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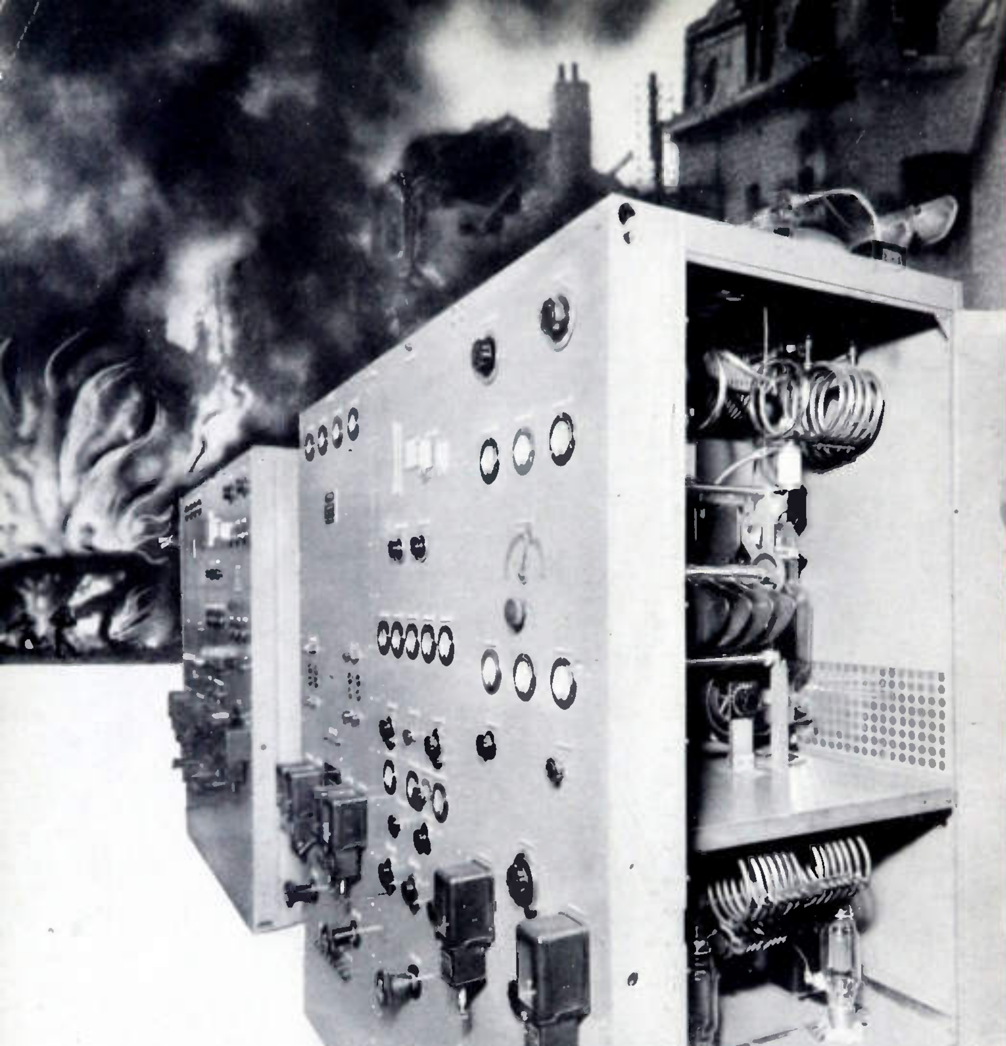

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Mobile Television Equipment*

R. L. CAMPBELL†, NONMEMBER, I.R.E., R. E. KESSLER†, NONMEMBER, I.R.E., R. E. RUTHERFORD†, NONMEMBER, I.R.E., AND K. U. LANDSBERG†, ASSOCIATE, I.R.E.

Summary—While portability is a necessary requirement for outside-pickup equipment, several advantages result when portability is carried into the studio. To equip a studio of adequate size with fixed equipment for the operation of several cameras involves considerable time and expenditure. However, with portable studio equipment, the entire equipment installation can be located to suit studio needs, as well as moved to different studios or outside locations.

