it. This time the rear edge of the cabinet strikes the power transformer.

Remove the cabinet. Study the situation. Remove the tubes to get more clearance. Lower cabinet. Ah, it clears the transformer, but strikes the condenser shaft. Remove cabinet and place it upside down on table. Invert chassis and lower it into cabinet. There, just tip it slightly and it clears both transformer and condenser shaft. Oh, oh—it slipped. Now it's gone down too far. Invert the whole works. Now lift up on cabinet. You can't. That's because condenser shaft and transformer have a tight grip on cabinet. Front of cabinet bulges. Bang it sharply on table. Cabinet gives suddenly and comes off chassis. So does dial pointer.

Straighten dial pointer. That's good enough. Lower chassis into cabinet, but not too far. Something is holding it back. Why, of course; the power cord should have been threaded through that grommet first. Now once more. At last it's down there, but the condenser shaft is still causing the cabinet to bulge. That's because it is not in line with the hole. It should be moved one-half inch to the left. It doesn't slide. Better take it off again. That time you only ripped off the dial light wire.

Resolder it. Take a rest. That tin cabinet was on there once and it isn't going to whip you. There, on it goes and all shafts lined up. Just goes to show what clear, calm thinking will do. Attach knobs. Attach aerial and ground, plug in power cord and snap on the power. Nothing happens. That's because you forgot to put the dial light back in the socket. Oh, well, you always have the room light on anyway. But wait, no hum, no hiss. Well, naturally—you took out all the tubes, remember?

That narrow opening in the back is just large enough for your hand. There—six of them are in. You have one left. That's the one that goes up on the shelf near the dial light. You can't reach it. Someone put it in there; you ought to be able to. The awful truth dawns. It has to be inserted before the cabinet is attached!

" Hey, Marjie, what picture are they showing at the Strand to-night? "

—*Q.S.T.* Vol. 25. No. 9. September, 1942, page 18.

Taking Care of Transmitters

The following suggestions for the care and maintenance of transmitting valves are taken from a useful booklet: "Prolonging Tube Life," issued by the Heintz Kaufman Co., California. The original is illustrated with amusing sketches of which three are reproduced here.

THE life of a tube in normal service depends upon the number of watts it is required to dissipate on the plate. If the plate loss in watts is reduced, the life goes up proportionately. In other words, tube life may be expressed as "watt-hours of plate dissipation," and any reduction in *watts* results in a gain in *hours*. Therefore, it is advisable to adjust every circuit so that the highest efficiency is obtained.

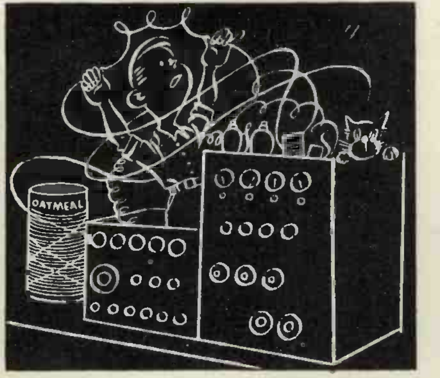
Keep circuits properly tuned. A small amount of detuning in the plate circuit causes a rapid increase in the plate dissipation of the tube. Circuits often detune as the transmitter heats up, and readjustment is then necessary.

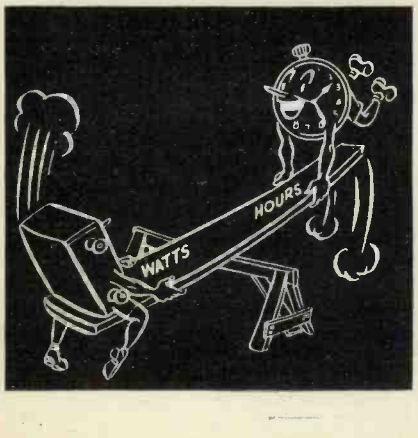
In Class B audio amplifiers, the "no-signal" plate current can often be reduced without resulting in harmful distortion. This reduction saves precious "watt-hours."

