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INTERNATIONAL CONFERENCES OF CAIRO, 1938

BY

C. B. JOLLIFFE
Engineer-in-Charge, RCA Frequency Bureau

THE regulation of all forms of radio which are capable of causing international interference is based on the International Telecommunications Convention signed at Madrid, Spain, December 9, 1932, by practically all nations of the world. This treaty was ratified by the United States June 27, 1934. The International Telegraph Convention of St. Petersburg, 1875, and the International Radio Convention of Washington, 1927, were replaced by the International Telecommunications Convention of Madrid, 1932.

By means of four sets of regulations, General Radio Regulations, Additional Radio Regulations, Telegraph Regulations, and Telephone Regulations, which are authorized and annexed to it as separate documents, the Madrid Convention combines under one document all regulations relating to electrical communications. A nation may accept one or more of the regulations as it desires, provided it also adheres to the Convention. The United States is a party to the Convention and the General Radio Regulations. The Convention proper contains the fundamental principles of regulation which are not subject to frequent change and it is, therefore, the basic document. It is subject to change only by a conference of properly accredited representatives which would be convened upon request by twenty or more nations. It is not expected that it will be necessary to change this basic treaty for many years.

It is provided in the Convention, however, that each set of regulations which are administrative in scope and which must be changed from time to time because of developments in the respective fields, will be reviewed at intervals of approximately five years by administrative conferences of representatives of governments which are parties to the regulations.

The next administrative conferences for the review of these regulations are to be held at Cairo, Egypt, beginning February 1, 1938. Two conferences meet simultaneously; the International Radio Conference which will consider the revision of the General Radio Regulations and the Additional Radio Regulations, and the International Telegraph and Telephone Conference which will consider the revision of the International Telegraph Regulations and Telephone Regulations. The two conferences will be run concurrently and for the most part representatives of the governments to one conference will be the same as the representatives to the other conference.

The United States Government, being a party to the General Radio Regulations, is authorized to be represented in that conference in full capacity, while in the conference dealing with telegraph and telephone regulations its delegation has the status of observers. Representatives of radio operating companies are admitted in an advisory capacity to both conferences, but are not permitted to vote on any question. In past conferences the delegations of the United States have received full cooperation from the representatives of the radio operating companies of the United States.

The official language for all documents of the conference is French, and the French draft is the official text of the final documents. For the purpose of discussion in the conference the use of both French and English is authorized by the Convention, a speech made in one language being immediately translated into the other. This enables persons attending the conference to understand the deliberations and take part in all of the discussions.

The secretarial work incident to the operation of all international conferences on communications is carried on by the International Telecommunications Bureau, the offices of which are located at Berne, Switzerland. This organization, which has been in existence for many years as the Bureau of the International Telegraph Union, is continued by the Convention and has been very useful not only in this capacity, but also in its major capacity of providing a central clearing house for information concerning all communication matters.

The preparatory work for the Cairo conferences has been going on for several months. All the nations who desired have submitted their proposals for changes in the various regulations. These proposals have all been combined in a single volume for radio and a single volume for telegraph and telephone by the International Telecommunications Bureau and these volumes circulated throughout the world. These proposals constitute the basis of an agenda for the meetings to be held at Cairo. Each country uses these as a basis for study prior to the departure of the delegations for the conference and they are used as bases of discussion by the conference itself. The countries are not precluded from making additional proposals and supplements to the books of proposals are issued from time to time, and proposals will probably be made by the various delegations during the course of the conference.

The principal problem which the Radio Conference will have to face is that of allocation and use of frequencies by the radio stations of the world. The allocation of bands of frequencies to services is the heart of the General Radio Regulations. The present allocation, which was arrived at after much deliberation at the conference at Washington in 1927, was changed in only minor particulars at the conference of Madrid in 1932. Since the original allocation of 1927, radio has developed and expanded, especially in the use of high frequencies for long-distance international services. New services have developed and older ones have changed.

Many nations have become conscious of the usefulness of high frequencies (6,000 to 21,000 kc) for the purpose of broadcasting to persons living at great distances. Many of the receiving sets now in use permit listeners to hear broadcasting from all parts of the world. This service was practically unknown in 1927 and did not reach its present development until about 1934. The amount of frequency space which was originally allocated to broadcasting was relatively small and with its development the number of stations which have been crowded into this small frequency space has resulted in a great amount of interference, which threatens to destroy the service. Some countries, not finding interference-free space within the broadcast bands, have assigned frequencies to broadcasting stations in the bands allocated for fixed service or mobile service, thus causing interference to long-distance international services in these bands. The Cairo conference, or some subsequent conference, must find a solution to this problem so that broadcasting can be properly regulated and made useful on high frequencies without destroying other very important and useful services.

The entire use of the higher frequencies for aeronautics, both in the United States and in other countries, has developed since the 1927 allocation, and radio has proved itself a necessary adjunct to aviation. Because of the dependence on radio and the high speed of travel of modern airplanes, it is necessary that radio for aviation be given primary consideration. Whether this may be done best by providing exclusive bands of frequencies for aviation or fitting the needs of aviation into the bands allocated for mobile services in order that the operation of aircraft, particularly over the sea, may be coordinated with that of surface craft capable of giving them aid in time of distress, is one problem which must be given careful consideration and to which an answer must be found.

It is of primary importance that ships at sea be able to communicate with each other and with shore stations all over the world for the exchange of messages and in order to insure the safety of life and property at sea. Consequently the operation of maritime radio stations must be under international regulations. The first radio conference in 1903 recognized this principle and since that time the General Radio Regulations have governed how the radio services of ships must be conducted for the exchange of messages, also how operations must be carried on in time of distress in order to minimize interference and insure the best communications possible under very adverse circumstances. The distress signal is specified by the regulations and given the protection which it deserves. Experience has shown that the regulations surrounding distress have in general functioned properly and there are few proposals for changing this operation.

The General Radio Regulations of 1927 established an alarm signal which was required to be sent in advance of the distress signal. The purpose of this signal is to operate automatic devices, installed on other ships, which may be in the vicinity of the ship in distress and call the operators on these ships. This makes possible the participation in distress cases of many ships which are unable to maintain a continuous radio watch and thus might miss a distress call entirely. Experience with this alarm device has shown it to be very satisfactory and the regulations are so framed as to give this device proper protection.

Amateur radio is a service which is considered quite important by the United States Government for various reasons and the United States has always taken the lead in protecting the frequencies assigned to this service. The Radio Regulations have a minimum amount of regulation of the amateurs, leaving each country free in most respects to apply such regulations as it sees fit. The American amateur has justified the freedom which he has been given in the use of the frequencies and undoubtedly the United States will again endeavor to protect this service against inroads by other nations.

In order that operators on board ships which go into foreign ports may not be subject to individual examination or that their qualifications may not be questioned, the Convention sets forth in considerable detail the requirements for the examination of radio operators. This portion of the regulations is always subject to careful scrutiny to be sure that the operators on whom may devolve great responsibility for the safety of a large number of persons in times of distress may be properly qualified to operate radio equipment. The basic requirements have become practically uniform throughout the world and the operators are capable of communicating from ship to ship regardless of the nationality of the operators.

One of the primary requirements in the operation of radio service in a medium which is used in common by all the nations of the world is that all nations use equipment that is built and operated in accordance with good engineering principles. It is quite difficult to determine in an administrative conference, which is invariably pressed for time, what is good engineering practice and accordingly the regulations have set up the International Radio Consulting Committee (CCIR) for the purpose of studying technical and allied problems. This conference meets between the administrative conferences and studies various technical problems which are submitted to it by the various administrations. The answers to some of these problems are simply informative in nature and require no additional regulation. Other answers, however, are recommendations for regulations and need to be considered by the administrations from the standpoint of whether or not they should be inserted in the General Regulations. An example of this latter type is the tolerance which will be permitted to a station when operating on an assigned frequency. The last meeting of the CCIR which was held at Bucharest in 1937 recommended that there be inserted in the regulations certain permitted tolerances for various classes of stations. It now becomes the duty of the administrative conference to decide whether or not it desires to accept or reject this recommendation or modify the proposal.

The United States Government is not a party to the International Telegraph Regulations and these regulations are not applied by the Government to the communication companies. However, the communication companies which operate circuits jointly with administrations or companies of countries which are parties to the Telegraph Regulations must, in general, apply the provisions of these regulations in order to provide uniformity in operations. There have been many proposals that the United States become a party to the Telegraph Regulations and participate actively in the conferences which formulate them. However, up to the present the United States has not become a party to the regulations although its representatives have participated in many of the conferences as observers.

While these regulations do not establish the basic rates between two countries they do set up the relation between the rates of various classifications of messages and the basic rate and the regulations under which the various classifications of messages are handled. These relationships between the rates for the various classifications of messages and the regulations concerning their handling are extremely vital to American communication companies and to American business firms who provide a large portion of the international communication business. The United States submitted as a proposal to the Telegraph Conference in Cairo a "Proposal in Principle" which would govern the basis for adherence or non-adherence of the United States to the Telegraph Regulations. The following is a quotation of a portion of this Proposal:

> "An examination of the International Telegraph Regulations discloses that they contain certain provisions, such for example, as those of Article 10, the subject matter of which is of interest to the Government of the United States in view of its increased regulatory activities. On the other hand, they contain numerous provisions, such for example as those of Article 35, having to do with operation and management of the communication services, by which the Government of the United States would not consider it proper to bind itself.
>
> "If a satisfactory segregation of the provisions of the International Telegraph Regulations into these two classes be effected, the Government of the United States can give serious consideration to the signing and acceptance of those that are of direct interest to this Government, subject to modifications and changes in existing language designed to meet conditions in the United States. This segregation could be accomplished by drafting two sets of telegraph regulations corresponding to the two sets of radio regulations. If this be not practicable, the segregation may be accomplished by arranging the provisions of the telegraph regulations so that those the Government of the United States can accept are placed together in a first part and the remaining articles in a second part. If the latter alternative is adopted, a separate protocol could be drafted in accordance with which the Government of the United States might accept only those provisions contained in the first part of the telegraph regulations.
>
> "The Government of the United States recognizes that conditions in other countries may make it unnecessary or undesirable to observe this distinction within their own jurisdictions, and to this, of course, it has no objection; nor can it object to the inclusion of provisions not of concern to the Government of the United States, such as those applicable to the European Regime, in the portion of the regulations which it does not sign."

This is a basic proposal concerning not only the form, but also the applications of the Regulations and will, of course, be considered carefully and discussed fully by the representatives of the nations which are parties to the Telegraph Convention.

As in the past it is expected that approximately eighty nations will participate in these conferences and that after much discussion and many compromises final treaties will be formulated to which the nations will adhere with or without reservations. Insofar as the United States is concerned ratification of these regulations must be approved by the Senate, and when ratified they become the basis on which national regulation of international communications must be built.

The development of international regulation of electrical communications over a period of many years has kept pace with the development of communication and each international conference has brought the regulation into line with development. This work has been constructive and has always resulted in stimulating the development of international communications.

DIRECT-VIEWING TYPE CATHODE-RAY TUBE FOR LARGE TELEVISION IMAGES*

By

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Summary—*A new device for obtaining large bright television images of high contrast and high definition has been developed at the Camden Laboratories of the RCA Manufacturing Company. It is a direct-viewing cathode-ray tube 4½ feet long and 31 inches in diameter. It is of the continuously evacuated type and gives a picture 18 by 24 inches in size. The paper describes the design and construction of the new tube, the reasons for the development, the difficulties which were overcome, and the results obtained.*

EVER since high-definition television pictures were first demonstrated, newspaper writers and laymen have commented on the small size of the picture. Seemingly it has been of little interest that the size of the picture has had little to do with the amount of information communicated. In early work on high-definition systems a 9-inch diameter cathode-ray tube was used to produce a picture approximately 6 by 8 inches. Most of the present direct-viewing cathode-ray tubes are 12 inches in diameter and produce pictures approximately 7½ by 10 inches. Even so, larger pictures are wanted. Consequently, a great amount of effort and money has been spent here and abroad in the quest for methods of producing large television images having adequate brightness, contrast, and definition.

Many solutions to the problem of obtaining large television images have been proposed and several methods have been extensively explored. Interesting demonstrations have been given here and abroad. Frequent mentions of the projection cathode-ray tube method and also of the supersonic light-valve method are made in the current technical news.

The purpose of this paper is to describe another method of obtaining large television pictures, namely, the method of large direct-viewing cathode-ray tube development. This tube was built with the primary purpose of studying television pictures of large size (18 by 24 inches) under conditions where brightness, contrast, and definition were adequate and where the method of reproduction did not limit the performance of the system.

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* First presented before Fall Meeting of IRE, Rochester, N. Y., November 9, 1937.

The most important consideration in favor of the large direct-viewing cathode-ray tube is that the total amount of light obtainable from a luminescent screen is directly proportional to the area of the screen. This point will be clarified further.

At present the most widely used luminescent materials for screens in cathode-ray tubes are: the zinc orthosilicate (willemite) and the zinc sulphide. Both materials exhibit the property known as "current saturation." A current-saturation curve of a yellow willemite screen, bombarded by 10,000-volt electrons in a developmental projection tube, is shown in Figure 1.

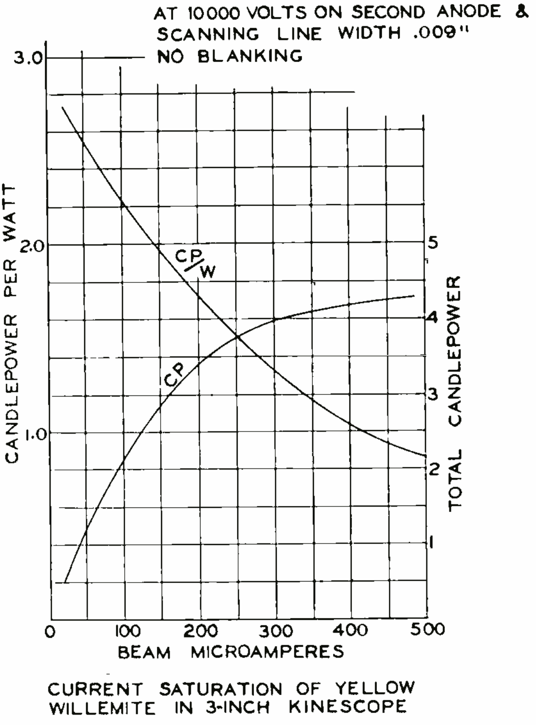

Fig. 1—Current-saturation of a willemite screen.

Measurements show that under the conditions of normal television scanning this saturation is a function of the area of the scanning spot and not of the total scanned area. But the area of the scanning spot is necessarily a function of the total area, if the detail of the picture is to be preserved; i.e. it cannot be larger than a certain fraction of the total area scanned. In actual practice, since the luminous spot is round, a certain overlap of the scanning lines is permissible.

As a limit, after which a serious loss of detail takes place, 50 per cent overlap may be taken. The present tentative standard calls for 441 lines per frame, about 10 per cent of which are blanked out during vertical synchronizing time. The observed picture, therefore, consists of 400 horizontal lines. Allowing 50 per cent overlap this calls for the line width of one-half of one per cent of the height of the reproduced picture as the limiting maximum line width.

It may be deduced from the curves of Figure 1 that at 10,000 volts the maximum useful brightness of this particular type of luminescent screen is 0.7 candlepower per square inch or 100 candles per square foot. The maximum useful beam current (while it is on) is 58 μa per square inch, but when the average power over a period of one complete white frame is considered, it is only 0.80 of the product of volts and amperes (max.).

Fig. 2—12-inch direct-viewing television cathode-ray tube for large light output at high contrast.

The factor of 0.80 is introduced because in actual operation the electron beam scans a given picture area for only 80 per cent of the time since 20 per cent of the time it is extinguished for the line and frame returns or fly-backs.

As to the minimum required brightness of the screen, opinions vary greatly. As a yardstick, the brightness of a motion-picture screen is often used. A committe of the Society of Motion-Picture Engineers concludes that the high-lights of the picture should have at least 11-foot lamberts or 3.5 candles per square foot if eye fatigue is to be completely avoided.* The recommendation, however, is that 0.86 to 1.65 candles per square foot be adopted as a temporary standard.

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* *Jour. SMPE,* Vol. 26 (May 1936) and Vol. 27 (Aug. 1936).

There is very good reason to believe that a television picture should have more light than that. The author's experience indicates that at no time has he seen a television image that was too bright in a normally lighted room. With the tube shown in Figure 2, with 1.1 ma in the beam at 10,000 volts on the second anode, high-lights of 40 candles per square foot were obtained. The picture was bright and permitted demonstrations in a brightly illuminated room, but no observer pronounced the picture as being too bright. In a dark room such a picture is definitely too bright.

The reason for low screen brightness being satisfactory for motion-picture theaters is that there is practically no stray light and the size of the image is very large. The theater hall is devoted to the showing of pictures and everybody there is looking at the picture. The television receiver is placed in a room which is used for other purposes. It may be the living room of a residence, a hotel lobby, or a restaurant. To be of maximum usefulness, a television receiver should not interfere with any other functions of the room. The willemite screen by itself, at 10,000 volts, is capable of giving a surface brightness as high as 100 candles per square foot or 314 foot-lamberts or apparent foot-candles. For a screen 18 by 24 inches it would require 25 ma at 10,000 volts. For the previously mentioned figure of 40 cp per square foot, only 6 ma at 10,000 volts are required. The lower the current

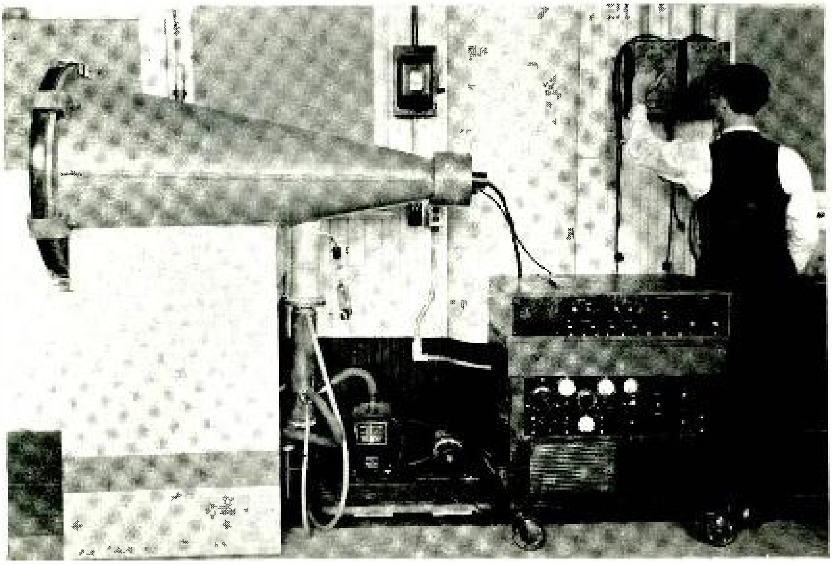

Fig. 3—Side view of 31-inch demountable TCR tube.

density of the luminous spot, the higher is the screen efficiency. At 2 ma and 10,000 volts a directly bombarded luminescent willemite screen of the type described will have a brilliancy of 14.6 cp per square foot or 46-foot lamberts which is nine times the upper brightness limit of the tentative SMPE standard.

During the first quarter of the present year the construction of a direct-viewing TCR tube with screen 18 inches by 24 inches was completed at the Camden Laboratory of the RCA Manufacturing Company, Inc. The tube is of the demountable, continuously-evacuated type and has a metal envelope with a Pyrex sight glass. Figure 3 shows a side view of this tube. The envelope is made of good grade steel ¼-inch thick with arc-welded seams and flanges.

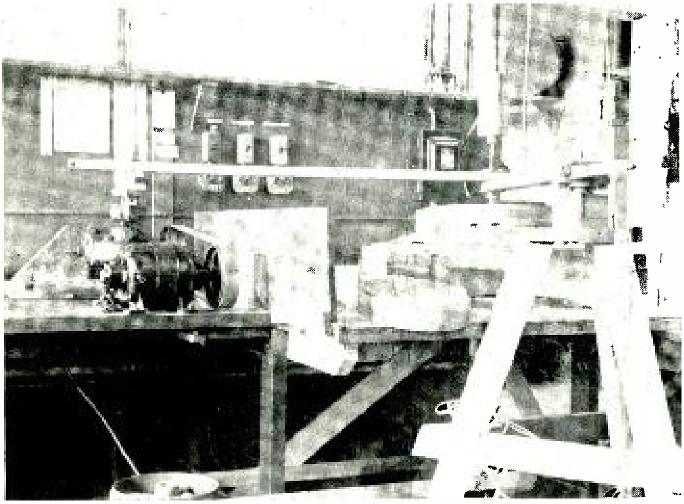

Fig. 4—Machine for grinding and polishing glass covers for 31-inch TCR tube.

It has the shape of a cone, and is 4.5 feet in length. The outside diameter of the larger flange is 31 inches. A three-stage oil-diffusion pump is directly connected to the tube through a special outlet. For fore-vacuum, a mechanical vacuum pump is connected to the diffusion pump by means of a length of rubber hose. The glass cover is convex outward, 31 inches in diameter and 2 inches thick. This thickness is required because the total atmospheric pressure on the glass is approximately 5½ tons. A special machine was constructed in the laboratory for grinding and polishing both surfaces of the glass. The technique used was that of grinding telescope lenses. A layout of the grinding machine is shown in Figure 4.

For vacuum-tight joints between the glass and metal as well as between metal flanges, pure gum rubber gaskets proved very satisfactory. The performance of the tube is quite satisfactory when vacuum of the order of 10^{-6} mm Hg is reached. Normally such a vacuum is reached after 48 hours of operation. The vacuum measurements are made by means of thermocouple and ionization gauges attached to the sleeve connecting the vessel and the diffusion pump.

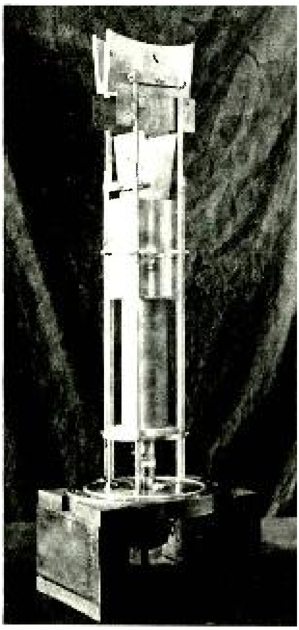

Fig. 5—Electron gun for 31-inch demountable TCR tube.

The tube was designed for 10,000 volts on the second anode. For safety reasons, instead of operating the metal envelope at 10,000 volts positive, it is grounded and the cathode is raised to the same voltage, but negative. This arrangement greatly facilitates the construction of the electron gun. The electron gun used in this tube is shown in Figure 5. It gave beam currents as high as 8 ma at 10,000 volts with corresponding brilliancy of the high-lights. However, the best overall performance was obtained with a gun giving 2 ma in a narrow beam with negligible defocusing and with -150 volts cut-off grid voltage.

The design of the power supply and video amplifier for the demountable tube offered many difficulties. The cathodes in the last stages of the video amplifier had to be operated at minus 10,000 volts and, of course, had to be capacity-coupled somewhere along the chain to the low-voltage stages. The two coupling condensers during the operation are charged to 10,000 volts and at the same time are required to pass low video frequency currents. All the meters and controls on the last stages of the amplifier had to be insulated for 10,000 volts. A view of the portable outfit containing the video amplifier, synchronizing and deflecting circuits, and high and low-voltage supply, is shown on the right-hand side of Figure 3.

A typical received picture is shown in Figure 6. The signal was taken from an Iconoscope pickup of a regular moving picture frame.

Fig. 6—Unretouched photograph of received television picture on demountable TCR tube.