The dolly-type equipment is described in some detail and systems for program control are discussed. Some of the design features discussed are portable and flexible synchronizing equipment, electronic view finders, oscilloscope monitors, and other operating facilities.

IN THE course of the development of television equipment, many of the improvements and simplifications resulting in better apparatus from the standpoint of performance and convenience of use are simply the applications of ideas developed in other allied fields which have been transferred to meet television design requirements. It may also be said that television-equipment design must follow, to some extent, the established precedents and engineering practices, e.g., radio broadcast equipment. When the precedent is followed too closely, however, difficulties are likely to appear in operating and maintenance because of the inherent complexity of the television system. In sound broadcast, there is one, and only one, electrical signal comprising the intelligence to be transmitted. In television, however, there are five separate electrical waves (and sometimes more depending on the type of system employed) which are combined and transmitted simultaneously to be used at the receiver in order to reproduce the picture. To make up this composite television-signal wave, several electrical waveforms not appearing in the final signal must be generated in order to obtain the television-system operation as we know it today. From this it can be seen that the operation of a television camera is by no means as simple an operation as setting up and operating a microphone for sound work.

With the above in mind, the purpose of this paper is to describe a type of television-camera equipment designed both for studio and outdoor use with respect to its function in a television operating plant. Particular components of the system to be described are mobile camera-control dolly, electronic view-finding system, flexible synchronizing equipment, sweep-driven control apparatus, interchangeability of units, cross control of camera dollies, and other operating features. Particular reference will be made to mechanical considerations as well as to some novel electrical features used in the equipment.

One application of this equipment would be for broadcast-studio operation. The economical factors involved in equipping a studio solely for television operation are likely to be out of proportion to the anticipated return on the investment in the case of most broadcast stations or other operating enterprises. Using the studio-type portable equipment, television programs can be presented with a minimum of installation difficulties. The cameras and camera-control equipment are merely rolled into the studio (together with adequate portable lighting fixtures) and the show is on. In the case of remote work, special events, etc., the same equipment can be wheeled into a small truck and unloaded and quickly set up for operation into a video line or relay channel.

A familiar and important requirement of portable equipment is weight. Considering the number of complex circuits involved in a television system, it can be seen that this problem is much more severe than in the case of equipment for remote sound work. Consideraable more apparatus is involved, and the question that immediately arises is, "Shall we have a few heavy units or shall we have several small, lightweight units?" In this equipment the latter was chosen for the following reasons:

1. The most logical electrical arrangement was to split the system into several units according to their functions.
2. Standard mechanical chassis arrangements could be adopted for ease of manufacture.
3. Servicing of all units was to be as convenient as possible.
4. No unit should be a two-man job to carry.
5. Future improvements can be added by replacing a unit at a time if desirable.
6. Television cameras using different types of pickup tubes may be used on the same equipment chains.

The camera and corresponding control equipment is arranged to operate in single or dual chains. In the case of a single chain, this equipment is divided into units as follows:

Synchronizing generator
Blanking sweep and power unit
Camera
Camera power supply
Electronic view finder
View-finder supply
Camera control
Shading generator (for use in conjunction with iconoscope cameras)
Camera monitor
Camera-monitor power supply
Camera-control power supply
Line amplifier

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† Allen B. DuMont Laboratories, Inc., Passaic, New Jersey.

Line-amplifier power supply
Line monitor
Line-monitor power supply

For a dual chain, the equipment required is:

1 Synchronizing generator
1 Blanking-sweep and power unit
2 Cameras

Fig. 1—Dual-camera-chain equipment.

2 Camera power supplies
2 Electronic view finders
2 View-finder power supplies
2 Camera controls
2 Shading generators (for use in conjunction with iconoscope cameras)
2 Camera monitors
2 Camera-monitor power supplies
2 Camera-control power supplies
1 Line amplifier
1 Line-amplifier power supply
1 Line monitor
1 Line-monitor power supply

In Fig. 1 is shown the apparatus outlined above arranged for dual-chain operation. On the camera-control dolly are the synchronizing generator, power units, camera-control units, monitors, and line equipment. With each camera connected to the main equipment dolly is the auxiliary camera power unit and the view-finding apparatus. This assembly is then connected back to the camera-control dolly by means of the camera cable, interlocked alternating-current power cable, and view-finder video cable.