Minimise stray circuit losses in Class C r.f. stages, and make sure that the loading on the tube is useful loading. To test for this, disconnect the useful load and check the unloaded plate current.

The grid current of a triode is a good indicator of the amount of r.f. grid driving voltage required. In ordinary Class C r.f. amplifiers the grid current should be roughly $\frac{1}{4}$ to 1/6 the d.c. plate current of the tube. For doubler or tripler service, where large grid-leaks on the order of 50,000 ohms are employed, the ratio of grid to plate current may fall off nearer 1/10.

A good experimental way to adjust to the proper amount of grid drive, is to reduce the drive until the efficiency of the tube starts to fall off. This will be indicated by a visible increase in plate heating. The grid drive should then be restored somewhat above this fall-off point.





When the tube is idle the filament should be turned off. When both the plate and filament voltage can be turned on simultaneously, the filament may be turned off in stand-by service also, since a thoriated tungsten filament is ready to operate in less than one second after the voltage is applied.

In these days no one can afford the luxury of an experimental set-up or a slightly "hay-wire" condition in the circuits and power supplies of a vacuum tube transmitter. Accidental circuit failures and accidental failures of component parts, will often destroy the tube.

Avoid excessive grid drive. Excess grid drive (grid current) wastes driving power, and shortens the life of the driver tube by making it do extra work. Excess grid current also overheats the grid of the tube, and shortens its life either by damaging the grid permanently, or by increasing the number of watts the tube must dissipate.

It is essential that the rated filament voltages be maintained at the tube. This voltage should be measured at the socket, and should not deviate more than plus or minus 5 per cent. from the rated value.

The life of a thoriated tungsten filament will be reduced to 2/3 of normal if the filament voltage is permitted to run 10 per cent. above its rated value. At 10 per cent. below rated value, the emission from the filament may fall off due to failure to diffuse enough thorium to the surface of the filament to maintain emission. A drop in emission may cause severe overheating of the plate, with a consequent reduction in the life of the tube, or even complete failure.

Make good electrical connexions to the tube. At ordinary frequencies, the standard connector clips are satisfactory. At u.h.f. the charging currents into the inter-electrode capacities become large enough so that special care must be taken. A split connector of aluminium or plated brass, with the two halves held together by a silver or similarly plated external spring, which will remain good at 200 to 300 degrees C., will prove most satisfactory.

The electrical instability of r.f. circuits increases the probability of damage and over-load to a tube. Parasitic oscillations can also cause damaging overloads, as well as inconvenience. Nearly any parasitic oscillation can be prevented. A good way to isolate and cure parasitics in an amplifier is to:

(a) Remove the normal excitation.

(b) Remove all fixed bias.

(c) Lower the plate voltage until the plate loss due to the static plate current flowing does not exceed the rated tube dissipation.

Under these conditions there should be no parasitic oscillation at any position of the tuning dials. A parasitic oscillation will be readily indicated by the presence of grid current. If such oscillations occur, then

(a) Find the frequency of the parasitic.

(b) Determine the parasitic circuit superimposed on the normal r.f. circuits.

(c) Adjust the parasitic circuit decreasing its excitation voltage until the oscillation ceases. Such changes need not seriously affect normal circuits.

ABSTRACTS OF ELECTRONIC LITERATURE

RADIO

A Study of Propagation over the Ultra-Short Wave Radio Link Between Guernsey and England on Wavelengths of 5 and 8 metres.

(R. L. Smith-Rose, D.Sc., Ph.D., and (Miss) A. C. Stickland, M.Sc.)

Analysis of the data obtained has shown that while the fading of the signals was similar in type on the two wavelengths, there was a difference in the secular variation of the amount of fading on 5 and 8 m. No diurnal or seasonal variation of signal intensity was noted, nor was there any definite diurnal variation in the amount of fading. The fading observed took various forms, ranging from a rapid type in which the changes occurred more or less periodically every three or four minutes, to a slow type of variation extending over one or more hours. The latter type was usually accompanied by a high level of signal intensity interrupted at long intervals by rapid and deep fading lasting from 10 to 30 minutes. The field strength records show a marked correlation between periods of negligible fading and the existence of low atmospheric pressure conditions generally, including lack of appreciable temperature inversion. The occurrence of fog and snow has a similar effect to that due to the prevalence of low pressure.