The photograph has not been retouched. It will be noted from the photograph that the sides of the image are straight and there is no apparent bulging of the image. The reason for this effect is that the 2-inch thick glass disc is used only as vacuum cover or a sight glass while the luminescent material is deposited on a flat glass sheet ¾-mm thick, which is fastened to the walls of the tube. The flat appearance of this type of luminescent screen is not its only advantage. The fact that it is flat greatly improves the overall contrast of the reproduced picture. On a concave screen, illuminated parts throw light directly on the blacks of the image, thereby reducing the contrast. The fact that the screen glass is thin improves the contrast in details by reducing the well-known "halation" or "the spurious ring" effect.

In conclusion it may be stated that with the tube described, large, bright television images of high detail and of high contrast are obtainable. This tube possesses real entertainment and communication value and may be shown in normally lighted rooms in day-time and at night.

The author acknowledges the valuable help and cooperation of his associates in carrying out the developments described, especially Dr. D. W. Epstein and Mr. K. R. Wendt, both of the General Research Division of the RCA Manufacturing Company, Inc. at Camden, N. J. The 31-inch glass disc was ground and polished with the assistance of Mr. D. D. Landis of the Photophone Division of the same company.

TELEVISION CATHODE-RAY TUBES FOR THE AMATEUR

BY

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THE work of amateurs and experimenters has always been important to radio progress. This is largely because many amateurs are highly endowed with a Missourian point of view and a driving curiosity which leads them to prefer the byways and unknown regions of their hobby rather than the beaten paths. Even though the individual findings of many in this group may be small, the total contributions and accumulated experience of all the radio amateurs and experimenters in the United States have been impressive.

That the radio amateur group has perhaps not yet contributed much to modern television progress is in no way due to lack of ability or resourcefulness on its part, but rather to the fact that this field of investigation requires devices outside the scope of the home workshop.

Of these devices, cathode-ray tubes suitable for television reception are perhaps the most important. Now that such tubes are available to the amateur, one of the major obstacles to his participation in experimental television activities is removed. The fact that the way of the amateur television experimenter is not easy should not be minimized. Even the hardy amateur who likes to pioneer, to build complex apparatus, to tear down and build anew, and who lives in one of the few favored areas which have television transmitters, should recognize that the transmissions in these areas do not as yet have a common standard and that transmitters are often off the air for long periods while changes are being made.

A photograph of two new cathode-ray tubes suitable for television, and known as Kinescopes, is shown in Figure 1. Both of these tubes are of the electromagnetic-deflection type and utilize a screen material which fluoresces brightly with a yellowish hue. The larger tube, RCA-1800, has a bulb-end 9 inches in diameter and accommodates a picture 5½ inches by 7⅜ inches in size, while the smaller tube, RCA-1801, has a bulb-end 5 inches in diameter and can show a picture 3 inches by 4 inches.

The electron gun in each of these tubes has been especially designed to give the small spot size required for high-definition television reception. The conical portion of the bulb of each tube is coated on the inner surface with a conducting material to prevent distortion of the beam and spot at the outer edges of the picture area. The maximum anode No. 2 voltage for the 1800 is 7000 volts and for the 1801, 3000 volts. Lower voltages can be used where it is desired to economize on power-pack cost, but at some sacrifice in either picture definition or brightness.

A diagram of a voltage-supply circuit for the 1800 is shown in Figure 2. This diagram gives suitable circuit constants. A bleeder current of two to five milliamperes is adequate. Adjustment of electrode voltages is provided by potentiometers in the bleeder circuit. The video signal and background-control bias are introduced between grid No. 1 and the cathode of the cathode-ray tube.

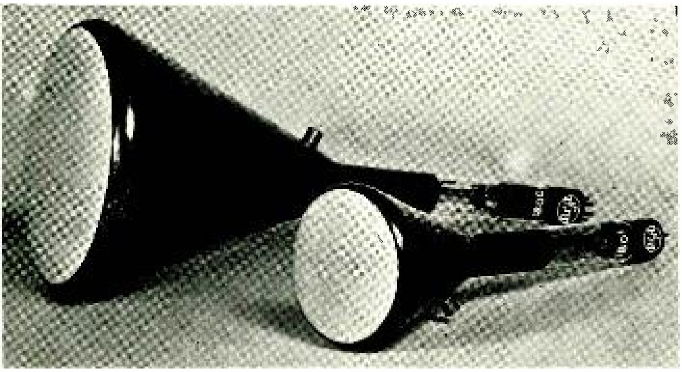

Fig. 1—Photograph of Kinescopes RCA-1800 and RCA-1801.

A typical vertical-deflecting circuit for television reception is shown in Figure 3. This circuit generates a saw-tooth deflecting current in the following manner. In the 6N7 tube, the triode unit shown on the left operates as a blocking oscillator. Oscillations are started in the circuit by the feedback action of transformer *T*. The flow of grid current which accompanies these oscillations causes a negative bias voltage to build up across the condenser C_1 and the grid leak consisting of resistors R_1 and R_2. After a very few oscillations, this bias voltage becomes sufficiently large to cause the plate current of the triode unit to cut off and, by so doing, to stop oscillations. When the oscillations cease, the charge on C_1 leaks off through R_1 and R_2 to a value such that the circuit can resume oscillation. The tube then blocks again, and the cycle repeats.

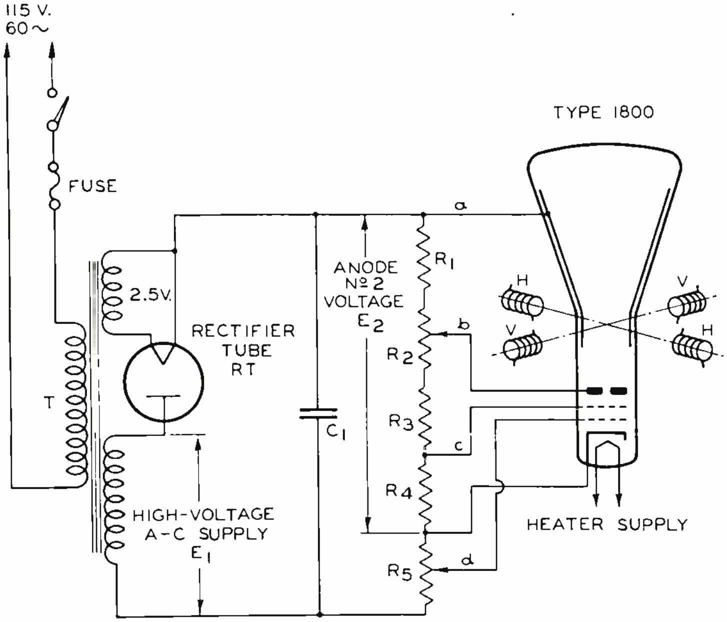

Fig. 2—Voltage-supply circuit for RCA-1800.

CIRCUIT COMPONENT	SPECIFICATIONS		
E_2	3000 VOLTS	4500 VOLTS	6000 VOLTS
C_1	2 μf, 3000 V.	2 μf, 4500 V.	2 μf, 6000 V.
R_1	1 MEGOHM, 6 WATTS	1 MEGOHM, 12 WATTS	1 MEGOHM, 25 WATTS
R_2	100000 OHMS, 1 WATT	100000 OHMS, 2 WATTS	100000 OHMS, 3 WATTS
R_3	135000 OHMS, 1 WATT	150000 OHMS, 2 WATTS	165000 OHMS, 4 WATTS
R_4	85000 OHMS, 1 WATT	70000 OHMS, 1 WATT	55000 OHMS, 2 WATTS
R_5	12000 OHMS, 0.5 WATT	10000 OHMS, 0.5 WATT	10000 OHMS, 0.5 WATT
E_1	2300 VOLTS RMS	3400 VOLTS RMS	4600 VOLTS RMS
RT	TYPE 879	TYPE 878	TYPE 878

a = ANODE No. 2
b = ANODE No. 1
c = GRID No. 2
d = GRID No. 1

H = HORIZONTAL-DEFLECTING COILS
V = VERTICAL-DEFLECTING COILS
T = POWER TRANSFORMER

Because the grid of the right-hand triode unit is connected to the grid of the first triode unit, the d-c plate current of the former unit rises suddenly to a large value during the period of oscillations. When oscillations stop, the d-c plate current becomes zero and remains zero until a new cycle starts. With the sudden increase in d-c plate current, the plate voltage drops abruptly because of the loss of voltage in resistors R_4 and R_5. During the period that the d-c plate current is zero, the plate voltage increases at a relatively slow rate, the rate being limited chiefly by the time constant of C_3, together with R_4 and R_5. Because the plate voltage thus goes through an abrupt decrease followed by a relatively slow increase, its variation has a saw-tooth form. This voltage applied to the grid of the 6C5 through C_4 and R_7 produces a saw-tooth current in the plate circuit of the 6C5. In this manner, saw-tooth current is caused to flow in the vertical-deflecting coils.

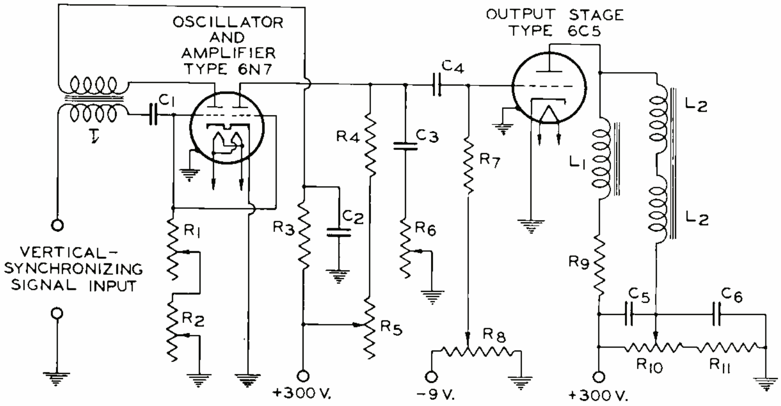

Fig. 3—Vertical-deflecting circuit for RCA-1800 or RCA-1801.

C_1, C_3 = 0.1 μf
C_2, C_4 = 0.25μf
C_5, C_6 = 10 μf
R_1 = 200000 OHMS } VERTICAL-SPEED
R_2 = 10000 OHMS } CONTROLS
R_3 = 100000 OHMS
R_4 = 2 MEGOHMS
R_5 = 2 MEGOHMS, VERTICAL-SIZE CONTROL
R_6 = 100000 OHMS, VERTICAL-PEAKING CONTROL
R_7 = 1 MEGOHM
R_8 = 100000 OHMS, VERTICAL-DISTRIBUTION CONTROL
R_9 = 10000 OHMS
R_{10} = 50000 OHMS, VERTICAL-CENTERING CONTROL
R_{11} = 50000 OHMS
L_1 = COUPLING CHOKE, 100 HENRIES
L_2 = VERTICAL-DEFLECTING COILS
T = FEEDBACK TRANSFORMER

In television circuits, it is essential that this saw-tooth wave be synchronized with that generated in the vertical scanning circuit of the Iconoscope at the studio. Synchronization is accomplished by a vertical-synchronizing signal which is transmitted with the picture signal and which consists of a series of voltage pulses. These pulses trigger the blocking-oscillator triode into oscillation. In practice, the circuit of the left-hand triode unit is adjusted to function at a rate slightly slower than the rate of the incoming synchronizing pulses. Hence, when the pulses are applied, the action of the blocking oscillator is speeded up to the proper rate for synchronization. Controls are provided so that the circuit can be adjusted for optimum operation.

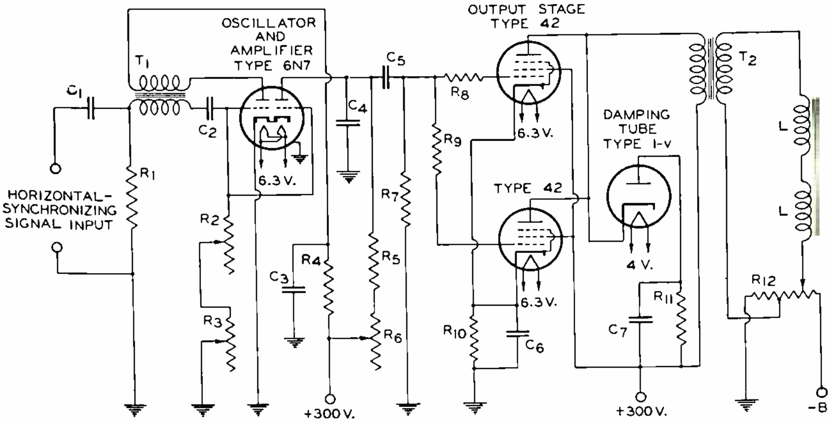

Fig. 4—Horizontal-deflecting circuit for RCA-1800.

C_1 = 0.005 μf
C_2, C_4 = 0.001 μf
C_3, C_5, C_7 = 0.05 μf
C_6 = 0.5 μf
R_1 = 500 OHMS
R_2 = 100000 OHMS } HORIZONTAL-SPEED
R_3 = 10000 OHMS } CONTROLS
R_4 = 100000 OHMS
R_5 = 50000 OHMS

R_6 = 0.25 MEGOHM, HORIZONTAL-SIZE CONTROL
R_7 = 0.5 MEGOHM
R_8, R_9 = 100 OHMS, NON-INDUCTIVE
R_{10} = 200 OHMS
R_{11} = 8000 OHMS, 5 WATTS
R_{12} = 5 OHMS, HORIZONTAL-CENTERING CONTROL
L = HORIZONTAL-DEFLECTING COILS
T_1 = FEEDBACK TRANSFORMER
T_2 = OUTPUT TRANSFORMER

A typical horizontal-deflecting circuit for the 1800 is shown in Figure 4. In this circuit, a 6N7 and an output stage generate a synchronized, saw-tooth current in a manner similar to that described for the vertical-deflecting circuit. However, the horizontal-deflecting circuit operates at a much higher frequency than the vertical-deflecting circuit. Because the horizontal-deflecting frequency is high, the deflecting current decreases very rapidly on the return portion of the deflecting cycle. This rapid decrease in current causes shock-excited oscillations in the plate circuit of the output stage. To damp out these oscillations, a Type 1-v tube is connected across the primary of the output transformer T_2. When oscillation starts, the primary first applies a high positive voltage to the cathode of the 1-v and then applies a negative voltage. As soon as the cathode becomes negative with respect to the plate of the 1-v, the 1-v conducts current. Thus, the flow of current through the 1-v quickly damps out the oscillations caused by shock excitation.

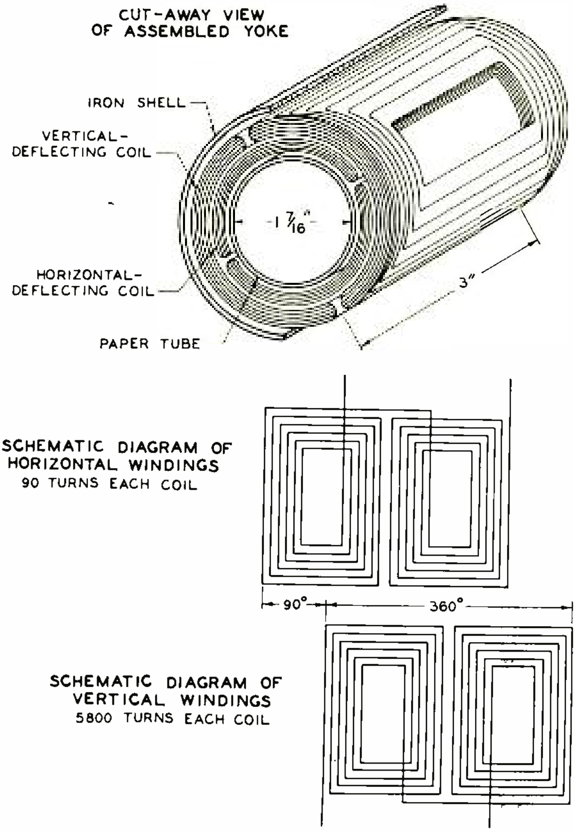

Fig. 5—Perspective view of a deflection yoke.

A deflection yoke is shown in perspective in Figure 5. This yoke is designed to minimize pin-cushion distortion and defocusing of the spot at the edges of the picture.

A NEW SYSTEM OF REMOTE CONTROL

BY

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ONE of the chief advantages of electrical machinery and electrically controlled devices is that they are readily arranged for control from points separated from the apparatus itself. The means of control usually involves merely the extension of wires and switches. However, there are cases where the use of extension wires is impossible or inconvenient. Furthermore, even where wires are permissible, it has usually been necessary, by the nature of the apparatus, to use arrangements which consumed some stand-by power even in inactive periods. One of these cases is the remote control of home broadcast receivers, where the use of control wires is highly objectionable and where continuous operation of control vacuum tubes or other control devices is undesirable.

With these facts in mind, an investigation was undertaken with a view toward providing an improved system of remote control. The development was given considerable impetus when it was found that a-c power lines could be used as a medium for the transmission of currents of low radiofrequency (less than 500 kc). This immediately suggested the possibility of using power lines as connecting links between the remote and controlled points.

I. NATURE OF THE PROBLEM

The general problem involving transmission of carrier currents on a-c supply lines has several aspects. First, the properties of the a-c line as a transmission medium must be determined, and the attenuation and line impedance at several frequencies must be known. Similarly, the power level required for the transmission of control signals is important, since the size and cost of apparatus are directly concerned.

The attenuation is probably the most important property of the a-c line when it is used as a carrier for r-f currents. If the attenuation is high, the power output of the control-signal source used as the remote-control unit must necessarily be quite large; this requires considerable equipment, in the form of power oscillators and amplifiers, at the control point.

The power required to actuate the control mechanisms at the controlled device is another factor which determines the required power capabilities of the remote-signal source, since, even with low line attenuation, the control power applied to the line at the remote point is a function of the energy consumed by the controlling mechanisms. This latter quantity will be determined generally by the type of circuit element employed to utilize the control energy. If power amplifiers were operating continuously to afford amplification of the control signals at the controlled point there would be no necessity for high-power levels at the remote unit. This would permit using the amplified control impulses directly to perform the desired functions, or they could be employed to cause relays to close certain power circuits. In this latter case the control energy from the line could initiate some secondary action, which might take the form of plate current passing through a relay in a power detector. The energy used to actuate the desired device could then be drawn directly from the detector power-supply.

In case stand-by power is considered objectionable, one might use a type of control device which could be actuated directly by the control signals. The copper-oxide rectifier is one circuit element which fulfils the requirements, but its operating power consumption is generally too large.

The choice of the proper frequencies to be used for control purposes, and the a-c line impedance at these frequencies, are both quite important. The frequencies employed must lie in a certain rather limited band, because of possible interference with radio receivers operating near the controlled device. The line impedance should not be too low, if proper impedance matching at the terminating points is to be accomplished readily.

II. TRANSMISSION AND POWER CONSIDERATIONS

Line impedance and attenuation were measured over the frequency range from 200 to 400 kc. This is the most useful range because its upper end is nearly the highest frequency which can be transmitted efficiently over the power line. Also the control frequencies should lie outside any bands used for broadcasting purposes, and also outside the i-f bands of radio receivers.

The line attenuation to these frequencies proved to be not very large. The maximum voltage attenuation on power lines in this laboratory was less than 10 to 1, as measured between the most remotely separated power outlets.

The average results of many line-impedance measurements made in typical buildings wherein a teledynamic control system of this type might be installed are as follows:

Average Line Impedance at 200 kc $= 25 + j40$ ohms
Average Line Impedance at 300 kc $= 40 + j60$ ohms

Loading the line with soldering irons, lamps and similar low-impedance appliances has little effect on the line attenuation, due to the low impedance of the line itself.

The maximum line attenuation to frequencies of 200 and 300 kc is about 20 db. This means that a signal source at the remote point with a power output of a watt or two would supply a resultant power of 0.01 to 0.02 watt at the controlled end of the a-c line, even under the most unfavorable attenuation conditions.

The possibility of using only a watt of r-f energy at the remote point is interesting, for, with this power rating, the remote unit may be made quite small physically. Since this unit will generally contain a radio-frequency oscillator as a source of control energy, the low power required makes the use of a small receiving-type tube quite feasible.

If one watt is taken as the normal r-f power supplied by the remote oscillator to the a-c line, a power attenuation of 100 means 0.01 watt available at the far end of the line. This may seem to be insufficient for control purposes, but it can be shown that it may be utilized to control, in many ways, non-power-consuming devices. Assume that the frequency of the control signals is 300 kc, and that the a-c line is properly terminated at the controlled point in an impedance of 72 ohms, (the resultant of $40 + j60$ ohms, the line impedance at 300 kc). Then the r-f voltage available at the line terminals is about 0.8 rms volt. The use of a resonant circuit, tuned to 300 kc, will result in an increase in amplitude of this voltage, due to resonant rise, to 30 or 40 volts rms, and at this level it may be applied to various circuits for control purposes.

The rise in voltage is not accompanied by any gain in power, consequently the increased voltage can not be used to operate a power-consuming device, such as a copper-oxide rectifier, because the tuned circuit across which the voltage appears is inherently a high-impedance arrangement. The r-f power drawn from the line should be used only to initiate some control action, whose energizing power is supplied from the 60-cycle power line.

A power amplifier might be so arranged that its output current could be controlled by the 30 or 40 volts of rf available due to resonant rise in the tuned circuit connected across the line. A relay, energized by the amplifier output current, could then be used to close any desired power circuit. This arrangement requires that the amplifier be in continuous operation, receptive at all times to control signals emanating from the remote point. The stand-by power consumed is somewhat objectionable, and requires that frequent replacement of amplifier tubes be made due to their continuous operation during quiescent periods of the controlled device. Accordingly, some means are needed for controlling the start of operation of the controlled device from the remote point without requiring the use of intervening amplifiers in continuous operation.

III. BASIC REMOTE-CONTROL CIRCUIT

Several tube and circuit arrangements were investigated to find a way to make this "cold-starting" feature possible. The requirements of no stand-by power eliminated immediately the use of a tube (such as a diode) with heater operating continuously, and pointed the way to the selection of a cold-cathode type of gas tube as the vital circuit element for use in this work.

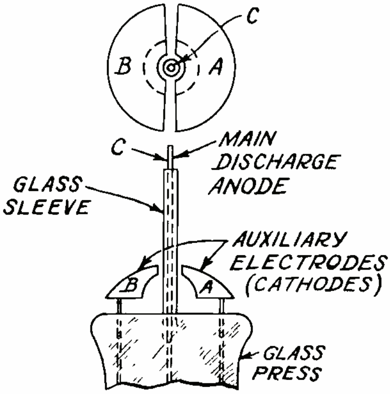

Fig. 1—Constructional details of gas tube.

The basic arrangement finally selected involves a gas tube and associated circuit elements which are so disposed that a small radio-frequency voltage transmitted over the power line to the pick-up circuits in the controlled device, and increased in amplitude by resonant rise in a tuned circuit, may be employed to add to the potential of one of the gas tube's elements and to initiate a gas discharge. This, in turn, causes a relay to close, and its contacts may then be used to connect the controlled device to the 60-cycle power line as a source of energy.

Figure 1 shows the constructional details of a gas tube which adequately fulfils the requirements. The tube consists of three elements, enclosed in an envelope containing gas at low pressure. The two similar elements, which are called auxiliary electrodes, are symmetrically disposed with respect to the third element. The two auxiliary electrodes act as cathodes, since the third element can act only as an anode to receive electrons from the other two. This is due to its construction and orientation with respect to the cathodes, which gives the tube a rectifying characteristic.

In a gas tube having the construction shown in Figure 1, the gap between *A* and *B* becomes ionized if the potential difference across it exceeds 70 volts. This is true regardless of the polarity of the voltage difference. Once started, the discharge is maintained as long as the potential difference exceeds 60 volts. The potential difference required to initiate a discharge between the anode *C*, and either of the cathodes *A* or *B*, in the absence of ionization in the gap *A-B*, is of the order of 200 volts. The discharge in gap *C-A* or *C-B*, once started, is maintained if *C* is 60 volts or more positive with respect to either of the two cathodes. If gap *C-A* or gap *C-B* is broken down, due to the presence of the required discharge voltage, a low-impedance current path exists in the main gap only if *C* is positive with respect to either of the cathodes. This action is due to the rectifying characteristic mentioned above.