For studio use the camera equipment proper is sometimes mounted on a studio platform dolly having a pedestal arranged to take the Akeley Gyro tripod head shown in the figure. The camera dolly platform supports the camera equipment and the cameraman, and it can be moved about the studio for camera "dolly" action shots.

Synchronizing Generator

The synchronizing generator used in this equipment is of the flexible fully electronic type and generates the DuMont synchronous wave (Fig. 2). The generator can be operated on any of the standards shown in

TABLE I

Lines per Frame	Fields per Second	Interlace
441*	60	2:1
525†	60	2:1
625*	30	2:1
343‡	120	2:1
441‡	120	2:1

* Experimental standards.
† Federal Communications Commission, (49851) "Television Report," May 3, 1941; also Donald G. Fink, National Television Systems Committee, Document No. 505L200M1.
‡ Color standards

Table I, and can be easily converted to other standards that may be desirable without affecting the standard chosen for regular operation. The synchronizing system may be switched to any one of these standards by means of a single wave switch and a few simple adjustments.

The complete generator is housed in two units, viz., the synchronizing-generator unit and the blanking-

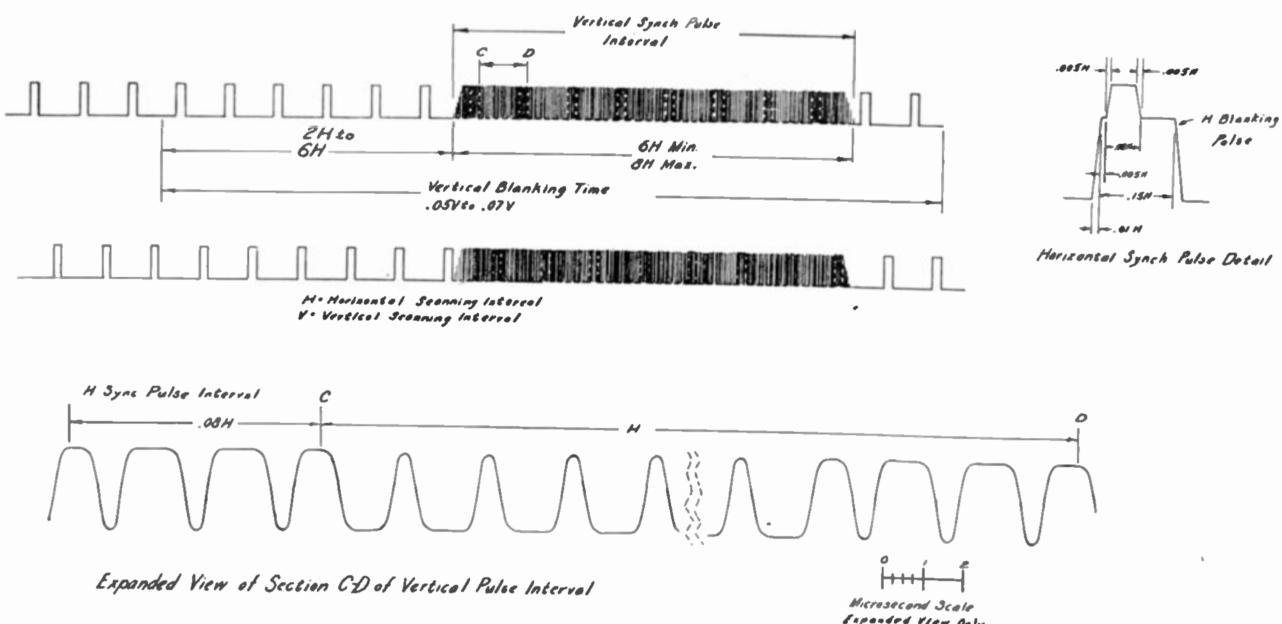

Fig. 2—DuMont synchronizing signal.

sweep and power unit. Fig. 3 shows the front panel of the synchronizing-generator unit with the cover removed. At the top of the unit is a monitor cathode-ray oscillograph which is connected to all circuits provided with front-panel adjustments. This cathode-ray oscillograph is of the "automatic" type; that is, the timing axis is automatically synchronized to the signal selected by the monitoring selector switch by means of an additional deck on the selector switch. Because of the many complex circuits involved in a synchronizing generator, and because it is desirable during operation to check the performance of the entire instrument without shutting down or throwing it out of adjustment, this monitor cathode-ray oscillograph is considered essential.