—*Jour. I.E.E.* to be published.

Transients in Frequency Modulation

(H. Salinger)

In a frequency modulation system a sudden jump in carrier frequency corresponding to a Heaviside unit signal will result in a transient depending on the receiving-filter bandwidth. If this bandwidth exceeds twice the maximum frequency swing, the shape and duration of the transient is shown to be about the same as in an amplitude-modulation system with the same bandwidth for narrower filters the transient lasts longer. These results are applied to several practical cases. The transient is favourably affected by using an amplitude limiter and by arranging the filter pass band so as to enclose the maximum frequency swing symmetrically.

—*Proc. I.R.E.,* Vol. 30, No. 8 page 378.

The Technique of Frequency Measurement and its Application to Telecommunications

(J. E. Thwaites and F. J. M. Laver)

Rigid frequency control has become a necessity in radio broadcasting and indeed in all forms of telecommunication, and the art of frequency measurement has likewise become of great importance. The improvement in frequency standards which has taken place in recent years is briefly outlined in this paper, and some of the methods which may be used to obtain accurate frequency measurements in terms of the improved standards are described.

—*Jour. I.E.E.,* Vol. 89, Part 3, No. 7 (1942), page 139.

THERMIONIC DEVICES

The Use of Secondary Electron Emission to obtain Trigger or Relay Action

(A. M. Skellett)

The use of secondary electrons to obtain trigger action similar to that of a thyratron is described. An experimental tube and the necessary circuits by which this action is achieved are discussed. This combination gives the features of a triode with a relay or on and off feature, resulting in an amplifier, oscillator, modulator or other vacuum tube device which may be turned on or off abruptly at high or low frequencies. In addition, it can be used to replace thyratrons in many of their circuits where very low impedance is not necessary and is capable of much greater speeds of operation in such applications.

—*Jour. App. Physics.,* Vol. 13, No. 8 (1942), page 519.

THEORY

The Characteristic Curve of the Triode

(E. L. Chaffee)

A method is described by which the entire static characteristic curves for the anode current of a power triode can be deduced from one experimentally determined curve. This experimental curve can be obtained at low power without danger of overheating the valve. Using the same procedure, the grid current curves and the total space current curves can be determined. A new log chart is described on which the static characteristic curves of a triode are presented by means of straight lines and two curves giving the division of space current between grid and anode.

—*Proc. I.R.E.,* Vol. 30, No. 8 (1942), page 383.

Simplified Inductance Chart

(E. S. Purington)

A reference sheet in which the inductance and Q of solenoidal or multilayer coils wound on a cylinder are readily determined when the dimensions of the coil are known. Effect of insulation in reducing space factor are considered together with skin effect considerations.

—*Electronics,* Vol. 15, No. 9, p. 61.

Variation of the Axial Aberrations of Electron Lenses with Lens strength

(E. G. Ramberg)

The variation of refractive power and spherical aberration with electrode voltages and field strengths is studied for two characteristic unipotential lenses, an immersion lens, and a magnetic lens. Conclusions are drawn herefrom regarding the variation, with lens strength and applied voltage of the resolving power obtainable with the lens as an electron-microscope objective. Scattered measurements by other authors agree satisfactorily with the results. The " relativistic aberration " of the electrostatic unipotential lenses, *i.e.,* the effect on the image of fluctuations in the over-all applied voltage, is calculated and shown to be of significance in the electrostatic electron microscope. Furthermore, the axial chromatic aberrations are computed for the four systems and the question of upper limits of the last two aberrations is discussed.

—*Jour. App. Phys.* Vol. 13, No. 9 (1942), page 582.