The main gap will become ionized at 70 volts on *C* if first the gap between *A* and *B* is broken down. Thus, the potential on *C* may be made any value less than 200 volts and no discharge in the main gap *C-A* or *C-B* will occur until gap *A-B* is broken down. When this happens, a large current can be made to flow in the anode circuit if the anode is more than 70 volts positive with respect to the cathodes when the auxiliary gap *A-B* becomes ionized.

One way to utilize the controlling effect of the r-f voltage transmitted from the remote unit is shown in Figure 2. Here the gas tube is the same as that described above. The main gap *C-B* is supplied with 60-cycle voltage from the power line, but, since the peak voltage is only $\sqrt{2}$ times the line voltage, or about 180 volts, there is no discharge, because of the 200-volt threshold breakdown potential in the main gap.

If, however, a volt or two of radio-frequency energy from the remote unit is present across the power line in Figure 2, it may be increased in amplitude by resonant rise in the series circuit *LC*, which is tuned to the carrier frequency of the remote oscillator. The resultant r-f voltage which appears across *L*, and consequently between *A* and

B, may then be over 70 volts peak, and this will cause gap *A-B* to become ionized. The potential required to ionize the main gap is thereby decreased to 70 volts, and current flows through the relay winding during the portions of the a-c cycle when *C* is more than 60 volts positive with respect to *A*.

The amount of r-f voltage required to initiate the discharge in gap *A-B* may be reduced, and the sensitivity of the device increased, by supplying element *B* with a 60-cycle potential as shown in Figure 3. The main anode current then passes between *C* and *A* through *L*, instead of from *C* to *B*, because of the large 60-cycle reactance of the capacitance connected between *B* and ground. The capacitance voltage divider may be adjusted to apply 50-peak, 60-cycle volts between *B* and *A*, thereby reducing the r-f voltage required to break down gap *A-B* to about 20 volts. The sensitivity could be further increased by applying a still larger "polarizing" 60-cycle potential to *B*, but this would make the device susceptible to line-voltage variations.

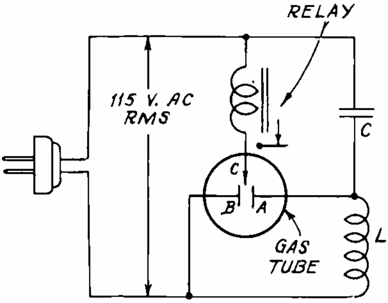

Fig. 2—Basic gas-tube circuit.

The use of a capacitance voltage divider, in place of a resistive potentiometer, for setting the polarizing potential of element *B*, permits the operation of the circuit with zero stand-by power.

The discharge in the main gap is not continuous, with operation on ac as shown in Figure 3. This would be true even if the auxiliary gap were continuously ionized. The reason for the intermittent nature of the main discharge lies in the sinusoidal variation of the voltage on element *C*, for, as soon as this voltage falls below sixty volts, the main discharge ceases, and is reinitiated only when the potential of *C* rises again during the next cycle to 70 volts. It is seen that the main discharge is not maintained even intermittently if the r-f voltage is removed from coil *L*, since the voltage across gap *A-B* is then insufficient to re-initiate the auxiliary discharge, which is also intermittent in nature, during the next cycle.

It is readily seen that the gas-tube discharge may be controlled from a remote point by transmitting a current of low radio-frequency over the power line. However, each gas tube at the controlled point requires a separate carrier frequency, and this means that if several functions are to be performed, the several gas-tube circuits must each be receptive to different carrier frequencies, if interaction is to be avoided and independent control achieved.

There is, however, an expedient at hand which may readily be used to decrease the number of carrier frequencies required. This method of control makes use of an r-f voltage which is modulated in intensity by the 60-cycle line voltage at the remote unit, and the added control feature is obtained by comparing the phase of the line voltage at the gas-tube circuit with the phase of the modulation envelope of the r-f voltage transmitted over the power line.

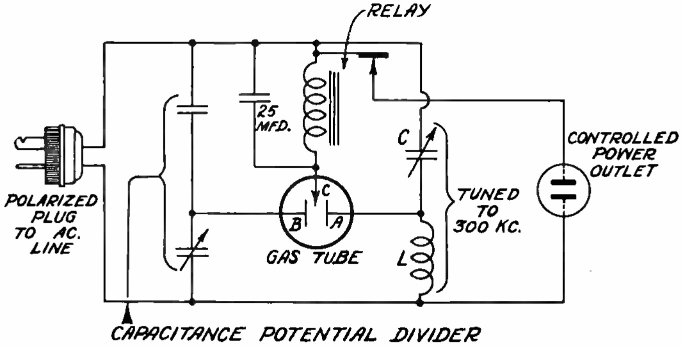

Fig. 3—Basic gas-tube circuit with means for improving sensitivity.

The source of energy for this type of remote control comprises a conventional feedback oscillator, whose plate-supply voltage is derived from the 60-cycle line, instead of from a d-c source. Application of 60-cycle ac as plate-supply voltage permits the tube to oscillate only in the portion of the a-c cycle during which the oscillator tube's plate is positive with respect to its cathode. Consequently, the oscillator output voltage which is fed into the line at the remote point is modulated in intensity, and consists of a series of pulses as shown in Figure 4. The interval during which oscillation takes place may be shifted by 180 degrees at 60 cycles if the phase of the line voltage is reversed before being applied to the oscillator circuit. This could be done by reversing the remote-unit power-supply plug. The r-f impulses then take the position shown dotted in Figure 4.

The r-f voltage, modulated in intensity, is picked up at the line terminals in Figure 3 and appears across coil *L*, as previously explained.

The main gap *C-A* can not be broken down in this case, even with sufficient r-f voltage available across coil *L*, unless the phase of the modulation envelope is properly related to the phase of the a-c line voltage on element *C*. The reason for this "selective ionization" is that the r-f voltage adds to the 60-cycle voltage across gap *A-B* only during a half-cycle at 60 cycles; hence, the gap *A-B* is ionized only during a portion of this interval, when the sum of the two voltages (r-f envelope plus 60-cycle polarizing voltage) is greater than 10 volts. If this auxiliary discharge occurs during the portion of the a-c cycle when element *C* is more than 70 volts positive, a low-impedance discharge takes place, and the relay in the anode circuit is energized. Reversal of the phase of the r-f envelope does not prevent ionization between *A* and *B*, for this still occurs, but only when *C* is negative with respect to *A*, and, thereby, the main-anode discharge current is not sufficient to actuate the relay.

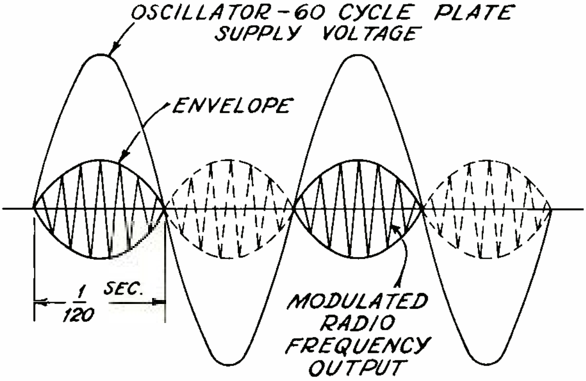

Fig. 4—Illustrating the modulated r-f output.

It can be seen that the gas tube must have a rectification characteristic if phase control of this type is to be possible. More current must flow in the main anode circuit when *C* is positive than flows when it is negative, otherwise the relay will be energized on either phase of the r-f envelope.

IV. USE OF SELECTIVE IONIZATION FOR CONTROL PURPOSES

Initiation of the gas discharge in a tube connected in a circuit similar to that of Figure 3 may be controlled by two variables; one is the carrier frequency of the remote oscillator, and the other is the phase of the r-f oscillator-output voltage envelope. Several similar gas-tube circuits may be set up at the controlled device and may be controlled independently from a remote oscillator, in which both the carrier frequency of the r-f voltage and the phase of the modulation envelope are variable.

Assume that two oscillators in a remote unit may be made to operate at either 200 or 300 kc as carrier frequency, and that the envelope phase may be controlled as desired; let the receptive circuits at the controlled unit contain four gas-tube arrangements, each similar in type to that shown in Figure 3; two of the gas tubes have their resonant circuits tuned to 300 kc, and the a-c potentials on the tube elements are so phased that the individual tubes are ignited by oppositely phased r-f envelopes; the second pair of gas tubes are arranged in the same manner for actuation by phase difference in the modulation envelopes, but their resonant circuits are tuned to 200 kc; let each gas tube have a relay in its anode circuit, and let the relays be numbered respectively 1, 2, 3 and 4. It is then possible to show, as is done in the following table, that ten different combinations are available by using two oscillator frequencies, 200 and 300 kc and the two opposite phases of the r-f envelope, *A* and *B*.

Oscillator No. 1		Oscillator No. 2		
Frequency	Envelope Phase	Frequency	Envelope Phase	Gas-Tube and Relay Number
300	A			1
		300	B	2
200	A			3
		200	B	4
300	A	300	B	1 and 2
200	A	200	B	3 and 4
300	A	200	B	1 and 4
200	A	300	B	2 and 3
300	A	200	A	1 and 3
300	B	200	B	2 and 4

The four relays may, therefore, be closed singly or in combinations of two at a time, depending upon the conditions existing at the remote unit. The relay contacts can be used to connect any electrical device to its source of power or to close other control circuits.

If three carrier frequencies are used with both phases of the modulation envelope, the number of possible combinations of frequency and phase is increased from ten to twenty-six. This would require six gas tubes at the controlled point, each with a multiple-contact relay in its anode circuit.

It is to be noted that the relays remain closed only as long as the r-f voltage is supplied to the line. This requires that the oscillator at the remote point deliver the required energy during the period that the relay is closed. For simple "on-off" switching the arrangement may be changed somewhat to eliminate the necessity for maintaining the r-f voltage during operating periods of the controlled device. This can be done by using two gas tubes, whose resonant circuits are tuned to the same frequency, but which are ignited by opposite phases of the modulating r-f envelope. One tube is used to close a small relay, which in turn closes a locking relay, whose contacts make the circuit for the controlled device. Removal of the rf from the line then has no effect on the locking relay. The other gas tube is arranged to be actuated by the opposite phase of the r-f envelope, and its relay is made to open the locking relay, thereby breaking the connection between the controlled device and the power line.

V. OTHER TYPES OF PHASE AND FREQUENCY DETECTORS

The "cold-starting" properties of gas tubes, with their lack of stand-by power requirements, makes them a very suitable device for "on-off" remote-control switching. They can be used to connect in circuit other types of phase and frequency detectors, which may contain vacuum tubes, and thereby require heater and plate-circuit power for operation.

One type of phase detector which is adaptable to this method of remote control consists of a high vacuum tube of the 6R7 type, which contains a triode and two small diodes. The circuit is as shown in Figure 5. Plate-supply potential is obtained directly from the a-c line, or through a step-up power transformer. The radio-frequency voltage present across the line is increased by resonance in the tuned circuit shown, and is applied to the control grid, in series with a suitable bias source.

The function of the bias is to maintain the tube at cutoff in the absence of control signals. This may be done by applying to the grid a 60-cycle voltage which opposes the a-c plate voltage in phase, and is slightly greater than the plate voltage divided by the amplification factor of the tube. No plate current flows under these conditions, but the grid current may be excessive during positive excursions of the a-c grid voltage, especially since the a-c source may have low impedance. To reduce this grid current, one may insert a high resistance in series with the grid; this action will have no effect on the detecting properties of the circuit.

Another way to maintain the tube at plate-current cutoff is to use a full-wave, rectified, unfiltered, 60-cycle voltage developed in the double-diode portion of the tube, as shown in Figure 5. The voltage is applied as grid bias, with the polarity arranged so that there is no positive grid-voltage swing with respect to cathode. The ratio of the amplitudes of the plate and grid voltages is somewhat less than the mu of the triode section. There is no grid current under these conditions, hence no series grid resistor is necessary.

The detector circuit operates in the following manner. Plate current flows through the relay winding when the phase of the r-f envelope is so arranged that the peak instantaneous grid voltage occurs coincidentally with the peak positive a-c plate voltage. If the r-f envelope is shifted 180 degrees in phase at the remote-control point, no current flows in the phase detector plate circuit because the plate is negative with respect to cathode (due to the 60-cycle plate voltage) when the pulse of r-f voltage is applied to the grid.

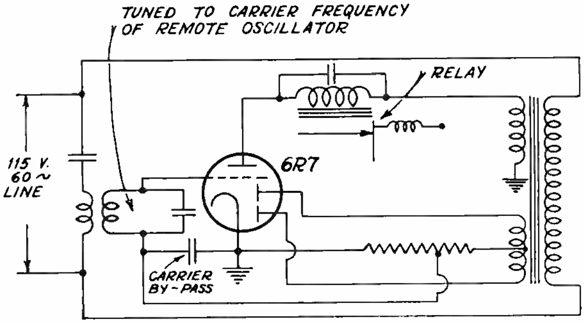

Fig. 5—High-vacuum type of phase detector.

The sensitivity of the device may be increased by rectifying the modulated r-f voltage present across the tuned circuit and extracting the low-frequency modulation components. The output of the diode rectifier used for this purpose is applied to the grid in series with the bias source. The increased sensitivity is due to the fact that the average value of the voltage applied to the grid is increased due to rectification in the diode.

The circuit of Figure 5 has the necessary qualifications for selection of the proper control signals, since it has both phase and frequency selectivity. Extension of this arrangement to include relay actuation on either phase of the modulation envelope may be made by using an additional triode, whose a-c plate voltage is opposite in phase to that on the tube described. More direct, however, is the use of a double-triode of the 6N7 type, as shown in the circuit of Figure 6. Here the a-c plate voltages are in phase opposition, and the relay which is actuated depends upon the phase relation between the r-f envelope and the a-c plate potential. Obviously both relays will be energized if both phases of the modulation envelope are fed into the line at the remote point.

A double-triode connected for phase detection as described above may be used in conjunction with only one carrier frequency; two double-triodes may be substituted for the four gas-tube circuits described in the first part of the paper, and, operation with two frequencies and both phases of the r-f envelope results in ten combinations of phase and frequency which will actuate four multiple-contact relays as described in Section IV. This, of course, would eliminate the "cold-starting" feature.

Small thyratrons or gas triodes may be substituted for high-vacuum tubes in this phase detector if it is desired to operate some power-consuming device directly from the anode current of the tube.

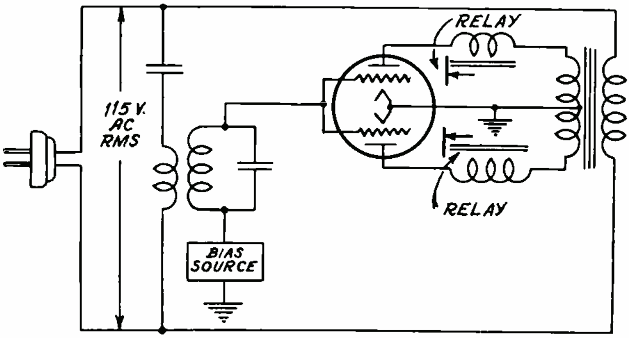

Fig. 6—Double-triode high-vacuum phase detector.

In this way a high-current solenoid or relay may be controlled by causing the gas discharge to be initiated by the properly phased r-f envelope.

VI. Applications of Remote-Control Circuits

The gas-tube circuit described above may be utilized for "on-off" control of any electrical device. Experimental models of the small remote-oscillator unit and the control circuit were constructed and were used for remote control (via the power lines) of lamps, bells and other electrical appliances.

In a remote-control arrangement of this type, the "on" switching is effected by causing the r-f impulses to initiate a discharge in the gas tube at the controlled point. The r-f energy is then supplied continuously to maintain the gas discharge during the operative period of the device. The controlled device is disconnected from its power supply by merely removing the r-f impulses from the line.

The necessity for maintaining the gas discharge while the controlled device is operating may be avoided by using a small locking relay in conjunction with two gas tubes and their associated circuit elements; one circuit is used for "on" switching, and the other is used for breaking the locking-relay contacts to turn the controlled device off.

VII. RADIO-RECEIVER CONTROL

This method of remote control may be readily applied to radio receivers which are already equipped for so-called "Electric Tuning". In receivers of this type the "Electric Tuning" is accomplished by means of a small motor and cam arrangement used in conjunction with push-button switches which are mounted on the front panel of the receiver. These switches are arranged to close the motor circuits to vary the tuning or volume level.

Automatic frequency control is generally used in "Electric Tuning" receivers, and this is a necessary feature, for even the best motor-and-cam arrangements are not sufficiently precise in their action to permit tuning with the required degree of accuracy.

For remote control, a unit, located in the receiver, contains phase and frequency detectors of the type previously described. These are arranged to be actuated by a small oscillator at the remote point, and r-f energy transmitted over the power line causes energization of relays connected in the anode circuits of the tubes in the receiver control unit. The contacts of the relays are connected in parallel with the push-button switches in the "Electric-Tuning" receiver, so that the closure of the relays, either singly or in combinations of two at a time, causes power to be supplied to the tuning-motor circuit as in regular "Electric Tuning".

One of the circuits in the receiver unit should contain a gas tube, if remote control of the "on-off" switching is desired. The gas-tube relay may be arranged to close a locking relay for "on" switching, and, once this function is performed, the gas tube may then be used for additional control functions.

Three other gas tubes may be used to actuate the three remaining relays, or high-vacuum phase detectors of the type previously described may be employed to control the closure of the required relays.

The remote unit consists of two triode oscillators (a double-triode 6N7 type of tube may be used) in which the carrier frequency and envelope phase are controlled by push-button switches mounted on the panel of the remote unit. The output of this device is fed into the power line at the point from which remote control is desired.

Use of two frequencies and two envelope phases in the remote unit permits control from the remote point of the "on-off" feature, and of the selection of any six broadcast stations. In addition, remote control of the volume level is available. The stand-by power at the receiver when it is not operating is zero; similarly, the power consumed in the remote unit when no tuning or volume operation is being controlled is merely that required in the heater circuit of the oscillator tubes.

Several remote-control arrangements of the type described in this paper may be operated independently and without interaction on adjacent power circuits if means are taken for mutual isolation of the power lines for the r-f control currents. This may be done easily by inserting a small r-f choke at the meter board of each power line. If this is impracticable, different sets of carrier frequencies may be used for each remote-control installation, say 250 and 350 kc for one installation and 200 and 300 kc for another.

These precautionary measures will, however, not be generally necessary, because of the relatively high natural attenuation existing between adjacent power circuits.

DESIGN TRENDS IN MOBILE RECEIVERS IN AMERICA[1]

BY

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IN GENERAL, radio receivers designed for mobile use are limited to the fields of automobile entertainment, police radio receivers, aircraft radio receivers, and marine radio receivers. This paper initially delivered, covered automobile entertainment receivers in a manner substantially identical to that published by Smith[2]; in consequence this material is omitted here.

Because of the fact that marine radio receivers may be extremely bulky—often approaching in dimensions and weight the characteristics of fixed point-to-point apparatus—they bear very little similarity to radio receivers designed for other mobile applications. This paper will, therefore, be limited to a discussion of aircraft and police radio receivers.

POLICE MOBILE RECEIVERS

Police radio communication has been assigned to the bands of 1610 to 1712 kilocycles, 2310 to 2490 kilocycles, and 30 to 40 megacycles. The medium-high-frequency bands are assigned on a permanent basis whereas the ultra-high-frequency assignments are experimental only. The 1610-to-1712 band is used for one-way state police systems while municipal police services are given assignments in the 2310-to-2490 band. Two-way communication systems are limited to the ultra-high-frequency band except in instances in which ultra-high frequencies are used as part of medium high-frequency systems for talk-back purposes from the car to fixed station.

One of the first mobile police receiver installations in the United States was made in the city of Detroit in 1928. These receivers were constructed much as the first entertainment automobile receivers and required separate "A" and "B" batteries. The remarkable success of the mobile-receiver installations caused the system to spread rapidly. As a result, several cities added equipment in 1929 and 1930 and in

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[1] This paper was delivered by the authors on July 30, 1937 before the Deutsche Gesellschaft für Technische Physik in Berlin.

[2] "Automobile Receiver Design", Jerome C. Smith, RCA REVIEW, April 1937.

1931, the State of Michigan installed a 5-kilowatt transmitter at Lansing. In 1932, the first experimental assignments were made in the ultra-high-frequency band. Since ultra-high frequencies offered the possibility of talk-back from the car on the same frequency or another frequency in the same band, as well as the possibility of car-to-car communication in large municipalities, this band has been extremely popular.

Shortly after assignments began to be made in the ultra-high-frequency band, a large number of municipalities began experimental use of two-way police communication. One of the first licenses issued for two-way police communication was to the city of Bayonne, N. J., in 1932. On March 1, 1937, there were 226 cities operating 1367 mobile police transmitters and receivers providing two-way communication.

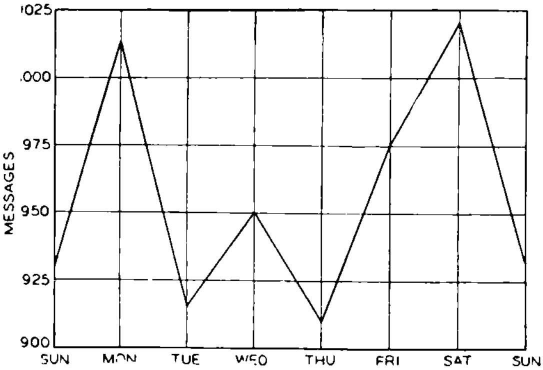

Fig. 1—Average daily load chart of a representative municipal police radio station.

The popularity of two-way police radio communication has increased materially during the past few years because of the speed with which a chase may be organized or an order executed. The first talk-back systems from the car to the central station were used merely to acknowledge orders or to assure the central station that an order had been received. However, it soon became apparent that violations could be quickly reported from patrol cars, and that in many cases, a chase could be organized from one of the pursuing control cars more efficiently than it could be handled through the central station alone. This fact has added materially to the popularity of ultra-high frequencies.

Figure 1 illustrates a daily load chart of a representative municipal police radio station. In this case the number of messages per day are plotted as a function of the day of the week. Figure 2 illustrates the average hourly load chart which is representative of an average police radio station. These figures indicate traffic demands on installations of this sort. They are reproduced through the courtesy of Messrs. E. L. White and E. C. Denstaedt, in whose article entitled "Police Radio Communication" these figures appeared. Credit is also given to the American Institute of Electrical Engineers' official journal, in the May 1937 issue of which the article was published.

The first conventional band mobile-police receivers were of the tuned-radio-frequency type and were continuously variable over a band of frequencies. A great many of these early receivers were modified entertainment automobile receivers. As mobile police installations became more popular, a demand arose for single-signal receivers which were fixed-tuned to the local transmitter, and were tuned only as a matter of routine maintenance rather than by the operator. It soon became apparent that the type of construction normally employed in entertainment automobile receivers was not entirely satisfactory for the services normally encountered in police usage. As a result, more rugged construction and greater stability brought about the design of receivers especially intended for police service.

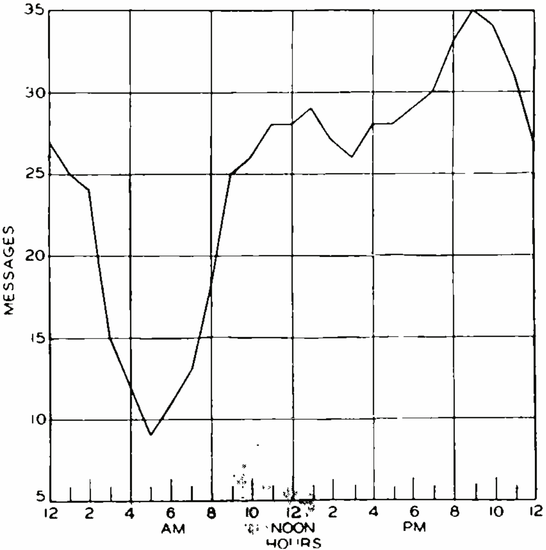

Fig. 2—Average hourly load chart of a representative municipal police radio station.