Fig. 4 shows a block diagram of the synchronizing system employed in the equipment. The synchronizing

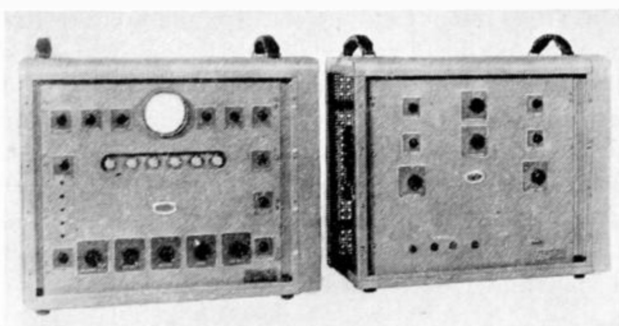

Fig. 3—Synchronizing generator with front cover removed.

generator can be divided into units according to the functions of the various circuits.

Unit No. 1 Monitor cathode-ray oscillograph
Frequency-divider circuit
Composite synchronizing-wave generator
Unit No. 2 Composite blanking
Master sweep generator
Power supply

The monitor cathode-ray oscillograph has been described above. The frequency-divider unit consists of transformer-coupled relaxation oscillators arranged to divide in accordance with the line- and frame-scanning standards selected. The switch to different standards is accomplished by means of a multiple-deck wave switch, connected to the oscillator and associated circuits, whereby the optimum circuit constants are selected for operation on the scanning standard chosen.

The composite synchronizing signal-generator circuit develops the synchronizing wave as shown in Fig. 2. Use of this type of signal makes it possible to minimize operating difficulties in the field insofar as synchronizing-generator performance is concerned. This is principally due to the fact that the composite synchronizing signal consists of two signals that are relatively simple to generate. Furthermore, improved vertical synchronizing performance is attained at the receiver.[1]

[1] National Television Systems Committee, Document No. 325R-200D31.

In the composite synchronizing-signal generator is the shaping circuit for horizontal pulses, the high-frequency carrier pulse generator for the field pulses, and the mixing and output circuits.

The blanking, sweep, and power unit contains those circuits indicated in its title. Power for all circuits in the generator is supplied from this unit by means of a well-filtered, regulated supply. From the generator

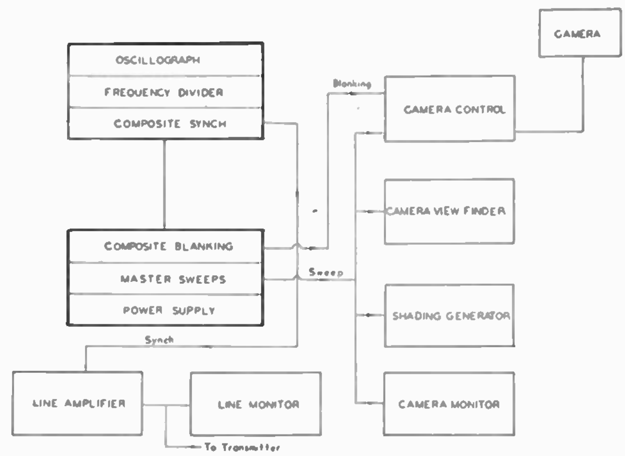

Fig. 4—Diagram of scanning and synchronizing system.

unit, driving pulses are fed to the sweep generators which control the scanning circuits on the cameras, monitors, and shading generators.