The Theory of Units

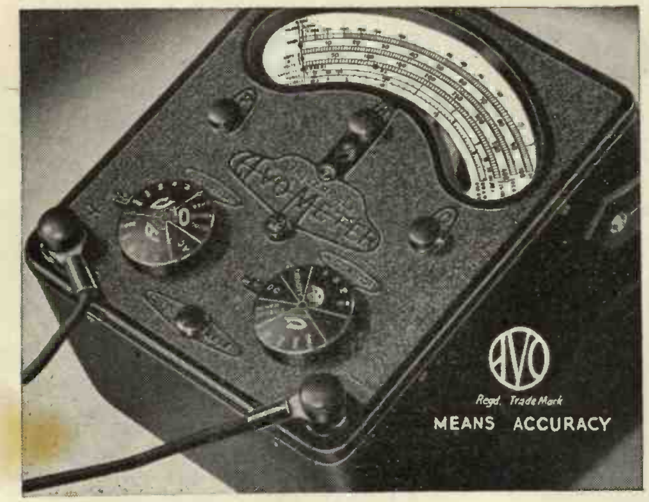
(L. H. Bedford)

In the formulation of electromagnetic theory there are a three stages at which arbitrary constants are conveniently introduced for the purpose of defining units. It is customary to assign immediately the value unity to certain of these constants, after which they are lost sight of. In the present treatment, a sketch of electromagnetic theory is given in which the suppression of these fundamental constants (k_1, k_2, k_3) does not occur. Maxwell's theory is shown to lead to a certain relationship between the k's involving the velocity of light. Subject to this restriction, the assignment of k-values is an arbitrary matter, the process of formulating a unit system being one of 2 degrees of freedom.

A table is constructed showing the assignment of k-values corresponding to the various known unit systems. This table can, amongst other things, be used to derive the numerical relationship between the unit quantities of the various systems.

Serious inconsistencies of method and nomenclature in connexion with the practical system of units are brought to light and proposals for rationalisation are considered.

—*Jour. Brit. I.R.E.* to be published.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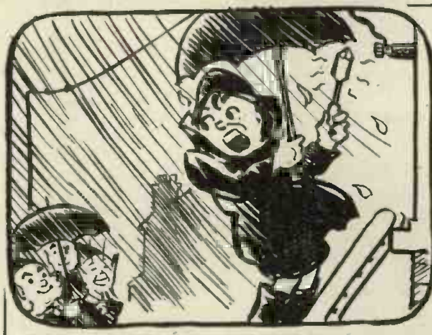

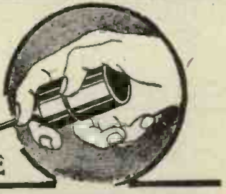

PATENTS RECORD

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INDUSTRY

Electric Motor-Control Systems

Relates to control systems for controlling the speed of an electric motor with respect to a reference speed, especially for machines operating on a length of material such as a web of wood veneer, paper or cloth.

In veneer stripping or cutting machines in which the veneer is stripped from a log and run out on a motor-driven roll conveyor the diameter of the log decreases as the veneer is stripped off. Therefore the linear speed at take-off decreases continuously even though the log is rotated at a constant speed. If the conveyor continues to run at a given speed undue tensioning and perhaps breakage will occur.

In the application of the invention to veneer stripping apparatus, the first of the pilot generators is mechanically coupled to the web as it leaves the log, *i.e.*, at the point where the linear speed of the web is decreasing. The second pilot generator is coupled to the driving roll which determines the speed of the web on the conveyor. The pilot generators are connected to a common output circuit so that the respective output voltages oppose each other. If the voltages vary with respect to each other because of a change in speed of one or the other of the generators, a circulating current, which is proportional to the speed difference, flows in the saturating winding of a reactor.

The resulting change in impedance increases or decreases the output of a valve circuit to correspondingly increase or decrease the field excitation of the main generator supplying the driving motor. Thus, the speed of the conveyor motor is raised or lowered to give a uniform linear speed of the web.

A simple anti-hunting circuit is provided to prevent over-correction of the generator voltage output and driving speed.