Because of vibration difficulties, it was some few years before motorcycles were equipped with radio receivers. Adaptation of radio receivers to motorcycles made even further demands on the construction of these equipments on account of the severe vibration encountered under operating conditions.

The receiver is mounted on the luggage carrier of the motorcycle, directly behind the rider. The antenna is a short rod projecting from the back of the receiver. These receivers were designed for use in the medium-high-frequency bands only. Figure 3 shows a close-up view of the receiver.

Vibration encountered on a motorcycle results from two basic sources; first, road shock; and second, engine vibration. Recalling that the motorcycle frame is spring mounted only on the front wheel, it will be seen that a piece of apparatus secured to the frame obtains very little protection from road shock by means of the motorcycle structure. To protect the apparatus from road shock, it is general practice to mount the entire receiver case on a shock-absorbing medium, such as rubber. The vibration caused by the engine shock excites the relatively light parts of the motorcycle, and, as a result, sustained high-frequency oscillations appear in the frame. The frequency of these oscillations, however, is a function of the engine's speed. Therefore any portion of the equipment may be subjected to a more or less sustained vibration, and the source of the vibration may be near the natural period of this portion of the equipment. Small lengths of wiring, tube elements, entire chassis (in fact any number of portions of the equipment) may build up an amplitude of vibration sufficient to

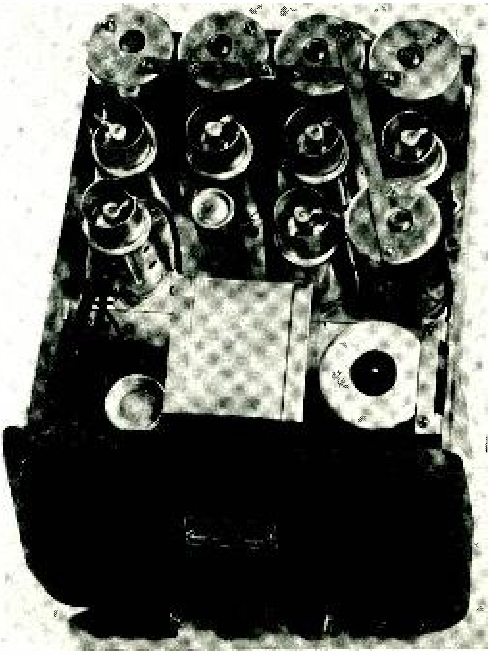

Fig. 3—Top view of medium-high-frequency motorcycle receiver.

cause destruction unless the parts are properly designed and the mountings so arranged that the mass of the set and the damped elasticity of the shock mounting function as a low-pass filter.

In the design of the receivers, it has been found desirable to use a very rugged chassis so that the chassis itself will have a vibration frequency above that at which it will be excited in use. The tubes are secured in their sockets by means other than the conventional tube sockets. Usually these take the form of clamps secured to the chassis and exerting a clamping action on the base of the tube. Wiring is carefully designed to avoid rubbing on screws or sharp edges. The wires are laced in position because the high-frequency vibration to which they are subjected would otherwise cause their gradual displacement from the original position and result in failure due to chafing of the insulation.

Ignition interference on the motorcycle was immediately recognized as a severe problem because the high-tension wiring is not protected to the extent that it is in automobiles. The operator of the motorcycle sits in such a position that his legs are closely coupled to the high-tension ignition system, and as a result his body radiates an intense field which may be picked up by the antenna. The treatment of the ignition system that has been found necessary consists of enclosing the entire high-tension system in a shield. A shield clip is placed over the spark plug and a suppressor is used within this clip. In another form of shielding, it has been found practicable to place the suppressor under the braid used to shield the high-tension wire and adjacent to the plug. The charging generator is by-passed. Both battery-distributor and magneto-type ignition systems are employed on motorcycles. In the case of battery-distributor installations, it is necessary to by-pass the ignition switch to prevent radiation on the low-tension system. It is also necessary to shield the distributor and coil, or magneto, as the case may be.

A considerable amount of noise is experienced on motorcycles, particularly on dry pavements, as a result of the generation of tire static. Bonding of the wheels to the frame by means of a rubbing contact by a stiff spring reduces this static to an acceptable degree.

It has been essential in the design of motorcycle receivers to keep the primary power consumption at a minimum since the motorcycle battery is of extremely low capacity. It is therefore necessary that an efficient loud speaker be used so that the power output of the receiver will not be inconsistent with the available size of battery. It is general practice to use highly efficient electro-magnetic loud speakers. Since the entire equipment is exposed to the elements, the receiver is necessarily made water-tight, and the loud speaker is of such a design as to prevent direct spray from reaching the elements, and to allow any accumulation of moisture to drain away. Also suitable materials and finishes are applied to resist corrosion.

After a considerable period of usage of the type of installation just discussed, experience has indicated that a saddle-bag type of installation would be more suitable. The earlier design placed the weight of the receiver apparatus considerably above the center of gravity of the motorcycle and tended to cause instability at high speeds. The motorcycle operators experienced difficulty in mounting the motorcycle when using the earlier type of installation, and were concerned over the danger of having a solid object directly to the rear in case of collision or other accidents. For these reasons, later designs included the change to the saddle-bag construction.

The power supply unit, consisting of a highly efficient self-rectifying vibrator, transformer and filter, is mounted in one of the two cases. The receiver proper is mounted in the other case. The antenna is a screen mesh supported on rubber insulators to the rear of the two cases. The selection of this location for the antenna is intended to place it as far as possible from the ignition system and to reduce coupling to the frame of the motorcycle to avoid pickup of parasitic noise. It is mounted on rubber insulators to avoid breakage of the insulation due to vibration. The antenna tuning system consists of tapped inductances shunted by fixed capacitors. The antenna and detector circuits are tuned by means of finely divided magnetite cores over any one of the two medium-high-frequency bands. A screw is removed from each coil assembly to change from the high-frequency band to the lower-frequency band. Because the antenna is of extremely high impedance (of the order of 15 $\mu\mu f$ capacity), a maximum antenna gain is obtained by direct coupling of the antenna to the high side of the input tuned circuit, and as a result, an excellent signal-to-thermal noise ratio is obtained. To conserve space, a multi-grid converter tube is used for the first detector and oscillator. The oscillator circuit uses a tuned anode with the proper choice of voltage to obtain very high stability with respect to power supply variations. The inductance is so designed as to have a low temperature coefficient. A special design of fixed air-dielectric capacitor is used in parallel with the tuning capacitor. A circuit having a low ratio of inductance to capacity is used. This is inherently stable and is temperature compensated to take care of changing circuit capacity and tube-element capacity with respect to temperature. The capacitor is constructed of two concentric tubes of dissimilar metals spaced by very stable low-loss insulation at each end to obtain the desired negative temperature coefficient. One stage of 260-kilocycle intermediate-frequency amplification is used. This intermediate frequency was chosen as low as practical so as to have a maximum of gain with a minimum current drain, and to obtain the desired degree of adjacent channel selectivity, but was chosen high enough to obtain approximately 60 db of image suppression.

Because of the very high impedance used, an antenna filter similar in principle to that used in modern entertainment automobile receivers, caused excessive attenuation. Very careful filtering has been applied to the "A" circuit.

The sensitivity of the receiver is limited by the first-circuit thermal noise. For this reason, it is customary to express the sensitivity as a signal-to-noise ratio. The total circuit drain from the 6-volt battery is 2.5 amperes. By means of the temperature-compensating device, the tuning of the receiver remains within approximately 1.6 kc over a temperature range of 30° C to 66° C.

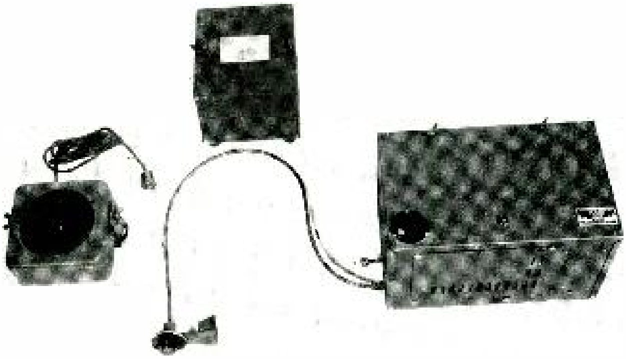

Fig. 4—Medium-high-frequency automobile receiver.

Police automobile-receiver design for the medium-high-frequency bands has followed closely that of the entertainment automobile receiver design except that, as previously mentioned, ruggedness and stability are improved. A normal police installation uses the car receivers so many hours a day that practically a continuous rating must be applied to the receiver components and the power supply. It is also necessary that the receiver be made readily removable from the car and easily serviced.

The majority of police car receivers employ a dynamotor operated from the car battery to supply the necessary "B" potential. This dynamotor is designed for very long periods of service with a minimum of attention.

A typical medium-high-frequency receiver is shown in Figure 4. The receiver mounts to the fire wall of the car by means of a single bolt. A highly efficient electromagnetic loudspeaker unit is mounted in the header of the car. The circuit, shown in Figure 5, consists of a radio-frequency stage, first detector, separate heterodyne oscillator, one stage of 175-kc intermediate-frequency amplification, a diode-triode tube in which the functions of second detector, automatic volume control and squelch are embodied, an audio-amplifier stage and output stage.

The oscillator circuit is temperature compensated in a manner similar to the motorcycle receiver. The triode section of the second detector is arranged so that it draws current until one of the diodes receives a predetermined rectified current from the desired carrier. When this triode is biased off, it permits normal grid bias to occur in the first audio-amplifier stage and thus renders the set operative. A control located on the front of the receiver permits the operator to select any desired level for the operation of the squelch between 12 to 100,000 microvolts input. These receivers are normally operated with the squelch circuit in use to avoid fatigue of the operator. The sensitivity of the receiver is limited only by first-circuit noise and the sensitivity, expressed in terms of 6 db signal-to-noise ratio with carrier on, is less than 3 microvolts. The maximum output is approximately 2 watts and the primary power supply is approximately 6.5 amperes at 6 volts.

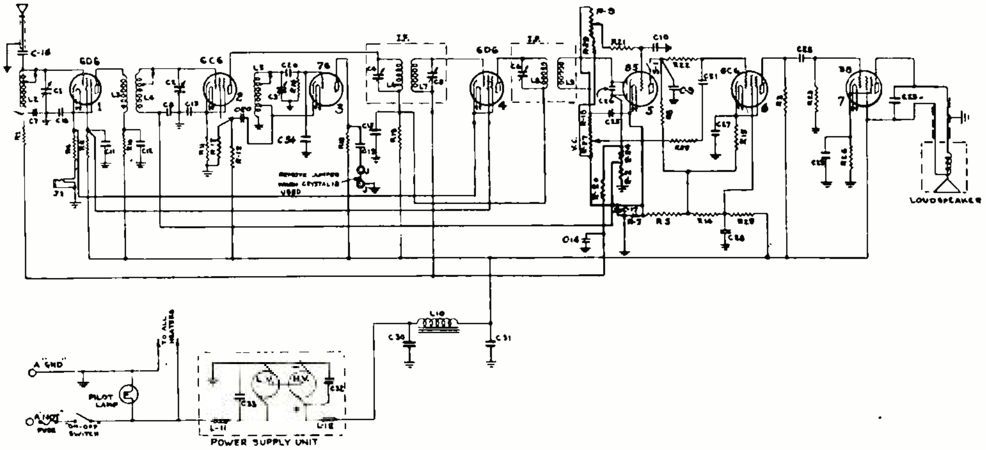

Fig. 5—Schematic circuit diagram of receiver shown in Figure 4.

When the ultra-high-frequency band was first assigned for police usage, superheterodyne receivers were not available for use at these frequencies. Inasmuch as considerable pioneering effort was indicated, super-regenerative receivers were first used. These receivers were not only inexpensive but, in addition, had sufficiently low selectivity so that temperature compensation and other refinements were not required. Since there were few police radio installations in this band in any given vicinity high selectivity was not required. Moreover, most of the ultra-high-frequency police transmitter installations at that time did not employ crystal control. Consequently, a high degree of selectivity would have added considerably to the maintenance of the equipment. The limited selectivity permitted continuous reception for reasonable distances from the transmitter, irrespective of the frequency drift. During this period, practically all ultra-high-frequency police transmitters operated with carrier on continuously. Under these circumstances, the limiting action obtainable in super-regenerative circuits was extremely valuable. After the art developed, however, crystal-controlled transmitters came into use in which the carrier was radiated only during the period of modulation. This immediately obsoleted the super-regenerative receivers and receivers embodying more refinements of design were introduced.

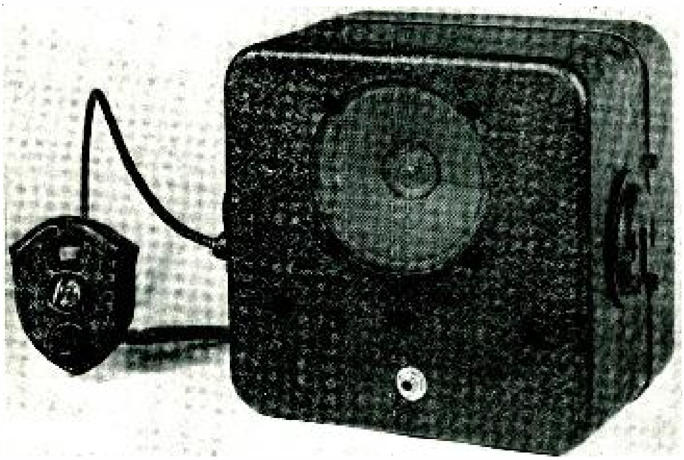

Fig. 6—Super-regenerative ultra-high-frequency police receiver.

Figure 6 shows a super-regenerative receiver of this period. A dynamotor power supply was employed with an on-and-off switch mounted on the steering column. The output of this receiver was approximately 1 watt.

(To be continued)

RECENT MARITIME RADIO LEGISLATION IN THE UNITED STATES

By

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and

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On May 20, 1937, the President of the United States signed an Act of the 75th Congress known as Public Law No. 97. This Act not only amended several sections of the Communications Act of 1934 but also wrote into that act many new sections. The primary purpose of Public No. 97 is to further safety of life and property at sea. Consideration of its provisions will be assisted by a brief review of earlier radio legislation and international agreements affecting commercial shipping.

For many years the Ship Act of 1910[1], as amended July 23, 1912[2], was applicable to ocean-going as well as Great Lakes steamers of American registry plying between ports 200 miles or more apart and carrying, or licensed to carry, 50 or more persons, including passengers or crew or both. It was required that the radio installation be capable of transmitting and receiving messages over a distance of at least one hundred miles, and that an auxiliary source of power be provided to permit the operation of the transmitter for a period of at least four hours in the event of the failure of the vessel's main power supply. It was provided that such radio equipment be in charge of two or more qualified radio operators, with one or the other on duty at all times when the vessel was being navigated. "Public No. 97" repealed these general provisions with respect to ocean-going vessels but it left them still expressly applicable to vessels operating on the Great Lakes.

The United States is a party to the International Telecommunication Convention of Madrid, 1932, and the General Radio Regulations annexed thereto. That Convention and its General Radio Regulations made far-reaching provisions relating to the operation and administration of radiotelegraph and radiotelephone services, and the allocation of radio frequencies. However, its effect upon ship radio equipment may be said to be slight, and further discussion is therefore unnecessary for the purposes of this paper.

The Federal Communications Commission, by virtue of authority

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[1] Public Law No. 262, 61st Congress, approved June 24, 1910; 36 Stat. 629.

[2] Public Law No. 238, 62d Congress; 37 Stat. 199.

granted to it by the Communications Act of 1934, has promulgated various specifications and regulations applicable to ship-board radio installations. Most of these regulations are found in its General Orders and its Rules and Regulations.

The International Convention for the Safety of Life at Sea, held in London in 1929, adopted a number of new and important regulations affecting radio in commercial marine service. The United States Senate ratified the Convention on June 19, 1936, and this ratification was deposited with the British Government on August 7, 1936. On November 7, 1936, three months later, the provisions of the Convention became effective for vessels of United States registry.

It is proposed to treat first the provisions of the Safety of Life at Sea Convention rather than Public No. 97, since the former constitute the fundamental regulations, the requirements of which Public No. 97 parallels and, in some instances, exceeds.

The radio provisions of the Convention apply to all ships of *international* voyages, except vessels of less than 1600 tons gross. A cargo ship is "any ship not being a passenger ship," and a passenger ship is defined as any ship which carries more than 12 passengers. Any person carried on a ship, other than its officers and the crew actually required to man and operate it, persons employed to carry on the business of the ship, and persons who have been rescued from shipwreck, distress or similar situations, is deemed a passenger. The Safety Convention made the following requirements as to equipment and the standing of watches:

EQUIPMENT

The chief radio installations required by the Safety Convention are:

1. A main radiotelegraph transmitter with a normal range of at least 100 nautical miles to operate on 500 kc, 375 kc and at least one working frequency in the band 350-485 kc with A-2 (modulated) emission, or B (spark) emission if transmitter installed prior to January 1, 1930.
2. An emergency transmitter similar to the main transmitter except that a normal range of at least 80 nautical miles is required.
3. A main radio receiver, 350 to 515 kc and 100 to 200 kc for reception of A-1 (continuous) waves, A-2 (modulated) waves and B (spark or damped) waves.
4. An emergency receiver, 350 to 515 kc, with provision for using a crystal rectifier (detector).
5. For emergency operation, a source of power supply independent of the ship's main power supply, with a capacity for six hours of operation.

The emergency transmitter and receiver are not required if the main transmitter and receiver also comply with all the requirements for the emergency equipment.

An independent communication system between the radio room and the bridge, emergency lights, seconds-hand clock, spare parts, and tools constitute other equipment required by the Safety Convention.

Passenger vessels of 5000 tons gross and upwards are required to be fitted with an approved radio direction finder.

Where the number of lifeboats carried on a ship exceeds thirteen, one is required to be a motor boat, fitted with an approved radiotelegraph installation. If the number of lifeboats exceeds nineteen, two of them must be motor boats, each fitted with an approved radiotelegraph installation.

WATCHES

The Safety Convention introduced the requirement of a continuous radio watch on cargo ships of over 5500 tons gross and on passenger ships of over 3000 tons gross. The use of an approved auto alarm on the distress frequency of 500 kc at all times when the qualified radio operator or a certified watcher is not on duty is permitted. Cargo ships under 5500 tons and passenger ships under 3000 tons are required to maintain hours of watch as prescribed by their respective administrations, except that cargo ships from 3000 to 5500 tons gross not fitted with auto-alarms are required to maintain at least eight hours' watch per day. In general, watch requirements of ships compulsorily equipped in accordance with the Ship Act were not affected because the Ship Act required a continuous watch by one of two or more operators, and the use of an auto-alarm for any part of the continuous watch was not permitted.

The Safety Convention provides for possible exemption from its requirements in cases where the route or conditions of the voyage would make a radio installation unreasonable or unnecessary. Public No. 97 covers similar situations which are discussed further on in this paper.

Ratification of the Safety Convention by the United States affected about 640 U. S. vessels engaged in international voyages. Nearly all of the ships were already equipped with main, or main and emergency radio installations, and a reasonable period of time was allowed for installing any additional equipment required.

About 600 American ships that were not affected by the Safety Convention or the Ship Act are now covered by Public No. 97. The enactment of Public No. 97 brought into existence the first uniform minimum radio requirements for all ocean-going U. S. ships, both passenger and cargo, on international or domestic voyages, which come under the provisions of the Act. While Public No. 97 is patterned largely after the radio provisions of the Safety Convention, the new Act provides for higher standards than the Safety Convention in many respects. Before considering the specific requirements of Public No. 97, brief mention will be made of the existing radio regulations that are applicable to ship radio operation. At the present time, the following laws, treaties and regulations apply:

The Ship Act of 1910, as amended in 1912, and the Federal Communications Commission's Rules and Regulations to carry out the provisions of such Act (now applicable solely to vessels navigating the Great Lakes).
The International Telecommunication Convention, Madrid, 1932, and its annexed General Radio Regulations.
The Communications Act of 1934, as amended, and as further amended by Public No. 97.
The general Rules and Regulations of the F. C. C. (excluding Rules 281 and 296 and subparagraph "d" of Rule 284).
The General Orders of the F. C. C.
The International Convention for the Safety of Life at Sea, London, 1929.
The Ship Radiotelegraph Safety Rules of May 21, 1937, issued by the F. C. C.

The Ship Radiotelegraph Safety Rules of May 21, 1937, were issued by the F. C. C. to give effect to Public No. 97. These rules specify the classes of vessels covered by the Act, those specifically excepted by the Act, and the conditions governing requests for exemptions for ships on certain routes. Requirements are also given for radio equipment, operators and watches, inspection procedure, frequency allocations, ship station licenses, etc. The May 21, 1937, Safety Rules, while primarily drafted to encompass the requirements of the new Act, also provide for compliance with the Safety Convention, the Telecommunication Convention of Madrid, 1932, and the General Radio Regulations annexed thereto, and the Commission's general Rules and Regulations.

With the exception of ships specifically excepted, the Communications Act and the Commission's Safety Rules apply to any U. S. vessel navigated in the open sea outside of a harbor or port, and to any U. S. or foreign vessel which leaves or attempts to leave a U. S. harbor or port for a voyage in the open sea, other than a cargo ship of less than 1600 gross tons. Vessels not required to comply are: a ship of war; a ship owned and operated by the United States Government (except vessels of the United States Maritime Commission, the Inland and Coastwise Waterways Service and the Panama Railroad Company); a foreign "Safety Convention ship" with a valid certificate or exemption; yachts under 600 tons not subject to the radio provisions of the Safety Convention; vessels in tow; and finally, a vessel navigating solely on the Great Lakes or on any bays, sounds, rivers or protected waters within the jurisdiction of the United States, or a vessel leaving or attempting to leave a U. S. port or harbor for a voyage solely on these waters.

The definition of a vessel as a passenger or cargo ship is the same as that of the Safety Convention; namely, any ship is classed as a passenger ship if it carries or is licensed to carry more than 12 passengers, while a ship not in this class is considered as a cargo vessel. Special consideration is given to yachts under 600 gross tons which do not go on international voyages and hence are not subject to the Safety Convention. Such yachts may carry or be licensed to carry more than 12 passengers and need not comply with the new Act. A yacht of 1600 gross tons or over is classed as a cargo ship if it carries not more than 12 passengers and is certificated accordingly. Any yacht (or other vessel) up to 1600 tons, not classed as a passenger ship, need not comply regardless of the nature of the voyage.

For convenience, the principal sections of Public No. 97, now a part of the Communications Act of 1934, that relate to radio equipment and radio operators on board ship are reproduced at the end of this paper. The conditions under which a request for exemption will be considered are set forth in Section 352 of that reproduction. A U. S. vessel, which is shown to comply with the Safety Rules after inspection by the Federal Communications Commission, will obtain a ship station license certified to that effect. In addition, U. S. ships which are subject to the Safety Convention are required to carry a Safety Certificate (for passenger vessels) or Safety Radiotelegraphy Certificate (for cargo vessels) or an Exemption Certificate (for exempted vessels). These certificates are issued by the Bureau of Marine Inspection and Navigation of the Department of Commerce, after the F. C. C. has certified to the Bureau that the radio installation complies with the requirements or that the vessel has been exempted. After the initial inspection, each ship radio installation must be inspected at least once each year in order to obtain a renewal of the certificate and of the license certification. Routine additional inspections are also made at frequent intervals.

A passenger vessel under Public No. 97 and the subsequent Safety Rules of the F. C. C. is required to be fitted with a radiotelegraph installation comprising separate main and emergency transmitters and separate main and emergency receivers.