Horizontal and vertical blanking voltages are derived from the respective sweep-signal generators, and shaped in the blanking-generator circuit. They are

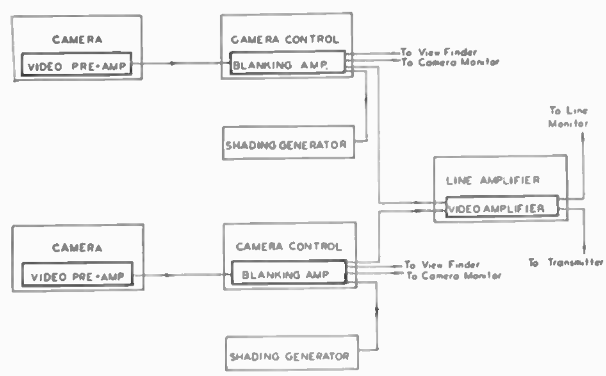

Fig. 5—Diagram of video system.

next mixed to form a composite blanking wave which is fed to the camera-control unit.

Low-impedance outputs are provided on the synchronizing-generator unit to feed a single or dual camera chain with the following signals: 1. horizontal sweep, 2. vertical sweep, 3. composite blanking, and 4. composite synchronizing. By means of the synchronizing distribution unit, several camera chains may be controlled from one generator if it is considered so desirable. For normal operation on a dual chain, and with reasonable cable lengths, the synchronizing distribution unit can be eliminated.

Video System

Fig. 5 diagrammatically shows the video system employed in a dual chain. Referring to this diagram, the video signal generated in the iconoscope output re-

Fig. 6—Camera equipment (iconoscope).

sistor is fed to the pre-amplifier in the camera where correction for capacitance of the iconoscope output circuit is accomplished by means of a peaking stage in this amplifier. A cathode-follower output stage on the pre-amplifier feeds through the main cable to the camera-control amplifier, which will be described later.

Camera

Fig. 6 shows the camera equipment. In the camera are the video pre-amplifier (Fig. 7), camera sweep circuits, a type 1850 iconoscope, camera blanking circuits, and protective circuits. Power for these circuits is fed from a separate cable from the camera power unit. The amount of power dissipated in the camera itself is such that the heat generated by the tubes would be excessive, especially when used in a "hot" studio or out in the sun. Therefore, it has been found desirable to isolate those tubes generating most of the heat and place them on a deck on the exterior of the camera. The lens mechanism is operated by means of a handle at the side, and provisions are made for interchanging lenses in the approximate range of 6½ inches, $f/2.5$ to 16 inches, $f/3.5$.

Camera Control

The circuits in the camera control unit are the video blanking amplifier, camera horizontal sweep-control and keystoning circuit, camera vertical sweep-control circuit, pedestal control, iconoscope beam control, iconoscope rim-light control, and the monitor and view-finder video supply circuits. The camera cable terminates in the rear of this unit, and all signals feeding the camera pass through the camera-control unit. (Note: The video signal to the view finder is fed over a separate small coaxial cable.) A test circuit for checking the plate currents of amplifier tubes in the camera control is connected by means of a switch to a meter on the front panel. The camera video amplifier is comprised of five stages and two blanking clippers.

Of interest in the camera-control unit are the blanking circuit and the pedestal-control circuit. The former utilizes a low-impedance diode limiter for clipping the blanking pedestal after mixing, and beyond this point in the amplifier is the pedestal control which is a similar diode circuit, but has a variable direct-current bias control for adjusting the amplitude of the pedestal in accordance with lighting conditions.

The video output circuit of the camera control consists of a high-level cathode-loaded stage which feeds the line amplifier and a low-level cathode-loaded stage for feeding the monitor, view finder, and shading-generator cathode-ray oscillograph. Fig. 8 shows the interior view of the camera control on the wiring side. Power for the camera-control unit is obtained from a separate regulated supply to which the camera power and view-finder power units are interlocked.