—*The British Thomson-Houston Co., Ltd. Patent No.* 540,035.

CATHODE-RAY TUBES

Improved Electron Camera

To overcome the disadvantages of a discontinuous (mosaic type) photo-emissive electrode and to avoid the distortion introduced by a double-sided mosaic it is proposed to make an electron camera in the form of a tube with a semi-circular extension tube, at the far end of which is mounted a flat semi-transparent photo-cathode on which the optical image is focused.

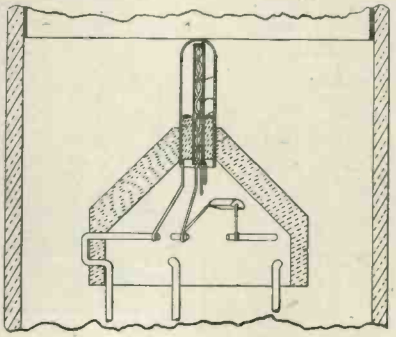
At the end of the curved tube nearest the electron gun is a target electrode of mosaic structure offset from the axis of the beam from the electron gun.

A coil surrounds the whole envelope, the toroidal portion serving to guide the photo-electrons round the curved portion of the tube on to the mosaic, where they produce an electron image.

Electrons from the gun pass through an aperture at the side of the mosaic plate and passing round the curve tube, are reflected back from the photo-cathode which is at a high negative potential to the source of electrons. These electrons are then caused to scan the mosaic on the same side as that on which the electrostatic image is formed and so discharge the mosaic in the usual way.

—*Marconi's W.T. Co. and H. A. Iams. Patent No.* 546,519.

Improved Cathode Ray Tube



To provide a cathode-ray tube of simplified and improved construction especially suitable as the image reproducer in television systems.

This comprises a thermionic cathode, a first electron lens consisting of a cylindrical apertured grid and a cylindrical first anode. The grid has an external diameter less than 0.15 times that of the internal diameter of the anode. The grid cathode are preferably as small as manufacturing facilities permit, the cathode electron emission being at least as high as 50 milliamperes per square centimetre of emissive surface. The decrease in the area of the electron-emissive surface of the cathode results in a substantial reduction of the heater wattage.

An additional advantage is that the grid to cathode capacity is materially decreased by the relatively small dimensions of the electron gun components, and is claimed to be of the order of one or two $\mu\mu$F.

—*Hazeltine Corp. (Assignees of R. C. Hergenrother). Patent No.* 546,846.

December Meetings

I.E.E. Wireless Section

On Wednesday, December 2, at 5.30 p.m. in the Lecture Theatre of the Institution of Electrical Engineers, Savoy Place, W.C.2, a paper on "The Electrical Amplifying Stethoscope and Phono-electrocardioscope" will be read by Dr. G. E. Donovan, M.Sc., M.B., B.Ch.

Institute of Physics—" Newton Tercentenary "

A meeting of the Branch will be held in the Lecture Theatre of the Royal Institution, Albemarle Street, Piccadilly, London, W.1, on Wednesday, December 9, at 5.0 p.m., to celebrate the three-hundredth anniversary of the birth of Sir Isaac Newton. Papers will be read as follows:—

(a) " Life of Newton," by H. Lowery, D.Sc., F.Inst.P.
(b) " Some Historical Aspects of Newton's Published Works," by H. F. Buckley, D.Sc., F.Inst.P. (National Physical Laboratory).

This meeting, which is of general interest, will be open to visitors.

The Television Society

The annual general meeting of the Television Society will be held on Saturday, December 5 at 2.30 p.m. in the small theatre of the Institution of Electrical Engineers, Savoy Place, W.C.2.

At the conclusion of the meeting (which is for members only) a short review of " Progress in Colour Television " will be given by the Hon. Lecture Secretary, Mr. G. Parr.

An informal reunion will be held at the conclusion of the meeting, which non-members may attend by invitation only.

I.E.E. Students Section

At the next meeting to be held on December 7 at 7 p.m., a paper will be read by Dr. D. S. Anderson on " The Failure of the Technician in his Role of Citizen."