The main transmitter must have a normal range of at least 200 nautical miles with a minimum antenna power of 200 watts on 500 kc and be capable of A-2 (modulated) emission or B (spark) emission on 500 kc, 375 kc, and at least one working frequency between 350 and 485 kc. This does not prohibit also the use of A-1 (continuous wave) emission and additional working frequencies in the 350-485 kc band, as well as additional frequencies or other transmitters in the assigned low- or high-frequency bands. The emergency transmitter is required to have a range of at least 100 nautical miles, a minimum antenna power of 50 watts, and other characteristics the same as the main transmitter. In addition, the emergency transmitter and emergency receiver must have a source of power supply, independent of the shipboard main power supply or any other electrical system, which is capable of operating the emergency equipment for at least six consecutive hours.

The main receiver must be capable of reception of A-1, A-2, and B emission in the band 350 to 515 kc and the band 100 to 200 kc. The emergency receiver is required to cover the band 350 to 515 kc with A-1, A-2, and B reception, and in addition must be provided with a crystal detector or a crystal receiver which will permit reception in case of failure of all vacuum tubes or power supply for the emergency receiver.

An independent means of communication between the radio room and the bridge is required. A suitable clock provided with a sweep-seconds hand and an emergency light or lights are required in the radio room. Various spare parts, tools, spare antenna material, a three-range d-c voltmeter, fuses, hydrometer, and instruction books are also necessary. The use of an automatic alarm, in lieu of a qualified operator, is not permitted on U. S. passenger ships, since a continuous operator watch is compulsory. Motor lifeboats on passenger vessels are required to be fitted with a complete radiotelegraph installation as outlined previously under Safety Convention requirements. The lifeboat radio equipment must meet the F. C. C. specifications, which call for A-2 emission on 500 kc with a minimum of 75 watts plate input to the stage supplying power to the antenna. The antenna must be an inverted "L" or "T" and not less than twenty feet above the water line, with the maximum practicable length. The lifeboat receiver must cover the band 350 to 550 kc with reception of A-1, A-2, and B waves. A suitable storage battery in the lifeboat to provide continuous operation for at least six hours is required. An approved type of radio direction finder is also specified for each passenger ship of 5000 gross tons and over.

The rules for cargo vessels are in general similar to those outlined above for passenger vessels with the following exceptions: A direction finder is not required. The use of an approved automatic alarm is permitted on cargo vessels, provided at least an 8-hour watch per day in the aggregate is maintained by a qualified radio operator. The automatic alarm is required to be in operation whenever the operator is not on watch while the ship is being navigated outside a harbor or port, including periods when the direction finder is in use. A separate emergency transmitter and receiver are not required on cargo vessels, provided the main installation complies with all the provisions affecting the emergency installation. Where a main transmitter on a cargo vessel is thus affected, this is sometimes accomplished by providing storage-battery, emergency-power supply to operate the transmitter at 50 watts output. The main receiver normally uses storage batteries for filament supply and dry batteries for plate supply, and if also arranged for crystal-detector reception, will comply with the requirements for the emergency receiver, provided six hours reserve battery capacity is available.

The Safety Rules also provide for the use of type "B" emission (spark sets) with 1000 watts input to the transformer for the main transmitter and 500 watts input to the transformer for the emergency transmitter. Spark set operation however is limited to transmitters that were installed prior to January 1, 1930. No new installations of spark sets are permitted on U. S. vessels. The high technical standards of the Safety Rules can be appreciated when it is realized that most of the foreign administrations do permit the installation of spark sets.

The principal sections of the Communications Act of 1934, relating to radio equipment and radio operators aboard ship, that became a part of Title III of that Act by the passage of Public No. 97, are as follows:

PART II—RADIO EQUIPMENT AND RADIO OPERATORS ON BOARD SHIP

SHIP RADIO INSTALLATIONS AND OPERATIONS

Sec. 351. (a) Except as provided in section 352 hereof, it shall be unlawful—

(1) For any ship of the United States other than a cargo ship of less than sixteen hundred gross tons, to be navigated in the open sea outside of a harbor or port, or for any ship of the United States or any foreign country, other than a cargo ship of less than sixteen hundred gross tons, to leave or attempt to leave any harbor or port of the United States for a voyage in the open sea, unless such ship is equipped with an efficient

radio installation in operating condition, in charge of and operated by a qualified operator or operators, adequately installed and protected so as to insure proper operation, and so as not to endanger the ship and radio installation, as hereinafter provided, and in the case of a ship of the United States, unless there is on board a valid station license issued in accordance with this Act;

(2) For any passenger ship of the United States of five thousand gross tons, or over, to be navigated outside of a harbor or port, in the open sea, or for any such ship of the United States or any foreign country to leave or attempt to leave any harbor or port of the United States for a voyage in the open sea, unless such ship is equipped with an efficient radio direction finder apparatus (radio compass) properly adjusted in operating condition as hereinafter provided, which apparatus is approved by the Commission;

(b) A ship which is not subject to the provisions of this part at the time of its departure on a voyage shall not become subject to such provisions on account of any deviation from its intended voyage due to stress of weather or any other cause over which neither the master, the owner, nor the charterer (if any) has control.

Sec. 352. (a) The provisions of this part shall not apply to—

(1) A ship of war;
(2) A ship of the United States belonging to and operated by the Government, except a ship of the United States Maritime Commission, the Inland and Coastwise Waterways Service, or the Panama Railroad Company;
(3) A foreign ship belonging to a country which is a party to the Safety Convention and which ship carries a valid certificate exempting said ship from the radio provisions of that Convention, or which ship conforms to the radio requirements of such Convention or Regulations and has on board a valid certificate to that effect;
(4) Yachts of less than six hundred gross tons not subject to the radio provisions of the Safety Convention;
(5) Vessels in tow;
(6) A vessel navigating solely on the Great Lakes, or on any bays, sounds, rivers, or protected waters within the jurisdiction of the United States, or to a vessel leaving or attempting to leave any harbor or port of the United States for a voyage solely on the Great Lakes, or on any bays, sounds, rivers, or protected waters within the jurisdiction of the United States.

(b) The Commission may, if it considers that the route or the conditions of the voyage or other circumstances are such as to render a radio installation unreasonable or unnecessary for the purposes of this part, exempt from the provisions of this part any ship, or any class of ships, which falls within any of the following descriptions:

(1) Passenger ships which in the course of their voyage do not go more than twenty nautical miles from the nearest land or more than two hundred nautical miles between two consecutive ports;
(2) Cargo ships which in the course of their voyage do not go more than one hundred and fifty nautical miles from the nearest land;

(3) Passenger vessels of less than one hundred gross tons not subject to the radio provisions of the Safety Convention;

(4) Sailing ships.

OPERATORS, WATCHES, AUTO-ALARM

Sec. 353. (a) Each cargo ship required by this part to be fitted with a radio installation and which is not fitted with an auto-alarm, and each passenger ship required by this part to be fitted with a radio installation, shall, for safety purposes, carry at least two qualified operators.

(b) A cargo ship, required by this part to be fitted with a radio installation, which is fitted with an auto-alarm in accordance with this title, shall, for safety purposes, carry at least one qualified operator who shall have had at least six months' previous service in the aggregate as a qualified operator in a station on board a ship or ships of the United States.

(c) Each ship of the United States required by this part to be fitted with a radio installation shall, while being navigated outside a harbor or port, keep a continuous watch by means of qualified operators: PROVIDED, HOWEVER, That in lieu thereof on a cargo ship fitted with an auto-alarm in proper operating condition, a watch of at least eight hours per day, in the aggregate, shall be maintained by means of a qualified operator.

(d) The Commission shall, when it finds it necessary for safety purposes, have authority to prescribe the particular hours of watch on a ship of the United States required by this part to be fitted with a radio installation.

(e) On all ships of the United States fitted with an auto-alarm, said apparatus shall be in operation at all times while the ship is being navigated outside of a harbor or port when the operator is not on watch.

TECHNICAL REQUIREMENTS

Sec. 354. The radio installation and the radio direction-finding apparatus required by section 351 of this part shall comply with the following requirements:

(a) The radio installation shall comprise a main and an emergency or reserve installation: PROVIDED, HOWEVER, that on a cargo ship, if the main installation complies also with all the requirements of an emergency or reserve installation, the emergency or reserve installation may be omitted.

(b) The ship's radio operating room and the emergency or reserve installation shall be placed in the upper part of the ship in a position of the greatest possible safety and as high as practicable above the deepest load water line, and the location of such room or rooms shall be approved by the Bureau of Marine Inspection and Navigation, Department of Commerce.

(c) The main and emergency or reserve installations shall be capable of transmitting and receiving on the frequencies and types of waves designated by the Commission pursuant to law for the purpose of distress and safety of navigation.

(d) The main installation shall have a normal transmitting and receiving range of at least two hundred nautical miles, that is to say, it must be capable of transmitting and receiving clearly perceptible signals from ship to ship over a range of at least two hundred nautical miles by day under normal conditions and circumstances.

(e) Sufficient power shall be available at all times to operate the main radio installation efficiently under normal conditions over the range specified in subsection (d) of this section.

(f) The emergency or reserve installation shall include a source of energy independent of the propelling power of the ship and of any other electrical system and shall be capable of being put into operation rapidly and of working for at least six continuous hours. For the emergency or reserve installation, the normal range as defined in subsection (d) of this section shall be at least one hundred nautical miles.

(g) There shall be provided between the bridge of the ship and the radio room, and between the bridge and the location of the direction finding apparatus, when the direction finding apparatus is not located on the bridge, an efficient means of communication independent of any other communication system of the ship.

(h) The direction finding apparatus shall be efficient and capable of receiving clearly perceptible radio signals and of taking bearings from which the true bearing and direction may be determined. It shall be capable of receiving signals on the frequencies prescribed for distress, direction finding, and radio beacons by the General Radio Regulations annexed to the International Telecommunication Convention in force and in new installations after the effective date of this part, such other frequencies as the Commission may for safety purposes designate.

LIFEBOATS

Sec. 355. Every motor lifeboat, required to be equipped with radio by treaty or convention to which the United States is a party, by statute, or by regulation made in conformity with a treaty, convention, or statute, shall be fitted with an efficient radio installation under such rules and regulations as the Commission may find necessary to promote the safety of life.

NOTE:—Previous issues of the RCA REVIEW described various types of shipboard radio equipment which have been developed to meet the new requirements:—"Safety of Life at Sea", July, 1936, by Charles J. Pannill; "Automatic Alarm", January, 1937, by Byrnes and Martin; and "300-Watt Marine Radio Telegraph Transmitter", April, 1937, by Byrnes.

EFFECTS OF SPACE CHARGE IN THE GRID-ANODE REGION OF VACUUM TUBES*

BY

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Summary—*The effects of space charge in the region between grid and anode of a vacuum tube, for the case where the planes of the grid and plate are parallel, are determined from the results of a simple analysis. The main effects of the space charge are (a) to introduce departures from the linear potential distribution of the electrostatic case; (b) to set an upper limit, under certain conditions, for the anode current; (c) to introduce instabilities and hysteresis phenomena in the behavior of the tube; and (d) to increase the electron transit time in this region.*

Four modes of potential distribution which may exist in this region are treated: (1) Neither potential minimum nor virtual cathode exist; (2) potential minimum exists; (3) space-charge-limited virtual cathode exists; and (4) temperature-limited virtual cathode exists (negative anode potentials). For each of the various states of operation, expressions are derived for the distribution of potential and electric intensity throughout the region; the time of flight of electrons from grid to anode, and from grid to the plane of zero potential; and the location and magnitude of the minimum potential. An expression is also derived for the dependence of the anode current on the space current, grid-anode distance, grid voltage, and anode voltage. Curves are plotted from these expressions, and it is shown how the behavior of a large variety of practical tubes can be predicted and explained with their aid. The assumptions which underlie the theory are stated, and the effects of the neglected phenomena are discussed qualitatively.

Anode-current vs. anode-voltage and anode-current vs. space-current curves representing observations made on a specially constructed tetrode are presented by way of experimental verification of the theoretical results.

For purposes of illustration, application is made of these results to elucidate the theory of the type of power-amplifier tube which employs a minimum potential, formed in front of the anode as a result of the space charge of the electrons, to minimize the passage of secondary electrons from anode to grid. In addition, it is shown how the decrease of anode current with increasing space current, which occurs when a space-charge-limited virtual cathode is formed in the grid-anode region, may be utilized to provide negative transconductance amplifiers and oscillators.

I. INTRODUCTION

THE effects of space charge in the region between grid and anode are of fundamental importance in any complete theory of vacuum-tube behavior. This is particularly true for the type of tube in which the anode is preceded by a grid electrode operated at a positive potential. Among such tubes, we may mention the Barkhausen-Kurz triode oscillator, in which the grid is operated at a high positive potential and the anode is either negative or only slightly positive; the common tetrode r-f and a-f amplifiers, in which the screen

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grid is operated at a relatively high positive potential, and the anode potential is caused to vary from a high positive potential to a smaller positive potential, which may be less than the screen-grid potential; and more recent types of power-amplifier tubes, such as the RCA 6L6, in which a minimum potential which occurs in the space between screen grid and anode at relatively low anode potentials is employed to minimize the passage of secondary electrons (liberated at the anode) to the screen grid.[1, 2, 3, 4]

In this paper we wish to investigate theoretically, somewhat more completely than has been done hitherto, the properties of the grid-anode region.* The theory is also applicable, with suitable modifications, to the region between two grids, a problem which has also received some attention.[5] To this end, we postulate an idealized situation in which electrons pass through the interstices of a plane-accelerator grid, the effective potential of which is V_g, towards a parallel-plane anode separated a distance a from the grid and operated at a potential V_a. We assume that all of the electrons have been emitted from the cathode with zero velocity, and that they pass through the grid with a uniform velocity corresponding to the potential V_g. In general, we cannot assume that all of these electrons will pass on to the anode, for the potential distribution in the grid-anode region may be such that some, or even all of the electrons, will be turned back towards the grid at some intermediate point in this region. These returning electrons may be expected to execute excursions about the grid, a portion being absorbed on each passage through the grid.[6, 7, 8, 9] In view of the uncertainty of the exact fractional absorption of the electrons by the grid on each complete passage, and to simplify matters, we shall assume that only two groups of electrons contribute to the space current: one group consists of the electrons which initially pass through the grid towards the anode, corresponding to a current I_0; and the second group consists of the electrons which may be turned back at an intermediate point in the grid-anode region, corresponding to a current I_0-I_A. Generally speaking, we are interested in the dependence of the anode current upon V_a, V_g, a, and I_0; the time of flight of the electrons; the distribution of potential and electric intensity throughout the region between the grid and anode; and the location and magnitude of the minimum potential which may occur in this region.

As stated above, the problem is only an approximation to the actual

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[1, 2, 3, 4, etc.] Numbers refer to bibliography.

* While we were preparing the results of our investigation for publication, our attention was called to four recent papers dealing with this subject. (Bibliography 13, 14, 15 and 16).

case. The complete treatment of the problem would require taking into account such things as the Maxwellian velocity distribution of emitted electrons; the tangential velocities of the electrons due to grid-wire deflections; the distribution of velocities of the electrons which pass through the grid, due to the variation of the potential in the plane of the grid; the presence of secondary electrons liberated at the anode and accelerator grid; multiple passages of electrons about the grid, etc. However, the solution of the problem set forth will ordinarily suffice for a first-order answer, since the effects of these neglected factors can frequently be inferred, at least, from the physics of the situation.

II. THEORY

The equation which must be solved to determine the properties we seek is:

$$\frac{d^2V}{dx^2} = 4\pi\rho \qquad (\rho \text{ negative}) \tag{1}$$

together with

$$\tfrac{1}{2}\, mv^2 = eV \tag{2}$$

and

$$I = \rho v \tag{3}$$

In this problem we do not concern ourselves with the effects of variations of the potential with time. Consequently we may assume that both forward-moving and returning electrons always have the same velocity at any one point, and thus that the total space charge in the grid-anode region depends only upon the absolute values of the corresponding forward and return currents. For these reasons, we may define the quantities used in (1), (2) and (3) as

$V =$ potential at any distance, x
$x =$ distance from grid
$\rho =$ charge density at x
$e =$ electron charge
$m =$ mass of electron
$v =$ absolute value of velocity at x
$I =$ sum of the absolute values of the current densities at x

Further on we shall have occasion to introduce the following additional symbols:

$E =$ electric intensity at x
$H = 16\pi \left(\frac{m}{2e} \right)^{1/2} I_0$, where I_0 is the current density corresponding to the forward-moving electrons
$\tau =$ transit time for an electron passing from grid to anode
$\tau' =$ transit time for an electron passing from grid to point of zero potential

$$P = \frac{3H^{1/2}x}{4V_g{}^{3/4}}$$

$$P_a = \frac{3H^{1/2}a}{4V_g{}^{3/4}}, \qquad \left[P_a = \frac{654\, I_o{}^{1/2} \left(\dfrac{\text{amp.}}{\text{cm}^2} \right)^{1/2} a \text{ (cm)}}{V_g{}^{3/4} \text{ (volts)}^{3/4}} \right]$$

$$P_c = \frac{3H^{1/2}c}{4V_g{}^{3/4}} \qquad Q_g = \frac{E_g}{H^{1/2}\, V_g{}^{1/4}}$$

$$Q = \frac{E}{H^{1/2}\, V_g{}^{1/4}} \qquad Q_c = \frac{E_c}{H^{1/2}\, V_g{}^{1/4}}$$

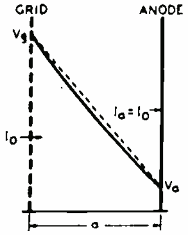

Fig. 1—Potential distribution between grid and anode typical of the case when neither potential minimum nor virtual cathode exist. All of the electrons which pass through the grid are collected at the anode.

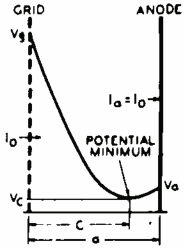

Fig. 2—Potential distribution illustrating the conditions which exist when a potential minimum is formed in the grid-anode region. All of the electrons which pass through the grid are collected at the anode.

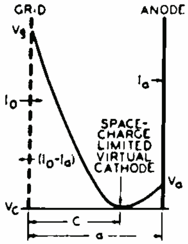

Fig. 3—This potential distribution represents the case when a space-charge-limited virtual cathode exists. Only part of the electrons which pass through the grid are collected at the anode; the others are returned towards the grid at the virtual cathode.

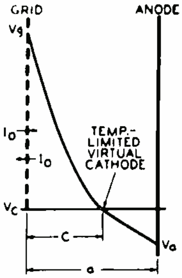

Fig. 4—This potential distribution illustrates the case when a temperature-limited virtual cathode exists. All of the electrons which pass through the grid are subsequently returned towards the grid at the virtual cathode.

The boundary conditions for the various states of operation are given in Figures 1, 2, 3, and 4, which depict typical potential distributions for these states of operation.

By combining (1), (2) and (3), and performing the first integration of (1) from any point x_1, to another point x, we get for the electric intensity at x,

$$\text{E} = \pm \left[E_1{}^2 - HV_1{}^{1/2} + HV^{1/2} \right]^{1/2} \tag{4}$$

A second integration gives

$$\frac{3H^2}{4}(x - x_1) = \pm [E_1^2 - HV_1^{1/2} + HV^{1/2}]^{1/2} \cdot [HV^{1/2} - 2(E_1^2 - HV_1^{1/2})] - E_1 [3HV_1^{1/2} - 2E_1^2] \tag{5}$$

The time of flight of an electron from x_1 to x is

$$\tau - \tau_1 = \int_{x_1}^{x} \frac{dx}{v} = \left(\frac{m}{2e} \right)^{1/2} \cdot \int_{x_1}^{x} \frac{dx}{V^{1/2}} \tag{6}$$

From (1), (2) and (3)

$$V^{-1/2} = \frac{1}{4\pi \left(\dfrac{m}{2e} \right)^{1/2} I} \cdot \frac{dE}{dx}$$

so that (6) becomes

$$\tau - \tau_1 = \frac{1}{4\pi I} \int_{x_1}^{x} \frac{dE}{dx} \cdot dx = \frac{1}{4\pi I} \cdot (E - E_1) \tag{7}$$

These general equations will now be applied to the various special states of operation.

1—*Neither Potential Minimum Nor Virtual Cathode Exist.*—This mode of operation is depicted in Figure 1. The situation is intermediate between the case when no electrons are present in the grid-anode region, for which condition the potential distribution is linear, and the case when enough electrons are present in this region so that the electric intensity is zero at the grid or anode. In this mode, all of the electrons which pass through the grid are collected at the anode.

Usually, I_o, a, V_g, and V_a are specified, and it is required to determine the distribution of the electric intensity and potential throughout the region and the time of flight of an electron from grid to anode. To start with, we must find E_g. This can be done by substituting the conditions that $V = V_a$ when $x = a$, and $E = E_g$ and $V = V_g$ when $x = 0$ in (5). Thus,

$$\frac{3H^2 a}{4} = \pm [E_g^2 - H\,(V_g^{1/2} - V_a^{1/2})]^{1/2} \cdot [HV_a^{1/2} - 2(E_g^2 - HV_g^{1/2})] - E_g [3HV_g^{1/2} - 2E_g^2] \tag{8}$$

This equation allows the determination of E_g in terms of I_o, a, V_g, and V_a. To put the equation into somewhat more convenient form for plotting, divide both sides by $H^{3/2}V_g^{3/4}$. We get an expression involving the dimensionless parameters, P_a and Q_g:

$$P_a = \pm \left[Q_g^2 - 1 + \left(\frac{V_a}{V_g} \right)^{1/2} \right]^{1/2} \cdot \left[\left(\frac{V_a}{V_g} \right)^{1/2} + 2 - 2Q_g^2 \right] - Q_g \left[3 - 2Q_g^2 \right] \tag{8a}$$

A curve of P_a vs. $\dfrac{V_a}{V_g}$ for various values of Q_g may now be obtained very simply from (8a). This is shown in Figure 5.

For any given set of the four variables, Q_g can be located in Figure 5. For values of $\dfrac{V_a}{V_g}$ greater than unity, Q_g will range for this case, from zero to positive values; for values of $\dfrac{V_a}{V_g}$ which lie between unity and zero, Q_g will range from zero to negative values.

The potential distribution can now be determined by means of (5) by substituting the condition that $E_1 = E_g$ and $V_1 = V_g$ when $x_1 = 0$. Doing this, and rearranging, there results

$$P = \pm \left[Q_g^2 - 1 + \left(\frac{V}{V_g} \right)^{1/2} \right]^{1/2} \cdot \left[\left(\frac{V}{V_g} \right)^{1/2} + 2 - 2Q_g^2 \right] - Q_g \left[3 - Q_g^2 \right] \tag{9}$$

This will be recognized as being similar to (8a) except that P replaces P_a, and V replaces V_a. Therefore, the same data can be utilized to study the variation of $\dfrac{V}{V_g}$ with P, (to which the distance from the grid, x, is proportional) for various values of Q_g. Figure 5 thus represents, also, generalized potential distribution curves.

The electric intensity at any point may be determined by means of (4) by substituting the condition that $E_1 = E_g$ when $V_1 = V_g$. This gives, after some rearrangement,

$$Q = \pm \left[Q_g^2 - 1 + \left(\frac{V}{V_g} \right)^{1/2} \right]^{1/2} \tag{10}$$

With the aid of (8a) and (9), (10) can be used to study the variation of Q (which is proportional to the electric intensity at any point, x) as a function of P, for various values of $\frac{V_a}{V_g}$.