View Finder

In motion-picture production, probably the most important technician is the cameraman. His successes or failures are very probably due to his ability to visualize how the particular scene will appear when projected on the screen before the shot is taken. By means of the

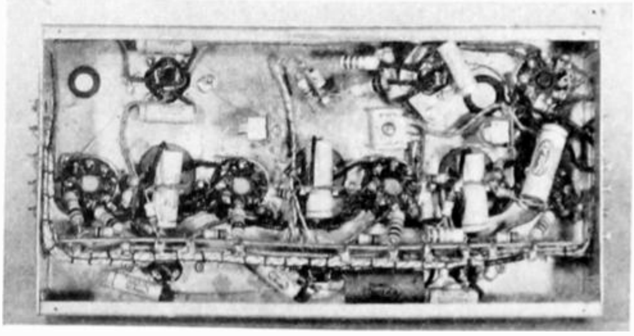

Fig. 7—Camera pre-amplifier.

electronic view finder, the television cameraman has an instantaneously developed picture before him at all times. View finding by means of matched lenses is an alternative method by which the cameraman can monitor his work. This method is expensive, however, and does not lend itself readily to quick interchangeability of lenses, sometimes required during a program. For these reasons the electronic method of view finding was chosen. Besides being able to determine the pictorial value of the scene before the camera, the electronic view finder is used as the focusing monitor. Thus, the cameraman can adjust the optical focusing instantaneously, and since he is in control of the camera, he can anticipate to some extent the position of the focusing handle and thus maintain the optical focus at all times. As an auxiliary to the electronic view finder, a framing device of some variety or other, or a Mitchell finder, is sometimes attached to the camera for the purpose of providing finding facilities outside of the field taken in by the camera.

The electrical arrangement of the view finder is as follows: A high-intensity 5-inch electrostatic-type cathode-ray tube is sweep-driven from signals to the

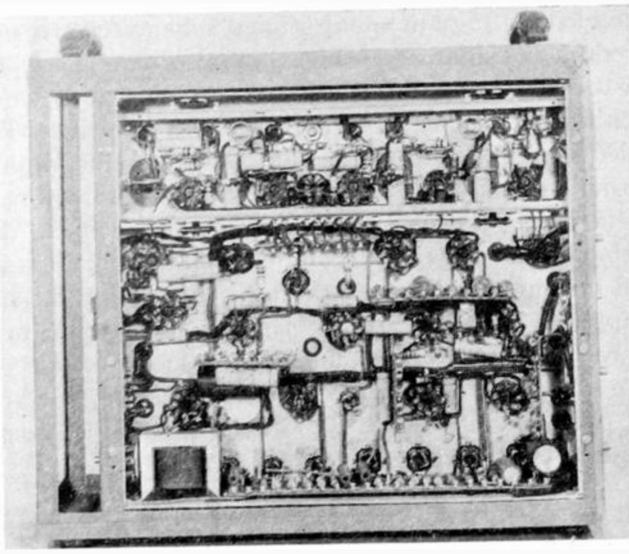

Fig. 8—Camera-control unit, wiring side.

camera. The sweep voltages are applied to plates of the cathode-ray tube by means of amplifiers located within the view-finder unit. The video signals fed to this unit are tapped off a monitor line in the camera control and fed to a video amplifier in the view-finder unit. Power and control circuits located in the view-finder supply unit are fed to the view finder by means of an interconnecting cable. (Controls are provided on the view-finder unit for maintaining the adjustments of brightness, contrast, and electrical focus, similar to those employed in television receivers.) Figs. 9 and 10 show the internal arrangements of the view finder and view-finder supply units, respectively.

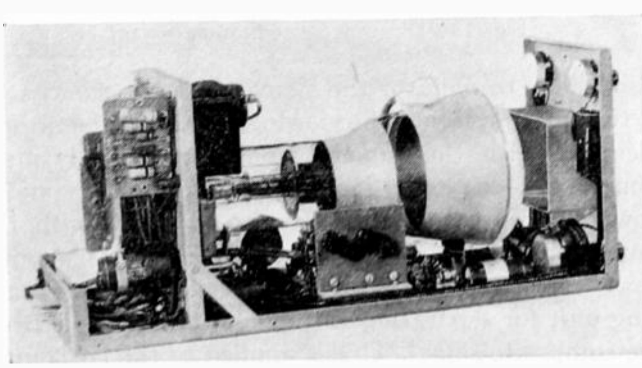

Fig. 9—View finder, interior view.