Tea will be served from 6.30 p.m.

On December 12 a dance will be held at the Lysbeth Hall, Soho Square, W.1, from 3-7 p.m.

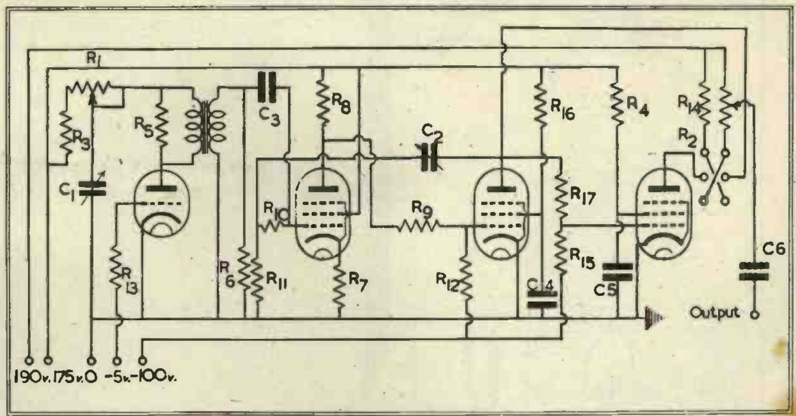
Tickets, price 2s. 6d. single, 4s. 6d. double, can be obtained on application to Mr. A. N. Lord, 8 Hadden Way, Greenford, Middlesex.


RELIANCE

A Pulse Generator for Circuit Testing

A USEFUL pulse generator circuit is described by H. P. Manning and V. J. Young in the May issue of *The Review of Scientific Instruments.* It produces a pulse which may be conveniently varied in duration, voltage and frequency so as to simulate actual operating conditions in a Geiger counter or similar apparatus. It is also useful for checking amplifier response.

C_1 and R_1 make possible a continuous frequency range from 0.8 to 30,000 pulses per sec. C_2 allows a choice of pulse breadths varying from 40 microseconds to 3 milliseconds. R_2 varies the pulse voltage from zero to 150 volts. The reversing switch changes the polarity of the pulse without changing its shape or disturbing the ground connexions. All five controls are essentially independent of each other. The valves are (from left to right) 885; 6C6; 42; 42.



Values of Components.

R_1	1-megohm potentiometer
R_2	5000-ohm potentiometer
R_3	500,000 ohms
R_4	15,000 ohms
R_5	1,000 ohms
R_6	50,000 ohms
R_7	10,000 ohms
R_8	100,000 ohms
R_9	1 megohm
R_{10}	100,000 ohms
R_{11}	100,000 ohms
R_{12}	500,000 ohms
R_{13}	25,000 ohms
R_{14}	50,000 ohms
R_{15}	600,000 ohms
R_{16}	12,500 ohms
R_{17}	1 megohm

C_1	2 μf 1 μf 0.4 μf 0.15 μf 0.065 μf 0.03 μf 0.01 μf 0.003 μf 0.0007 μf	9-position switch
C_2	0.01 μf 1450 μμf 500 μμf 250 μμf < 100 μμf	5-position switch
C_3	0.001 μf	
C_4	= C_5, 0.5 μf	
C_6	= 0.05 μf	

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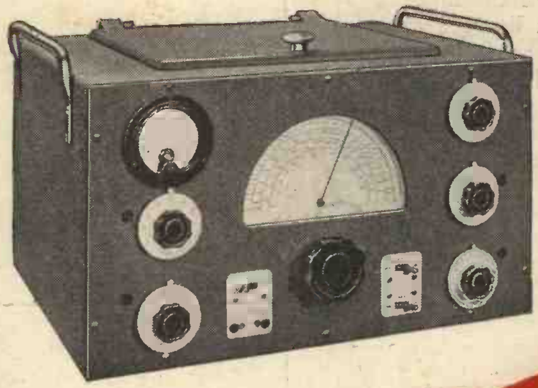

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