The time of flight, τ, of an electron passing from grid to anode is, from (7)

$$\tau = \frac{1}{4\pi I_o} (E_a - E_g) \tag{7a}$$

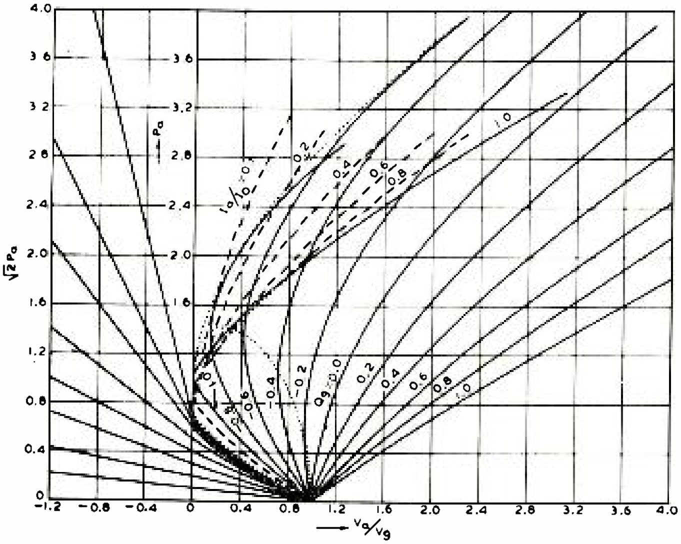

Fig. 5—Generalized potential distribution in the grid-anode region. This plot may be used to determine the particular mode in which the tube is operating, the anode current, and the electric intensity and potential distribution within the grid-anode region. Here,

$$P_a = 654 \cdot \frac{I_o^{1/2} \left(\frac{\text{amp.}}{\text{cm}^2}\right)^{1/2} \cdot a\ (\text{cm})}{V_g^{3/4}\ (\text{volts})^{3/4}}.$$

Equation (7a) can be put into more convenient form for plotting by making use of (4). After some rearrangement this gives

$$\tau = \frac{\tau_o}{P_a} \left\{ \pm \left[Q_g^2 - 1 + \left(\frac{V_a}{V_g} \right)^{1/2} \right]^{1/2} - Q_g \right\} \tag{11}$$

In (11) $\tau_o = \left(\frac{m}{2e} \right)^{1/2} \cdot \frac{3a}{V_g^{1/2}}$ represents the time of transit of an electron in a space-charge-limited diode, the electrodes of which coincide with the grid and anode, and for which the electric intensity at the cathode is zero and the anode voltage is V_g. The ratio $\frac{\tau}{\tau_0}$ is plotted in Figure 6, the necessary data for the curves being obtained from Figure 5.

It is of interest to note that when $I_0 \rightarrow 0$, the limiting case for which the potential distribution is linear, (11) becomes

$$(\tau)_{I=0} = \left(\frac{m}{2e}\right)^{1/2} \cdot \frac{2a}{V_g^{1/2}} \cdot \frac{1}{\left(\frac{V_a}{V_g}\right)^{1/2} + 1} \tag{11a}$$

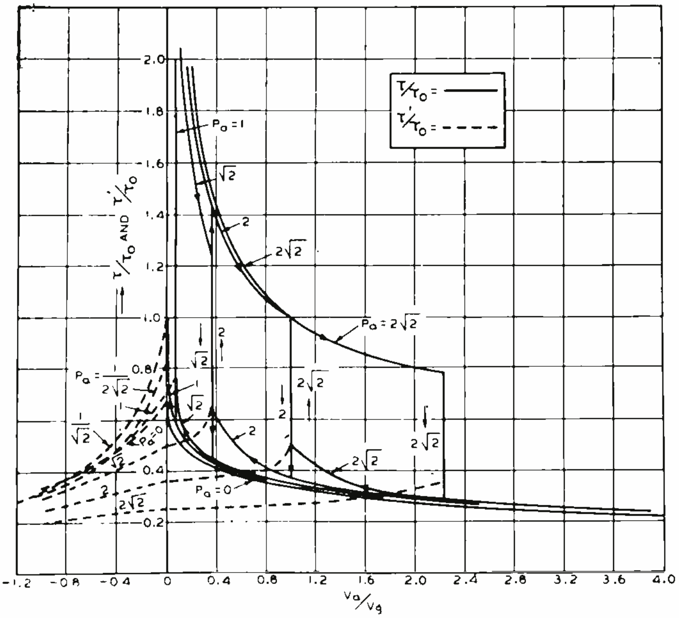

Fig. 6—Dependence of transit time upon anode voltage for various values of parameter P_a. τ is the transit time for electrons which travel from the grid to the anode. τ' is the electron transit time from grid to virtual cathode. $\tau_0 \text{ (sec)} = \frac{3a \text{ (cm)}}{5.95 \cdot 10^7 \, V_g^{1/2} \text{ (volts)}^{1/2}}$ is the electron transit time in a space-charge-limited diode.

The other limiting cases for the potential distributions under consideration here occur when $E_g = 0$, representing a potential minimum at the grid, or when $E_a = 0$, representing a potential minimum at the anode. In the first case

$$(\tau)_{E_g=0} = \left(\frac{m}{2e}\right)^{1/2} \cdot \frac{3a}{V_a^{1/2} + 2V_g^{1/2}} \tag{11b}$$

while in the second case

$$(\tau)_{E_a=0} = \left(\frac{m}{2e} \right)^{1/2} \cdot \frac{3a}{V_g^{1/2} + 2V_a^{1/2}} \tag{11c}$$

2—*Potential Minimum Exists.*—This mode of operation is shown in Figure 2. The situation existing here is intermediate between the case when enough space current flows through the grid so that the electric intensity is zero at the grid or anode, and the case when the conditions are such that the electric intensity *and* the potential are just zero at either electrode or in the intervening region. In this mode, also, all of the electrons which pass through the grid are collected by the anode.

Here again, I_0, a, V_g and V_a are usually specified, and it is required to find the distribution of the electric intensity and potential throughout the region, the time of flight of an electron from grid to anode, and the value and location of the minimum potential.

We start by substituting the condition that $V = V_g$ and $E = E_g$ when $x = 0$, and $V = V_a$ and $x = a$ in (5). This leads to an expression identical with (8a); the variation with $\frac{V_a}{V_g}$ of P_a for various values of Q_g is shown in Figure 5. For this mode Q_g ranges between 0 and -1 for all values of $\frac{V_a}{V_g}$, as shown in Figure 5. It is important to note that for one value of $\frac{V_a}{V_g}$ there can be two values of Q_g, corresponding to one value of P_a. We shall discuss this peculiarity further on.

In the same way, the expression for the potential distribution can be shown to be formally identical to (9), and the variation of $\frac{V}{V_g}$ with P for various values of Q_g can also be studied from Figure 5. For this mode of operation, $\frac{V}{V_g}$ passes through a minimum, representing the potential minimum.

The expression for the electric intensity is derivable, again, from (4); this is identical with (10). Now E is negative to the left of the potential minimum, zero at the potential minimum, and positive to the right of the potential minimum.

The time of flight, τ, of an electron from grid to anode is again given by (11), and the ratio $\dfrac{\tau}{\tau_0}$ is also plotted in Figure 6. It is of interest to note that for the limiting condition, when $V_c = 0$, (11) gives

$$\tau = \left(\frac{m}{2e}\right)^{1/2} \cdot \frac{3a}{V_g^{1/2}} \cdot \frac{1}{\left[1 - \left(\dfrac{V_a}{V_g}\right)^{1/4} + \left(\dfrac{V_a}{V_g}\right)^{1/2}\right]} \tag{11d}$$

The ratio of (11a) to (11d) gives us an estimate of the increase of the transit time for two important limiting cases which may be encountered. This is

$$\frac{(\tau)_{Vc=0}}{(\tau)_{I0=0}} = \frac{2}{3} \cdot \left\{1 - \frac{1}{\left(\dfrac{V_a}{V_g}\right)^{1/4} + \left(\dfrac{V_g}{V_a}\right)^{1/4}}\right\} \tag{12}$$

Thus when $V_a = V_g$, the transit time for the case when a zero-potential minimum is just formed is three times larger than the transit time for very low values of anode current.

It is of considerable interest, in any consideration of this particular mode of operation, to determine the value and location of the minimum potential. This may be done indirectly by making use of Figure 5. Alternately, an implicit expression for the value of $\dfrac{V_c}{V_g}$ may be derived from (4) and (5). This is

$$P_a = \pm\left[1 - \left(\frac{V_c}{V_g}\right)^{1/2}\right]^{1/2} \cdot \left[1 + 2\left(\frac{V_c}{V_g}\right)^{1/2}\right] \pm \left[\left(\frac{V_a}{V_g}\right)^{1/2} - \left(\frac{V_c}{V_g}\right)^{1/2}\right]^{1/2} \cdot \left[\left(\frac{V_a}{V_g}\right)^{1/2} + 2\left(\frac{V_c}{V_g}\right)^{1/2}\right] \tag{13}$$

The variation of $\dfrac{V_c}{V_g}$ with $\dfrac{V_a}{V_g}$, for several values of P_a is shown in Figure 7.

In the same way, by making use of (4) and (5), the location of the minimum potential can be expressed in terms of $\frac{V_c}{V_g}$ and P_a, i.e.,

$$\frac{c}{a} = \frac{1}{P_a} \cdot \left[1 + 2 \left(\frac{V_c}{V_g} \right)^{\frac{1}{2}} \right] \cdot \left[1 - \left(\frac{V_c}{V_g} \right)^{\frac{1}{2}} \right]^{\frac{1}{2}} \tag{14}$$

The dependence of $\frac{c}{a}$ on $\frac{V_a}{V_g}$, for various values of P_a is shown in Figure 8.

Figures 7-11 are of the greatest importance in delineating the properties of the grid-anode region. We shall make extensive use of these curves in our discussion.

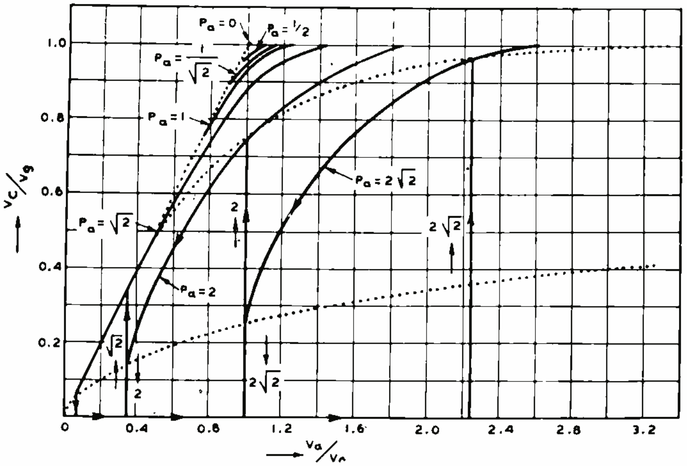

Fig. 7—Variation of the magnitude of the potential minimum with anode potential for several values of the parameter P_a.

Before proceeding to the next case, let us consider what happens when, with V_a, V_g, and a fixed, the current density I_0 is increased from zero until the potential minimum which is formed in the grid-anode region descends to the value zero. Figure 9, plotted from (13), shows the variation of $\frac{V_c}{V_g}$ with P_a^2 for various values of $\frac{V_a}{V_g}$. The rate of change of the minimum potential with respect to space current is given by the slope of these curves. Differentiating P_a in (13) with respect to $\left(\frac{V_c}{V_g}\right)^{\frac{1}{2}}$, considering $\left(\frac{V_a}{V_g}\right)^{\frac{1}{2}}$ fixed, and equating the derivative to zero tells us that $(P_a)_{\text{max.}}$ occurs when

$$\frac{V_c}{V_g} = \frac{\dfrac{V_a}{V_g}}{\left[1+\left(\dfrac{V_a}{V_g}\right)^{1/2}\right]^2} \tag{15}$$

Substituting (15) in (13), we find that the maximum current (density) which can be passed is

$$(I_a)_{\text{max.}} = \frac{1}{9\pi a^2} \cdot \left(\frac{2e}{m}\right)^{1/2} \cdot (V_g^{1/2} + V_a^{1/2})^3 \tag{13a}$$

The value of the potential minimum for this maximum current is given by (15). Equation (15) is plotted as a dotted line in Figure 7. Figure 11 is a plot of $\left(\dfrac{I_a}{I_o}\right)$. P_a^2 vs. P_a^2 and shows the anode current increasing directly with the space current until the maximum value, given by (13a), is reached. At this point the potential minimum descends abruptly to zero, and a new state of operation sets in.

3—*Space-Charge-Limited Virtual Cathode Exists.*—This state of operation is shown in Figure 3. The situation existing here is such that at some plane in the region between the two electrodes, or possibly in the plane at either electrode, the electric intensity *and* the potential are zero. Such a point may be called a space-charge-limited virtual cathode, since it turns out that it behaves in many respects like a space-charge-limited thermionic cathode. It is pertinent to call attention to the difference between the virtual cathode of the present case and the potential minimum of the preceding case. In general, the potential minimum merely calls for a zero of electric intensity, but not of potential, since all of the electrons which flow through the grid are collected at the anode. However, when there is a point of zero potential in the region between the two electrodes, some of the electrons passing towards the anode may be turned back towards the grid at this point.

As a digression, we may study briefly the limiting case, when the anode current is exactly equal to the space current. We have, on putting

$\dfrac{V_c}{V_g} = 0$ in (13),

$$I_a = \frac{1}{9\pi a^2} \cdot \left(\frac{2e}{m}\right)^{1/2} \cdot (V_g^{3/4} + V_a^{3/4})^2 \tag{13b}$$

This is smaller than the current predicted by (13a), the ratio of (13a) to (13b) being

$$\frac{\left[1+\left(\dfrac{V_a}{V_g}\right)^{1/2}\right]^3}{\left[1+\left(\dfrac{V_a}{V_g}\right)^{3/4}\right]^2},$$

the maximum value of this ratio occurring when $\dfrac{V_a}{V_g}=1$. The ratio is equal to unity when $\dfrac{V_a}{V_g}=0$ or ∞. The difference between (13a) and (13b) is the basis of the hysteresis phenomena characteristic of the curves shown in Figures 5-11, and will be discussed further on.

We return now to the general case, for which some of the electrons are turned back towards the grid, and assume for simplicity that such electrons never again re-enter the grid-anode region.* The anode current is then, in view of (1), (2), (3), and the requisite boundary conditions, given by

$$I_a=\frac{1}{9\pi}\cdot\left(\frac{2e}{m}\right)^{1/2}\cdot\frac{V_a^{3/2}}{(a\text{-}c)^2} \qquad (16)$$

The total space charge in the region between the grid and the virtual cathode depends upon both the forward-moving and returning electrons, and since these electrons have the same velocity at any one point, we may write

$$2I_o-I_a=\frac{1}{9\pi}\cdot\left(\frac{2e}{m}\right)^{1/2}\cdot\frac{V_g^{3/2}}{c^2} \qquad (17)$$

If we divide (17) by I_o, and put $H=16\pi\left(\dfrac{m}{2e}\right)^{1/2}\cdot I_o$ and $P_a=\dfrac{3H^{1/2}a}{4V_g^{3/4}}$ as before, we get, on rearranging,

$$\frac{c}{a}=\frac{1}{P_a\left(2-\dfrac{I_a}{I_o}\right)^{1/2}} \qquad (18)$$

It is of interest to note that when $I_a=0$, (18) becomes $\dfrac{c}{a}=\dfrac{1}{2^{1/2}P_a}$.

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* The ensuing treatment can easily be generalized to include the multiple-passage case. See, for example, 7, 9 of Bibliography.

That is, when $V_a = 0$, the virtual cathode (if one exists) is at a distance $c = \dfrac{a}{2^{1/2} P_a}$ cm from the grid.

In order to use (18) to plot the location of the virtual cathode as a function of I_o, a, V_g and V_a, we must first derive an expression for $\dfrac{I_a}{I_o}$ as a function of $\dfrac{V_a}{V_g}$ and P_a. This can be done by dividing (17) by (16) and making use of (18). There results

$$\left(\frac{V_a}{V_g}\right)^{3/4} = \left(\frac{I_a}{I_o}\right)^{1/2} \cdot \left[P_a - \frac{1}{\left(2 - \dfrac{I_a}{I_o}\right)^{1/2}}\right] \tag{19}$$

Equation (19) is used (a) to plot P_a vs. $\dfrac{V_a}{V_g}$ for various values of $\dfrac{I_a}{I_o}$, as shown in Figure 5; (b) to plot $\dfrac{I_a}{I_o}$ vs. $\dfrac{V_a}{V_g}$ for various values of P_a, as shown in Figure 10;* and (c) to plot $\left(\dfrac{I_a}{I_o}\right) \cdot P_a^2$ vs. P_a^2 for various values of $\dfrac{V_a}{V_g}$, as shown in Figure 11. In addition, by making use in (18) of the data furnished by Figure 10, we are enabled to plot $\dfrac{c}{a}$ vs. $\dfrac{V_a}{V_g}$ for various values of P_a, as shown in Figure 8.

The potential distribution within the grid-anode region corresponding to any particular value of c is $\dfrac{V}{V_g} = \left(\dfrac{c-x}{c}\right)^{4/3}$ to the left of c, and $\dfrac{V}{V_a} = \left(\dfrac{x-c}{a-c}\right)^{4/3}$ to the right of c. The corresponding electric intensities are $E = -\dfrac{4}{3} \cdot \dfrac{V}{c-x}$, and $E = +\dfrac{4}{3} \cdot \dfrac{V}{a-c}$, respectively.

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* Curves similar to those given in Figure 10 have been given by Tonks[7]; his curves however, are restricted to the virtual-cathode case and do not portray the important hysteresis-effects characteristic of the minimum-potential case.

The time of flight of an electron which travels from the grid to the virtual cathode is simply

$$\tau' = \left(\frac{m}{2e}\right)^{\frac{1}{2}} \cdot \frac{3c}{V_g^{\frac{1}{2}}} = \frac{c}{a} \cdot \tau_o \tag{20}$$

The time of flight of an electron which travels from grid to anode is

$$\tau = \left(\frac{m}{2e}\right)^{\frac{1}{2}} \cdot \left[\frac{3c}{V_g^{\frac{1}{2}}} + \frac{3\,(a\text{-}c)}{V_a^{\frac{1}{2}}}\right]$$

$$= \tau_o \cdot \left\{\left(\frac{V_g}{V_a}\right)^{\frac{1}{2}} + \frac{c}{a} \cdot \left[1 - \left(\frac{V_g}{V_a}\right)^{\frac{1}{2}}\right]\right\} \tag{21}$$

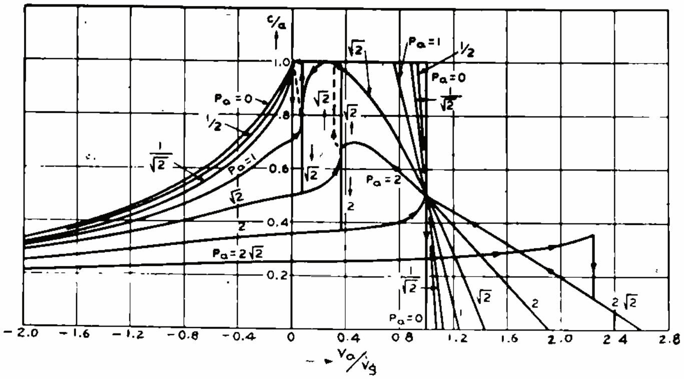

Fig. 8—Position of the potential minimum and virtual cathode as a function of the anode potential for various values of the parameter P_a.

The time of flight for each class of electrons can be plotted with the aid of (18), as shown in Figure 6.

4—*Temperature-Limited Virtual Cathode Exists (Negative Anode Potentials)*.—This state of operation is illustrated in Figure 4. The situation now is such that the electric intensity is negative throughout the region between the two electrodes, and at some intermediate plane the potential is zero. The electrons which move through the interstices of the grid towards the plane of zero potential are gradually decelerated, and finally turned back at this plane toward the grid. Since the electric intensity at this plane is always finite, the potential distribution between c and o resembles that of a temperature-limited diode. It turns out, moreover, that the properties of the plane of zero potential are in many respects identical to that of the temperature-limited cathode. Accordingly, we shall designate this plane as a "temperature-limited virtual cathode", by analogy with the "space-charge-limited virtual cathode".

Here I_o, a, V_g, and V_a are specified and it is required to determine the distribution of the electric intensity and potential throughout this region, the position of the virtual cathode, and the time of flight of an electron from the grid to the virtual cathode. We shall not attempt a rigorous treatment, but instead we shall base our analysis on the same assumptions as we have made for the preceding cases.

Since the anode is negative, no electrons will be collected. (This is a consequence of our assumptions regarding the initial velocities of the electrons emerging from the grid plane: in actuality there is always a very small anode current as a result of a distribution of initial velocities.) Furthermore, since the electric intensity must be continuous, and since there is no charge in the region $c < x < a$, the electric intensity, E_c, at the virtual cathode (and also to the right of this plane) will be equal to $\dfrac{V_a}{a\text{-}c}$·

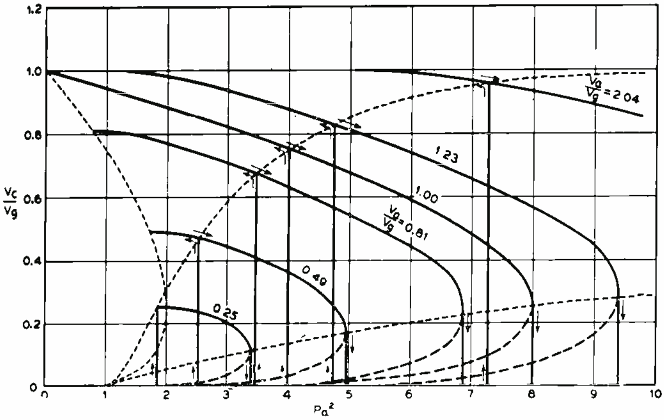

Fig. 9—Variation of the magnitude of the potential minimum with P_a^2 (to which the space current is proportional) for several values of $\dfrac{V_a}{V_g}$·

We now make use of (4) and (5), subject to the appropriate boundary conditions. In (4) we put $E_1 = E_c$, $E = E_g$, $V_1 = 0$, $V = V_g$, and $I = 2I_o$. There results

$$E_g = -\left[E_c^2 + 2HV_g^{1/2}\right]^{1/2} \qquad (22)$$

In (5) we put $x = c$, $x_1 = 0$, $E_1 = E_g$, $V_1 = V_g$, and $V = 0$ getting, with the aid of (22)

$$\frac{3H^2c}{2} = \mp\left[E_c^3 + (E_c^2 - HV_g^{1/2}).\,(E_c^2 + 2HV_g^{1/2})^{1/2}\right] \qquad (23)$$

Dividing through by $2H^{3/2}. V_g^{3/4}$ and making use of our abbreviated symbolism, we get

$$P_c = \mp \frac{Q_c^3 + (Q_c^2 - 1).(Q_c^2 + 2)^{1/2}}{2} \tag{23a}$$

Now,

$$P_c\,Q_c = \frac{3H^{1/2}c}{4V_g^{3/4}} \cdot \frac{V_a}{(a\text{-}c)\;H^{1/2}\,V_g^{1/4}} = \frac{3}{4} \cdot \left(\frac{V_a}{V_g}\right) . \frac{\dfrac{c}{a}}{1 - \dfrac{c}{a}} = P_a \,.\, Q_c \frac{c}{a}$$

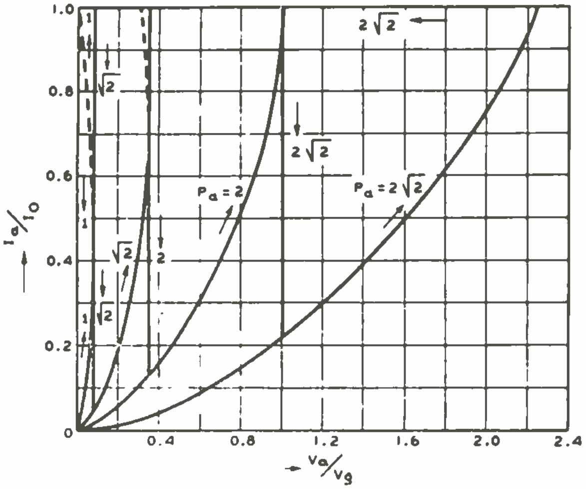

Fig. 10—Dependence of anode current upon anode potential for various values of the parameter P_a.

so that

$$\frac{c}{a} = \frac{P_c}{P_a} = 1 - \frac{3}{4} \cdot \frac{\left(\dfrac{V_a}{V_g}\right)}{P_a{\cdot}Q_c} \tag{24}$$

and

$$P_a = P_c + \frac{3}{4} \cdot \frac{1}{Q_c} \cdot \left(\frac{V_a}{V_g}\right) \tag{24a}$$

We may now plot P_c as a function of Q_c by means of (23a). The corresponding value of P_a for any given value of $\dfrac{V_a}{V_g}$ may then be found from (24a). This gives us sufficient information to plot $\frac{c}{a}$ as a function of $\frac{V_a}{V_g}$ for various values of P_a: such curves are included in Figure 8. We are also enabled to plot, from this information, P_a vs. $\frac{V_a}{V_g}$ for various values of Q_g: this is shown in Figure 5, which may be used again as a generalized potential distribution plot.