Shading Generator

The shading-control generator is a separate unit in the equipment and is used only in conjunction with iconoscope cameras. The shading signals are derived

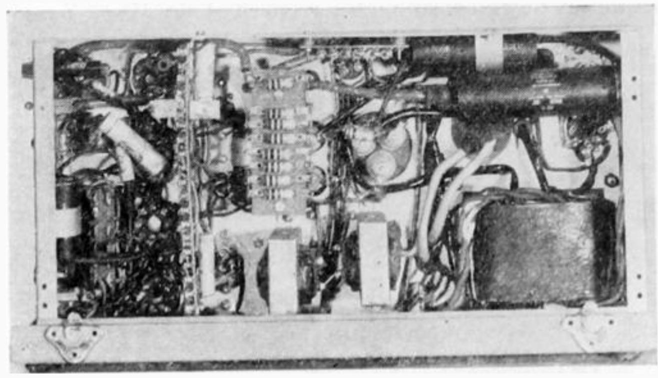

Fig. 10—View-finder power supply, interior view.

from the horizontal and vertical master sweep signals from the synchronizing generator. From these sweep signals the shading voltages generated in this unit are the horizontal saw-tooth, horizontal parabola, horizontal sine, vertical saw-tooth, vertical parabola, and vertical sine. These signals can be controlled both in

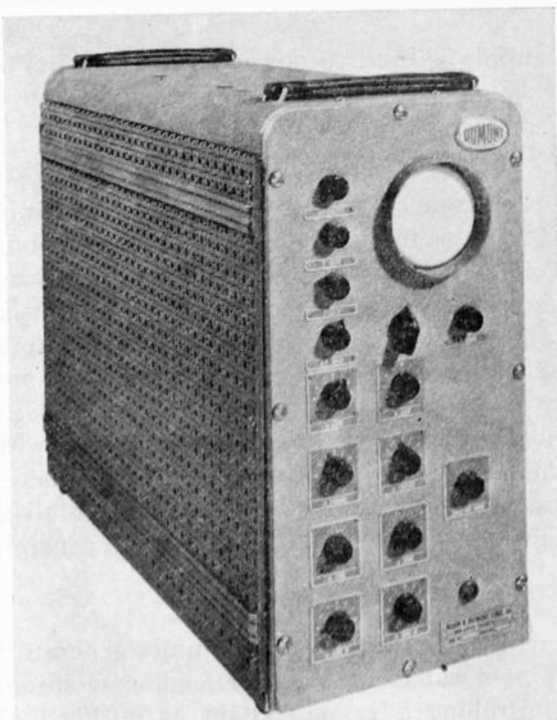

Fig. 11—Shading generator.

amplitude and phase so that many varieties of composite shading voltages can be obtained. These signals are mixed in a common amplifier whose output is fed into the iconoscope output circuit by means of a line in the camera cable. In the shading generator are the shading-generation, mixing, and output circuits, shading cathode-ray oscillograph, and the internal power unit. Video from the corresponding camera control is fed to the shading-generator cathode-ray oscillograph in order to monitor the shading signals. The time axis is driven from either the horizontal or vertical sweep depending upon the setting of a switch on the front panel. Thus, the operator selects the line-frequency sweep for checking horizontal shading, and the field-frequency sweep for checking vertical shading. A regulated supply is used to power all of the units in this circuit. Fig. 11 shows the shading generator.

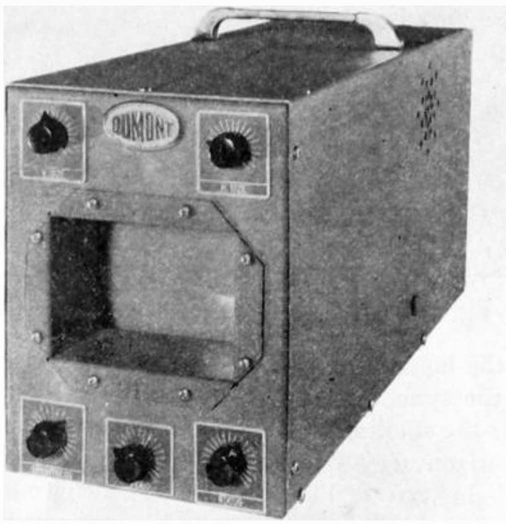

Fig. 12—Camera-monitor unit.