The electric intensity at any point is given by

$$Q = -\left[Q_g{}^2 + 2\left(\frac{V}{V_g}\right)^{1/2} - 2\right]^{1/2} \tag{25}$$

The time of transit of electrons from grid to virtual cathode is, from (7) and (25),

$$\tau' = \frac{\tau_0}{2P_a} \cdot [Q_g + (Q_g - 2)^{1/2}] \tag{26}$$

The ratio $\frac{\tau'}{\tau_0}$ vs. $\frac{V_a}{V_g}$ for various values of P_a is shown in Figure 6. The limiting case, $I_0 \rightarrow o$, of (26) is

$$\tau' = \tau_0 \cdot \frac{2}{3} \cdot \frac{1}{1 - \left(\frac{V_a}{V_g}\right)} \tag{26a}$$

VI. DISCUSSION OF THEORETICAL RESULTS

We now proceed to interpret the results of the foregoing analysis. In the interests of clarity we shall lead up to the general discussion of Figures 5-11 by providing a preliminary verbal description of the results of varying first the space current and then the anode voltage for several representative cases of interest.

A—*Preliminary Discussion of Several Particular Situations*—

1. Effects of Varying I_0, when $\frac{V_a}{V_g} = 1$.

To begin with, let us suppose that the effective grid voltage, V_g, the anode voltage, V_a, and the grid-anode distance, a, are fixed, and that it is required to determine the effects of increasing the space current, I_0. This situation corresponds, for example, to the case of a tetrode, the screen-grid voltage and the anode voltage of which are fixed, and the control-grid voltage of which is varied. To fix ideas, we choose $\frac{V_a}{V_g} = 1$: a similar interpretation will hold for any other ratio.

When the control-grid bias is adjusted to cut off the cathode current, the potential distribution is linear, as shown by (*a*) of Figure 12. Decreasing the bias causes electrons to flow into the grid-anode region,

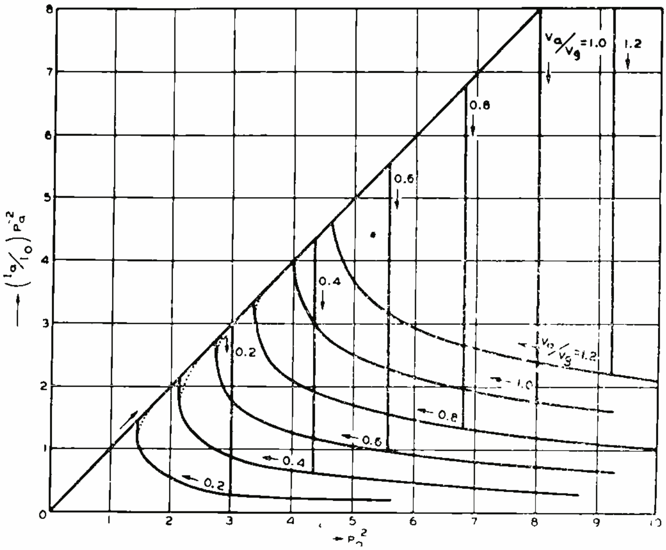

Fig. 11—Variation of the anode current with space current for several values of $\frac{V_a}{V_g}$.

and the potential is depressed, as shown by (*b*). A potential minimum is formed mid-way between the grid and anode. As the space current is increased, the potential minimum gradually descends until it reaches a value equal to 0.25 times the grid voltage, as shown in Figure 9. Up to this point the anode current increases directly with the space current, as shown in Figure 11. All of the electrons coming through the grid are collected at the anode. If, now, any additional electrons are permitted to pass into the grid-anode region, the potential minimum drops abruptly to zero, mid-way between the grid and anode. This is shown in Figure 9, the rate of change of the minimum potential with space current being infinite at this point. Since the minimum is now zero, some of the electrons are returned towards the grid, thus decreasing the space charge between the potential minimum and the anode, and increasing the space charge between the grid and the potential minimum. This causes an alteration of the potential distribution, the minimum (now a virtual cathode) shifting abruptly towards the grid until it reaches a distance from the grid equal to 0.265 times the grid-anode distance, shown by (*d*) of Figure 12. Thereafter, as the space current is increased, the anode current decreases gradually, in accordance with Figure 11, and the virtual cathode retreats toward the grid. The effect of increasing the space current on the time of flight of an electron from grid to anode, and from grid to virtual cathode, is shown in Figure 13.

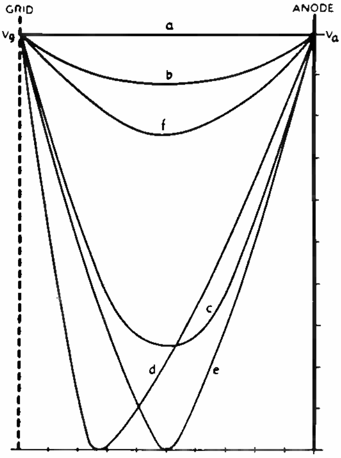

Fig. 12—Potential distribution for several values of space current. $\dfrac{V_a}{V_g} = 1.$

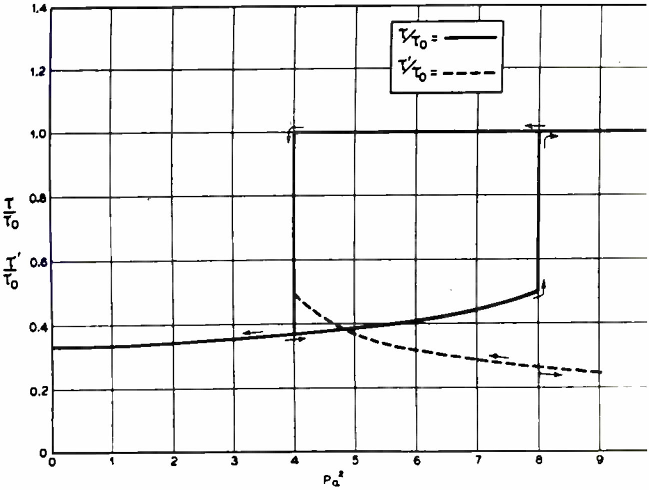

Fig. 13—Variation of electron transit time with space current. $\dfrac{V_a}{V_g} = 1.$

It will be observed that the transit time is initially one-third that of a space-charge-limited diode, the electrodes of which coincide with the grid and anode and for which the anode voltage is $V_g = V_a$. With increasing space current the transit time rises slowly until a current corresponding to $P_a^2 = 8$ is reached. At this point a virtual cathode is formed, and the transit time rises abruptly from $\tau = 0.500\tau_0$ to $\tau = \tau_0$. It remains constant thereafter at this value with further increase of space current. The transit time for the electrons which return toward the grid when a virtual cathode is formed is initially $\tau' = 0.265\tau_0$, and decreases slowly with increasing space current, since the virtual cathode retreats toward the grid. The time of flight of electrons from grid to virtual cathode and back to grid is twice τ'.

If the space current is now gradually decreased, the anode current will be found to increase, as shown in Figure 11. The operation now, however, does not take place entirely along the original curve. This is due to the fact that, originally, operation began with a potential minimum; now it takes place with a virtual cathode. When the space current reaches a value equal to one-half of the current at the transition value, where the potential minimum had previously shifted abruptly into a virtual cathode, the anode current becomes equal to the space current. The virtual cathode is now saturated. At this point the virtual cathode is mid-way between the grid and the anode, as shown by (*e*) of Figure 12. The slightest further reduction of space current causes the virtual cathode to shift abruptly into a potential minimum, the value of which is 0.75 times the grid voltage, as shown by (*f*) of Figure 12. Any further reduction of space current is accompanied by a proportionate decrease of anode current, the potential minimum gradually approaching (*a*) of Figure 12. The effect of decreasing the space current on the time of flight of an electron from grid to anode, and from grid to virtual cathode, is also shown in Figure 13.

The time of flight for electrons moving from grid to anode remains constant until the virtual cathode disappears; this occurs when the space current is decreased to a value corresponding to $P_a^2 = 4$. At this point the transit time drops abruptly to a value $\tau = 0.366\tau_0$. With further decrease of space current the transit time decreases slowly until the limiting value $\tau = 0.333\tau_0$ is reached for $P_a^2 \rightarrow 0$. The time of flight for the electrons which return from the virtual cathode increases slowly with decreasing space current, because the virtual cathode moves out slowly towards the anode. When $P_a^2 = 4$, the virtual cathode disappears and the transit time for the electrons which have been returning for space currents slightly greater than this value, is $\tau' = 0.500\tau_0$.

2. Effects of Varying I_0, when $\dfrac{V_a}{V_g} = 0.20$.

In the foregoing we have described in detail what occurs when the space current between the grid and anode is first increased, and then decreased. But this description was for the particular case when the grid and the anode voltages were equal. Similar phenomena, however, occur for any other ratio of $\dfrac{V_a}{V_g}$, with only slight differences. For example, suppose $\dfrac{V_a}{V_g} = 0.20$. Beginning with $I_0 = 0$, the potential distribution is linear. As I_0 is increased, the behavior of the grid-anode region is initially characteristic of the first mode of operation, i.e., neither potential minimum nor virtual cathode exist. Further increase of I_0 finally causes a potential minimum to occur at the anode, the value of this potential minimum being equal to the anode potential. The corresponding value of P_a is given by equation (13) as $P_a = 1.41$, the equivalent space-current density being $I_0 = 4.62 \times 10^{-6} \times \frac{V_g^{3/2}}{a^2} \frac{\text{amp.}}{\text{cm}^2}$. Reference to Figure 11 tells us that the anode current will increase directly with space current until the space-current density reaches a value

$I_0 = 7.00 \times 10^{-6} \frac{V_g^{3/2}}{a^2} \frac{\text{amp.}}{\text{cm}^2}$, corresponding to a value of $P_a^2 = 3.00$. This is the maximum anode-current density which can be passed: at this point the potential minimum, which has decreased to a value 0.0956 times the grid voltage, drops abruptly to zero and becomes a virtual cathode. The potential minimum, which had initially been formed at the anode, gradually recedes toward the grid until it reaches a distance equal to 0.776 times the grid-anode distance, given by equation (14). At this point it becomes a virtual cathode, and enables some of the electrons to return toward the grid. The resulting re-distribution of space charge causes the virtual cathode to shift abruptly until it reaches a distance equal to 0.418 times the grid-anode distance, given by equation (18). Any further increase of I_0 results in a decreased anode current, as shown in Figure 11, the virtual cathode retreating still further toward the grid. If the space current is then decreased, the anode current will increase, as shown in Figure 11, and the virtual cathode will move back toward the anode. This behavior will continue until the space-current density reaches a value

$I_0 = 3.38 \times 10^{-6} \times \frac{V_g^{3/2}}{a^2} \frac{\text{amp.}}{\text{cm}^2}$, corresponding to a value of $P_a^2 = 1.45$. At this point the virtual cathode is at a distance equal to 0.719 times the grid-anode distance, as given by equation (18). The slightest further reduction of space current permits enough electrons to pass abruptly through the virtual-cathode barrier toward the anode so that the space-charge distribution is radically altered, the virtual cathode disappears, and the anode current rises abruptly to the full value of space current. (An interesting observation about this example is that at this unstable point the virtual cathode does not become a potential minimum, but disappears.) From this point on, any further reduction of space current results in a proportionate reduction of anode current, as shown in Figure 11, and the potential distribution becomes more and more linear.

3. Effects of Varying V_a, when $P_a = 1.0$.

Let us now suppose that the grid-anode distance, a, the effective grid voltage, V_g, and the space-current density, I_o, are fixed, and that it is required to determine the effects of varying the anode voltage, V_a. This situation corresponds, for example, to the case of a tetrode, the screen-grid voltage and the control-grid voltage of which are fixed, and the anode voltage of which is varied.

To begin with, we assume that the control-grid voltage is set at a value which provides a space-current density corresponding to a value of $P_a = 1.0$. When the anode voltage is negative, the anode current is zero, and all of the electrons are returned toward the grid at the temperature-limited virtual cathode. The potential distribution for $\frac{V_a}{V_g} = -0.2$ is shown by (a) of Figure 14. The position of the virtual cathode is given by Figure 8 as $c = 0.645\ a$. When $V_a = 0$, then $I_a = 0$, and $c = 0.707\ a$, as shown by (b) of Figure 14. As V_a is made more positive, the anode current rises as shown in Figure 10 and the space-charge-limited virtual cathode moves toward the anode. (If there had been no movement of the virtual cathode, the anode current would rise as $V_a^{3/2}$.) This continues until $V_a = 0.068\ V_g$, while the anode current density rises to $I_a = 0.413\ I_o$, and the virtual cathode moves to $c = 0.793\ a$, as shown by (c) of Figure 14. The slightest further increase of anode voltage results in an abrupt saturation of anode current, the rate of change of anode current with anode voltage being infinite at this point, as shown by Figure 10. The virtual cathode disappears at this point, and the potential distribution is of the form shown by (c') of Figure 14, characteristic of the mode for which neither potential minimum nor virtual cathode exist. Further increase of anode voltage leaves the anode current, which is now equal to the space current, unaffected. However, when V_a reaches $0.75\ V_g$, a potential minimum appears at the anode, as shown by (d) of Figure 14. With increasing anode voltage this minimum recedes toward the grid, approaches V_g in value, and finally disappears, as shown in Figures 7 and 8.

If the anode voltage is now decreased, the anode current remains at saturation value until $V_a = 0$, as shown in Figure 10. The potential minimum re-appears at the grid when $V_a = 1.23\ V_g$, moves toward the anode as V_a is decreased, and reaches the anode when $V_a = 0.75\ V_g$, as before. At this point the minimum disappears. When $V_a = 0$, the potential distribution is abruptly altered, and a virtual cathode is formed at a distance $c = 0.707\ a$, as shown by (f), (f') of Figure 14.

The time of flight of an electron moving from grid to anode, and from grid to virtual cathode, is shown for increasing and decreasing anode voltages in Figure 6.

4. Effects of Varying V_a, when $P_a = \sqrt{2}$, $P_a = 2\sqrt{2}$.

Similar interpretations hold for any other value of the parameter P_a. As a matter of interest, the various forms of the potential distribution for the particular cases $P_a = \sqrt{2}$ and $P_a = 2\sqrt{2}$ are shown in Figures 15 and 16. For the first of these, as the anode voltage is increased from a negative value, the potential distribution passes from the state described by Case 4, which is characterized by a temperature-limited virtual cathode, (a), to that described by Case 3—the space-charge-limited virtual cathode case, (b), (c); thence abruptly to that described by Case 2, for which a potential minimum is formed, (c'), (d); and then finally to that described by Case 1, for which neither potential minimum nor virtual cathode exist. The behavior for decreasing anode voltage is successively illustrated by (d), (c'), (e), (f), (g), (g'), (a) of Figure 15. The abrupt decrease in anode current again occurs at a lower value of V_a than that required for the abrupt rise, as shown in Figure 10.

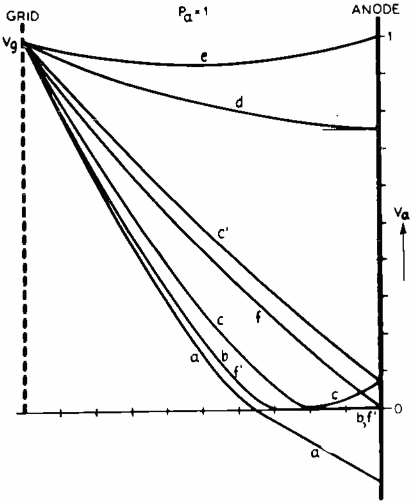

Fig. 14—Potential distribution for various values of anode voltage. $P_a = 1$.

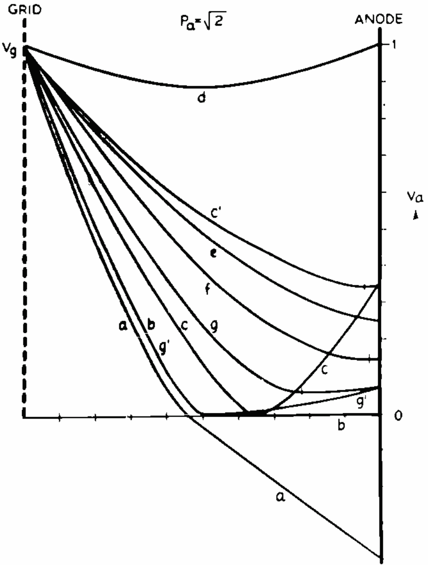

Fig. 15—Potential distribution for various values of anode voltage. $P_a = \sqrt{2}$.

B—*General Discussion of Curves.*—The four distinct modes of potential distribution which are treated in the foregoing analysis represent a somewhat arbitrary and simplified division of the potential distributions which may occur in the grid-anode region. This simplification is made possible by the original assumptions that the electrons have all been initially emitted from the cathode with zero velocity and that no velocity distribution has been introduced during the passage of the electrons through the structure. In the rigorous treatment of this problem the separate analyses of the four modes of potential distribution would merge into a single general analysis. However, such a treatment would be considerably more involved, and the main results at least can probably be anticipated by combining the various results found here for the individual states of operation on common graphs. This has been done in Figures 5-11.

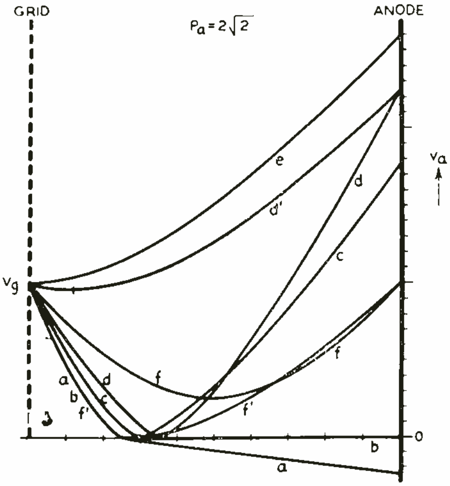

Fig. 16—Potential distribution for various values of anode voltage. $P_a = 2\sqrt{2}$.

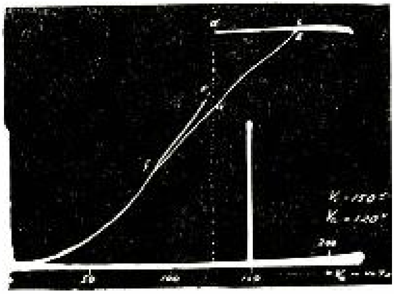

Fig. 17—Anode-characteristic oscillogram of special pentode showing the hysteresis and instability phenomena similar to curve marked $P_a = 2\sqrt{2}$ in Figure 10.

Figure 5 is a plot of P_a vs. $\dfrac{V_a}{V_g}$, for various values of Q_g. If I_o, V_a, V_g, and a are given, this plot enables one to determine the electric intensity at the grid, since P_a is proportional to $\dfrac{a I_o^{1/2}}{V_g^{3/4}}$ and Q_g is proportional to $\dfrac{E_g}{I_o^{1/2} V_g^{1/4}}$. As was pointed out in the analysis, these curves also represent a generalized potential distribution plot if P_a is replaced by P, which is proportional to $\frac{xI_0^{1/2}}{V_g^{3/4}}$, and $\frac{V_a}{V_g}$ is replaced by $\frac{V}{V_g}$. For the temperature-limited virtual cathode case (negative anode potentials) the scale of ordinates represents $\sqrt{2}\,P_a$ instead of P_a. This results from the fact that all the electrons which pass through the grid toward the anode are turned back toward the grid at the virtual cathode, and the space charge corresponds to a space current of $2\,I_0$ instead of I_0. The parameter Q_g is replaced by $\frac{I_a}{I_0}$ in the space-charge-limited virtual cathode case because this quantity is of greater practical significance. However, if the electric intensity at the grid is required for this case, it can be found very quickly from the relation

$$Q_g = -\left(2 - \frac{I_a}{I_0}\right)^{1/2} .$$

It will be observed that there occurs a curious overlapping of the various possible states of operation in different regions of the plot. For example, for $0 \leq \frac{V_a}{V_g} \leq 1$ and $\frac{1}{\sqrt{2}} \leq P_a \leq \sqrt{2}$, there may be a virtual cathode, a potential minimum, or neither potential minimum nor virtual cathode. Again, for $\frac{V_a}{V_g} \geq 0$ and $P_a \geq \sqrt{2}$, there may be a virtual cathode or a potential minimum. To put it in another way, the potential at any point in the region, for certain values of I_0, V_a, V_g and a, is multi-valued. This is a typical hysteresis phenomenon, and in order to determine which of the values of potential is the correct one it is necessary to know, as in all cases of hysteresis, the previous history of the region. This point will be discussed in further detail in connection with the other figures.

Figure 9 is a plot of $\frac{V_c}{V_g}$ vs. P_a^2, for various values of $\frac{V_a}{V_g}$. The curves shown in this plot indicate how the minimum potential varies with space current, since P_a^2 is proportional to I_0. The section of the plot which lies to the left of the lightly-dashed line connecting the points $\left(\frac{V_c}{V_g} = 1,\ P_a^2 = 0\right)$ and $\left(\frac{V_c}{V_g} = 0,\ P_a^2 = 1\right)$ delineates the operating region in which neither potential minimum nor virtual cathode occur. The lightly-dashed line which starts at the point $\left(\frac{V_c}{V_g} = 0,\ P_a^2 = 1\right)$ and passes through the point $\left(\frac{V_c}{V_g} = 0.25,\ P_a^2 = 8\right)$ is a parametric line which indicates the points at which the rate of change of the potential minimum with (increasing) space current is infinite. If the grid and anode potentials of a given tube are fixed and the space current is increased from zero, the potential minimum shifts abruptly at these points into a virtual cathode. The lightly-dashed line which starts at the point $\left(\frac{V_c}{V_g} = 0,\ P_a^2 = 1\right)$ and becomes asymptotic to $\frac{V_c}{V_g} = 1$ for very large values of P_a^2 is a parametric line which indicates the points at which the virtual cathode shifts abruptly, with decreasing space current, into a potential minimum. The heavily-dashed continuations of the individual curves represent mathematical solutions of equation (13) which are probably unrealizable, at least in the steady-state or low-frequency operation of the tube.

These curves exhibit the instabilities and hysteresis phenomena which are characteristic of the grid-anode region, and again illustrate the necessity for a knowledge of the previous history of the operation of the tube in order to predict its future behavior.

Figure 11 is a plot of $\left(\frac{I_a}{I_o}\right) \cdot P_a^2$ vs. P_a^2 for various values of $\frac{V_a}{V_g}$. These curves illustrate the variation of the anode current with space current for several particular values of the ratio of anode potential to grid potential. It will be observed that if the space current is increased from zero, the anode current bears a linear relation to the space current until a maximum value, given by equation (13a), is reached. At this point a virtual cathode is formed and the anode current is abruptly reduced. Thereafter, the anode current decreases continuously with increasing space current. If the space current is now decreased, the anode current will increase until the virtual cathode disappears. At this point the anode current becomes exactly equal to the space current, the phenomenon being an abrupt one for values of $\frac{V_a}{V_g} \leqq 1$. For $\frac{V_a}{V_g} \geqq 1$ this discontinuous phenomenon does not occur.

Further decrease of space current causes the anode current to decrease linearly with space current. The portions of these curves in which the anode current decreases with increasing space current are intimately associated with the presence of a virtual cathode, and can be utilized to provide a negative-transconductance amplifier or oscillator. This matter will be elaborated upon in section *D* of this discussion.

The dotted appendages of the individual curves in this graph represent mathematical solutions of equation (19) which are probably unrealizable in the steady-state or low-frequency operation of the tube. The terminations of these dotted sections upon the linear portion of unit slope, common to all of the curves, are given by equation (13b). On occasion these values have been assumed, incorrectly, as giving the maximum anode-current density which can be obtained.[7]

Figure 7 is a plot of $\frac{V_c}{V_g}$ vs. $\frac{V_a}{V_g}$ for various values of P_a. The curves shown in this plot illustrate the variation of the minimum potential with anode potential for several values of space current. The dotted line connecting the points $\left(\frac{V_c}{V_g}=0, \frac{V_a}{V_g}=0\right)$ and $\left(\frac{V_c}{V_g}=1, \frac{V_a}{V_g}=1\right)$ passes through the points on the individual curves at which a potential minimum is suddenly formed when the anode potential is increasing. The dotted line which becomes asymptotic to the value $\frac{V_c}{V_g}=1$ passes through the points on the individual curves at which the virtual cathode is abruptly transformed into a potential minimum when the anode potential is increasing. The third dotted line, which starts at the point $\left(\frac{V_c}{V_g}=0, \frac{V_a}{V_g}=0\right)$ and passes through the point $\left(\frac{V_c}{V_g}=0.25, \frac{V_a}{V_g}=1\right)$, is a parametric line which indicates the points on the individual curves at which the potential minimum shifts abruptly, with decreasing anode potential, into a virtual cathode.

The curve which represents the case $P_a = 0$ degenerates into a point located at $\left(\frac{V_c}{V_g}=1, \frac{V_a}{V_g}=1\right)$, since this is the only value of the ratio $\frac{V_a}{V_g}$ for which the electric intensity can be zero in the absence of space current. The curves which represent the case $P_a = 1$ consist of two discontinuous sections, the first being a portion of the axis of abscissas (this corresponding to the virtual cathode), and the second being the curved portion which begins on the first dotted line, which represents the potential minimum. For this case, when the anode potential is increased from negative values, a virtual cathode is first formed which presently disappears abruptly, and then for some greater value of anode potential, a potential minimum is formed. For values of $P_a \leqq \frac{1}{\sqrt{2}}$, neither potential minimum nor virtual cathode is formed with increasing anode voltage until $\frac{V_a}{V_g} \geqq 0.888$. Then a potential minimum is formed.

Figure 8 is a plot of $\frac{c}{a}$ vs. $\frac{V_a}{V_g}$ for various values of P_a. These curves show the position of the virtual cathode and potential minimum for several values of space current plotted against different anode potentials. The sections of the various curves which lie to the left of the axis of ordinates reveal the location of the temperature-limited virtual cathode. The sections of the same curves which lie to the right of the axis of ordinates indicate the location of the space-charge-limited virtual cathode and potential minimum. The dashed-line portions of the curves for $P_a = 1$ and $\sqrt{2}$ are mathematical solutions of equation (18) which are probably unrealizable in the ordinary operation of the tube. Figures 7 and 8 together give the location and magnitude of the potential minimum under various operating conditions. This information is of considerable value to a proper understanding of the operation of the tube.

Figure 10 is a plot of $\frac{I_a}{I_o}$ vs. $\frac{V_a}{V_g}$ for various values of P_a. These curves show the variation of the anode current with anode potential for several particular values of space current. It will be observed that as the anode potential is increased from zero, the anode current increases from zero relatively slowly at first and then more rapidly. This rapidity of increase of anode current is more pronounced for the smaller values of space current. The initial variation of anode current with anode potential is approximately a $\frac{3}{2}$-power law, such as obtains in the case of an idealized diode. As the anode potential increases, however, the returning current (corresponding to the electrons which return toward the grid from the virtual cathode) decreases because the anode current increases. This results in a movement of the virtual cathode toward the anode, and thus causes a further increase in the anode current. The net result is similar to that which would exist if the cathode of the idealized diode moved toward the anode with increasing anode potential, i.e., the current would increase more rapidly than predicted by the $\frac{3}{2}$-power law. It will be observed further that for the curves which represent the cases $P_a = 1$ and $\sqrt{2}$, the anode current saturates abruptly with increasing anode potential. This action indicates an instability in the plate characteristic for relatively low values of space current. For values of P_a greater than 2, the saturation of anode current occurs at a finite rate, i.e., the saturation is a stable phenomenon. If the anode potential is decreased toward zero, after the saturated condition has been attained, the anode current does not vary with anode potential entirely as before. For example, in the case $P_a = 2\sqrt{2}$, the anode current saturates with increasing anode potential at the value $\frac{V_a}{V_g} = 2.24$. With decreasing anode potential, the anode current remains saturated until the value $\frac{V_a}{V_g} = 1.0$ is reached, and thereafter follows the same law of variation as occurred for increasing anode potential. This "overhang" of anode current illustrates the hysteresis phenomena which occurs in the plate characteristic.

The dashed-line portions of the curves for $P_a = 1$ and $\sqrt{2}$, which indicate a negative-resistance anode characteristic, represent mathematical solutions of equation (19) which are probably unrealizable in the steady-state or low-frequency operation of the tube. The existence of such solutions has been used, unjustifiably, to form the basis of a negative-resistance explanation of Barkhausen-Kurz oscillations.[7] Such oscillations can be explained more satisfactorily by an analysis which takes into account, at the start, the effect of the transit time of the electrons.

Before leaving the discussion of Figure 10, it should be pointed out that curves representing values of P_a less than 1 have been omitted from this plot, since the analysis indicates the physically unrealizable phenomenon of a saturation current "overhang" into the negative region of $\frac{V_a}{V_g}$. The value $P_a = \frac{1}{\sqrt{2}}$ represents the limiting case for which, with increasing anode voltage, the anode current rises gradually. For values of $P_a \leq \frac{1}{\sqrt{2}}$, the anode current saturates abruptly at $\frac{V_a}{V_g} = 0$.

Figure 6 is a plot of $\frac{\tau}{\tau_0}$ and $\frac{\tau'}{\tau_0}$ vs. $\frac{V_a}{V_g}$ for various values of P_a. These curves show the variation with anode potential of the transit time, τ, for the electrons which travel from the plane of the grid to the anode, and of the transit time, τ', for the electrons which travel from the plane of the grid to the virtual cathode. The curves illustrate the magnitude of the effect on the transit time of the electrons which can occur in the presence of space charge.

IV. EXPERIMENTAL VERIFICATION

To obtain experimental verification, at least of a qualitative nature, of the main features of the theoretical results, a special pentode was constructed. The tube consisted of an indirectly-heated cathode having an area of 6.3 sq cm; a control grid, the potential of which was varied to alter the space current between the screen grid and anode; and a screen grid spaced 1.5 cm from a suppressor grid located in front of, and very close to an anode having an area of 9 sq cm. The suppressor grid was connected to the anode to prevent any secondary electrons, which might have been emitted from the anode, from seriously disturbing the space-charge conditions in the region between screen grid and anode. The electrodes were made slightly concave to minimize spreading of the electron stream, and large enough in area to minimize edge effects.

Oscillograms of the anode and transfer characteristics of this tube were taken on a special cathode-ray curve tracer built by Mr. O. H. Schade, of this laboratory.* The characteristics were determined in this way because the heavy currents would have damaged the tube if the tube performance had been observed by the usual point-by-point method.

Figure 17 is an oscillogram record of the variation of anode current with anode voltage, the control-grid voltage being 150 volts and the

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* This device was described by Mr. Schade before the Rochester Convention of the I.R.E. on November 20, 1935.

screen-grid voltage being 120 volts. It will be observed that the anode current increases uniformly, as the anode voltage is increased from zero, until the point marked "*c*" where saturation is reached, and thereafter changes very slowly with further increase of anode voltage. When the anode voltage is decreased, the anode current remains virtually constant until the point "*d*" is reached, at which point the anode voltage is considerably less than at the point "*c*". With further decrease of anode voltage, the anode current drops discontinuously and then decreases uniformly on the same curve as obtained with increasing anode voltage. This behavior is qualitatively similar to the theoretical curve marked $P_a = 2\sqrt{2}$ of Figure 10.

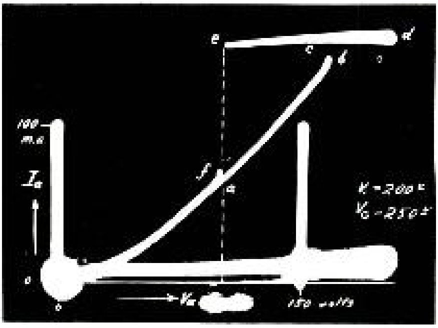

Fig. 18—Anode-characteristic oscillogram of special pentode showing the hysteresis and instabilities similar to curve marked $P_a = \sqrt{2}$ in Figure 10.

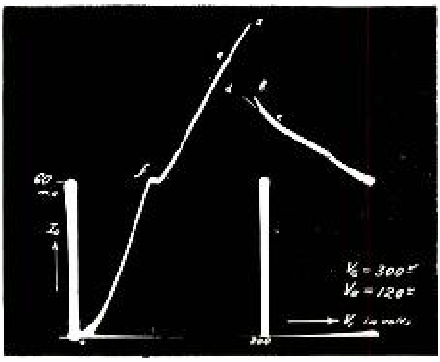

Fig. 19 — Transfer-characteristic oscillogram of special pentode showing the hysteresis and instabilities characteristic of the curve marked $\frac{V_a}{V_g} = 0.4$ of Figure 11.

Figure 18 is a similar oscillogram record, the control grid now being at 200 volts and the screen grid at 250 volts. This characteristic shows an abrupt increase of the anode current at the point "*b*" when the anode voltage is increasing, and an abrupt decrease of the anode current at the point "*e*" when the anode voltage is decreasing. This behavior is qualitatively similar to the theoretical curve marked $P_a = \sqrt{2}$ of Figure 10. The faint irregular traces which are present at the abrupt changes of anode current are probably mainly due to the unavoidable reactance of the tube connecting leads.

Figure 19 is an oscillogram record of the variation of anode current with control-grid voltage, to which the space current is proportional. The screen grid was maintained at 300 volts and the anode at 120 volts. This record is in qualitative agreement with the curve for $\frac{V_a}{V_g} = 0.4$ of Figure 11. It will be observed that the anode current increases uniformly, except at the point "*f*", with increasing space current until a virtual cathode is almost formed at the value corresponding to the point "*a*". Further increase of space current causes the formation of a virtual cathode between screen grid and anode, with an attendant sudden decrease of anode current, as indicated by the point "*b*". Still further increase of space current is accompanied by a uniform decrease of anode current. (The negative slope of the I_a vs. V_1 characteristic in this region represents a negative transconductance and can be utilized to form the basis of a novel type of amplifier and oscillator.) If the space current is now decreased, the anode current increases until the point "*d*" is reached, at which point an abrupt increase of anode current takes place. The point "*d*" corresponds to a lower value of space current than the point "*a*", in accordance with theoretical expectations, and marks the sudden disappearance of the virtual cathode and the formation of a potential minimum. Further decrease of space current is accompanied by a decrease of anode current, the same path now being followed as for increasing space current.

The curious kink in the early portion of the curve, indicated by the point "*f*", may be accounted for as follows: Since the control grid was operated at positive potentials in order to obtain sufficient space current for the experiment, it is to be expected that the possibility of primary and secondary emission from this grid is great. At relatively low values of control-grid voltage, corresponding to relatively low values of space current, these electrons contribute to the anode current. As the control-grid voltage is increased, the space current increases sufficiently so that the value of the potential minimum which has been formed in the region between screen grid and anode becomes less than that of the control grid. As a result the electrons emitted from the control grid are confronted by a potential barrier which they are unable to penetrate and consequently they execute excursions about the screen grid until captured. The slope of the I_a vs. V_1 curve is greater for values of V_1 corresponding to the range between zero and the point "*f*" than for the range above "*f*". This is accounted for by the explanation given above, because for values of V_1 less than those corresponding to "*f*" the electrons which are emitted from the control grid contribute to the anode current. Beyond this point these electrons do not contribute to the anode current, but instead because of their multiple excursions about the screen grid, they increase the value of space current and actually cause the anode current to increase at a lower rate with control-grid voltage than before. Approximate calculations which have been made support this explanation, but their reproduction at this point would be out of place.

V. ILLUSTRATIVE APPLICATIONS

The theoretical results which have been presented in the foregoing treatment are applicable to a wide variety of vacuum-tube problems. For purposes of illustration, we will now discuss two such problems, both of considerable practical interest.

1—*The Type 6L6 Beam Power Tube.*—This tube is a quasi-pentode power amplifier which employs a minimum potential, deliberately formed in front of the anode by utilizing the space charge of the electrons, to minimize the passage of secondary electrons from anode to screen grid. The existence, under certain conditions, of a potential minimum or virtual cathode in the region between screen grid and anode has been known for some time,[6,7,8] and the applicability of the phenomenon to the minimization of the (undesired) passage of secondary electrons from the anode to the grid has also been recognized for some time.[1,5,6,7] Recently, however, the subject has received considerable attention,[2,3,4,10,11] and it appears to us that a number of misconceptions which have arisen concerning the theory of operation of such a tube can be cleared up by means of the foregoing theory with certain modifications, of a quantitative nature, introduced by the factors which we explicitly neglected at the start.

The objectives which were sought in the development of this tube were improvements over a-f power output tubes then available with regard to power output, efficiency, power sensitivity, and distortion. These objectives could best be attained by a high-transconductance tube having an anode characteristic as close as possible to that of an ideal pentode.[4] Although the usual pentode anode characteristic exhibits no trace of the secondary-emission phenomena typical of the conventional tetrode, it is invariably marked by a relatively slow initial rise of anode current with anode voltage and by a rather gradual saturation of the anode current. These undesirable features can be charged mainly to the presence of the suppressor grid, and therefore it was decided to dispense with this grid and replace it, at least as far as its effect on preventing the flow of secondary electrons from the anode to the screen grid was concerned, by a potential minimum. The ultimate result of this development was the 6L6, the anode characteristic of which is reproduced in Figure 20.

In general, the main features of the curves shown in this figure agree qualitatively with those of the idealized curves shown in Figure

10. For example, the curves corresponding to the largest values of space current, i.e., highest values of control-grid voltage, tend to saturate at higher values of anode voltage and also exhibit the abrupt type of saturation predicted by the theory for values of P_a less than 2. The smooth type of saturation shown in Figure 10 for larger values of P_a, which was verified experimentally in Figure 17, is also evident in Figure 20. However, there are several discrepancies between the theoretical results for the idealized parallel-plane tube and the curves of Figure 20. One of these lies in the fact that none of the curves saturate immediately, whereas the theory indicates that for P_a less than $\frac{1}{\sqrt{2}}$ the anode current should rise to saturation value for the smallest positive anode potential. This lack of agreement can be attributed in small part to the Maxwellian distribution of velocities of electron emission, and in much greater part to the angular deflection of the electrons which occurs because of the wire grid structure.[9] This effect is most pronounced at the very lowest values of space current: at the higher space currents, where a virtual cathode is formed, this effect is entirely overshadowed by the behavior predicted by the theory. Another discrepancy is to be observed in the curve for $E_{c1} = -30$ volts, which exhibits negative anode conductance for anode voltages between 25 and 75 volts. This can doubtless be explained by the existence of a potential minimum, the value of which is too high to provide a suitable retarding field for the proper suppression of secondary electrons which are emitted from the anode. A further discrepancy is to be observed in the fact that all of the curves have a finite slope in the saturation region. This is probably due in main part, to the penetration of the anode field through the screen grid, and perhaps also to the slight contribution to the anode current of primary and secondary electrons from the screen grid. A more serious discrepancy, from an academic point of view, is the relatively narrow area of the hysteresis loops which are to be observed in the curves for $E_{c1} = +10$ volts and $E_{c1} = +15$ volts. This may be due to the edge effects introduced by the zero potential "beam-forming" plates.

It may be well to point out here that the behavior of the initial portions of all of the theoretical curves of Figure 10 and also the corresponding portions of the experimental curves of Figures 17, 18, and 20, is at variance with that deduced by J. H. O. Harries,[3] shown in his Figure 9 as curve *G A D E*. Harries invokes the electron-velocity spread at the grid to explain the actual curvature of the rising portion of the anode characteristic, but it appears from the theoretical results that the curvature can be present even in the absence of a well-defined distribution of velocities of the electrons which pass through the grid. Nor is the suggestion of S. Rodda[11] that the work of Below and Schulze indicates that this curvature is due primarily to the angular deflection of electrons in the neighborhood of the screen grid valid, except (as we have mentioned above) for very low values of space current. As Harries pointed out,[11] this is because Below explicitly neglected the possibility of the formation of a virtual cathode.[9] The suggestion of Bell[10] that the "detailed mechanism" of the anode characteristic of tubes, ostensibly utilizing a potential minimum to minimize the passage of secondary electrons from the anode to the screen grid, may be due largely to the inherently steep initial rise of the tetrode plus the somewhat reduced secondary emission of electrons from the anode is interesting, but hardly satisfying as a complete theory. Finally, in passing, we wish to call attention to the possibilities of obtaining misleading results when using long glass tubes such as that of the "sliding-anode" tube employed by Harries.[2,3] This has been pointed out by I. Langmuir[12], in a discussion of Lilienfeld's work.

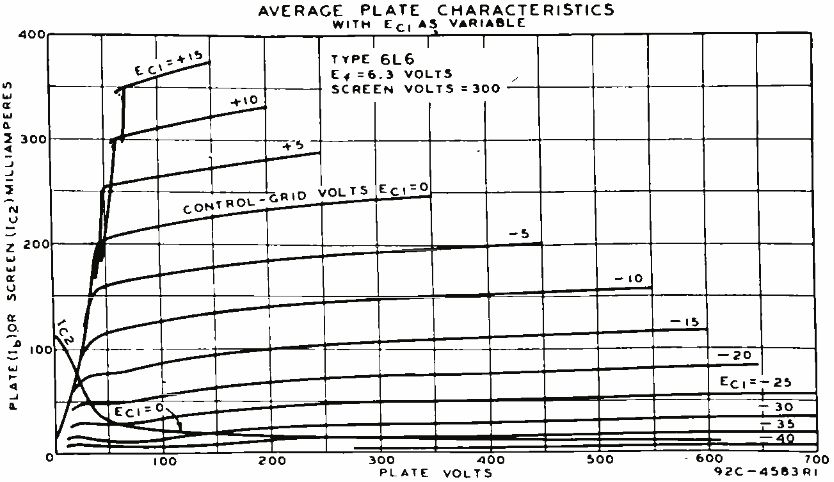

Fig. 20—Anode characteristic of RCA-6L6.

2—*Negative-Transconductance Amplifiers and Oscillators.*—To illustrate the application of the negative-transconductance features of Figure 11, in which the anode current decreases with increasing space current[6], a special pentode capable of operation with negative control voltage was constructed. The relative spacings of the electrodes were $K\text{-}G_1:\ G_1\text{-}G_2:\ G_2\text{-}G_3:\ G_3\text{-}A = 0.6:\ 0.6:\ 1.0:\ 8$. The first grid was used as a negative-control electrode, the second grid as a positive-accelerator electrode, the third grid as a positive electrode for controlling the velocity of the electrons entering the G_3-A region where a virtual cathode was formed, and the anode was used as the output electrode. The tube could be used as an amplifier and as an oscillator having several unique properties.

For an oscillator application, a typical circuit is shown in Figure 21. The control grid is coupled to the plate circuit by means of a large blocking condenser, C_{AG1}, the negative bias being supplied to the grid through a high resistance R_g. The grid excitation is adjusted by sliding the tap along the plate inductance. In this circuit the grid voltage is essentially in phase with the anode voltage, so that the tube oscillates only when the transconductance is negative. The condition for oscillation requires that the negative transconductance be greater than the sum of the internal anode conductance and the external circuit conductance, which includes the anode circuit proper and the reflected conductance of the grid resistor. Typical operating conditions were: $g_m = -2500$ μmhos, $g_p = 500$ μmhos, $g_c = 200$ μmhos, $V_a = 210$ volts, $I_a = 55$ ma, $V_3 = 200$ volts, $I_3 = 25$ ma, $V_2 = 200$ volts, $I_2 = 25$ ma, $V_1 = -12$ volts, $I_1 = -0.2$ milliamperes, $\lambda = 30$ meters, $V_{rf} = 15$ volts.

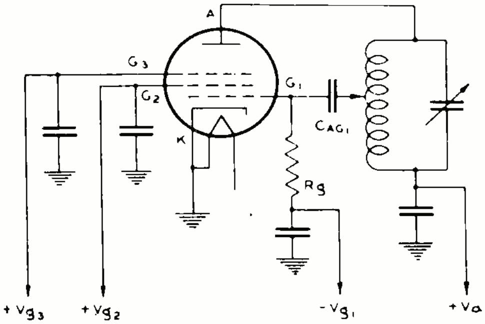

Fig. 21—Schematic circuit of negative-transconductance oscillator.

In order to obtain a larger plate swing without destroying the virtual cathode, and also to obtain a larger effective negative conductance by reducing the anode conductance, an additional screen grid may be placed in front of the anode. In this way the space-charge conditions in the G_3-G_4 region can be adjusted independently of the output circuit impedance by control of the potential of the screen grids G_3 and G_4. A tube of this type with relative electrode spacings of 0.6:0.6:1.0:8.0:2.0 showed an anode conductance of only 40 μmhos in the negative-transconductance region.

VI. CONCLUSIONS

The effects of space charge in the region between grid and anode are of fundamental importance in determining the behavior of vacuum tubes, particularly when the anode is preceded by a grid operated at a positive potential. The effects of space charge between the grid and anode of a parallel-plane vacuum tube have been determined in this study from the results of a simple analysis. The assumptions which underlie this analysis necessarily introduce modifications of the theory, but as the experimental verification indicates, these do not invalidate the main results, and their effects can be taken into account in a qualitative way.

The principal effects of the space charge in the grid-anode region are: (a) to introduce departures from the linear potential-distribution characteristic of the electrostatic case; (b) to set an upper limit to the current which can be collected at the anode, the limiting current density being

$$(I_a)_{\text{max.}} = 2.33 \times 10^{-6} \times \frac{(V_g^{1/2} + V_a^{1/2})^3}{a^2} \text{ amp per cm}^2$$

where V_g and V_a are the effective grid and anode voltages, respectively, and a is the distance between grid and anode; (c) to introduce instabilities and hysteresis phenomena in the behavior of the tube; and (d) to increase the electron-transit time in this region.

For the four modes of potential distribution which have been treated, expressions have been derived for the distribution of potential and electric intensity throughout the grid-anode region; the time of flight of electrons from grid to anode, and (when a zero-potential plane is formed) also from grid to this plane of zero potential; and the location and magnitude of the minimum potential. A formula has also been derived for the dependence of the anode current on the space current, grid-anode distance, grid voltage and anode voltage. These expressions have been plotted, using dimensionless parameters, in Figures 5-11.

The experimental verification of these theoretical results indicate that the theory can be employed in a qualitative way at least, to predict and explain the behavior of a large variety of vacuum tubes. This has been done for the 6L6 "beam power" tube, and for amplifier and oscillator tubes which make use of the decrease of anode current with increasing space current which the theory predicts.

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