Camera Monitor

On each camera chain there is a monitor unit connected by cable to the camera control corresponding to the camera being operated. This monitor is usually placed directly on top of the camera control or shading generator for the convenience of the operator. The camera monitor is powered from the camera-monitor supply by means of an interconnecting cable. Since electrically the camera monitor is identical with the view finder, it need not be described further here. Fig. 12 shows a camera-monitor unit while Fig. 13 diagrammatically shows the monitoring system in general.

Line Amplifier

Normally, the camera-control units generate video signals at a sufficient level for feeding monitor lines and controlling a camera chain as outlined above. However, the signals from the two cameras must be selected or mixed as the case may be and then mixed with the synchronizing signal to form the composite television signal. This is accomplished in the line amplifier which contains the video switching unit, synchronizing mixing amplifier, main output stage, four auxiliary output stages, and the monitor cathode-ray oscillograph.

Push-button switching of cameras is accomplished in the switching unit by selecting one or the other of the camera-control video signals. In addition to the two video signals, the composite synchronizing signal from the synchronizing generator is fed to the line amplifier. Just before the output stage, a mixing circuit is provided to introduce the synchronizing signal with the video. A synchronizing gain control is provided for maintaining the proper percentage of synchronizing signal to video. In the event that a separate synchronizing transmission is used, this signal is fed directly from the generator to the transmitter or relay apparatus.

The main output stage of the line amplifier is a heavy-duty cathode-follower stage which normally feeds a 75-ohm line at an approximate level of 6 volts. In addition to this stage, three low-level stages are provided for 75-ohm monitor lines, such as for program directing, auxiliaries, and local monitoring. The monitor cathode-ray oscillograph is for the purpose of monitoring the output signal on the various lines. The video signal applied to the cathode-ray oscillograph is normally connected to the main output line. However, by means of a plug-in arrangement at the back, this cathode-ray oscillograph can be used to check all input and output terminals on the unit. Power for the line amplifier unit (excepting cathode-ray oscillograph power which is a built-in unit) is obtained from a separate supply which is identical to that used for powering the camera-control unit. Fig. 14 shows the tube side of the line amplifier.

Line Monitor

The line monitor unit is used for checking the signal selected by the switching unit (Fig. 13). In addition to monitoring the video signal fed to the line, this unit also serves to monitor the synchronizing performance of the entire system. The viewing unit of the monitor is identical with the camera monitor previously mentioned, with the exception of the driven sweeps, and is powered from a supply unit also identical with the camera-monitor supply. In addition to this supply, however, there is a synchronizing separator and scanning unit for separating the synchronizing wave from the composite signal. This is applied to the horizontal and vertical sweep oscillators of this monitor in the

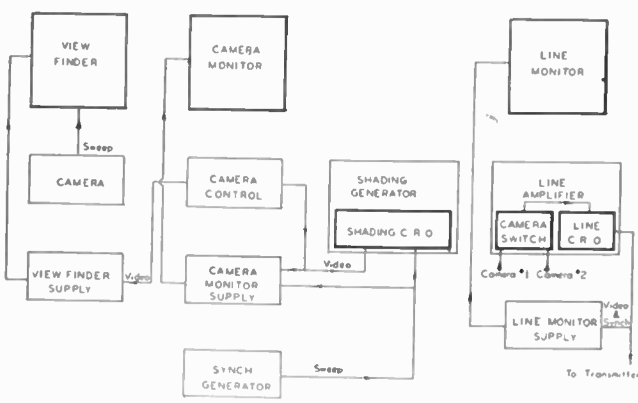

Fig. 13—Diagram of monitoring system.

same manner as in typical home television receivers. This line monitor, while intended primarily for operation with the DuMont synchronizing signal, is arranged to operate on synchronizing signals having rectangular field pulses as well as those of the radio-frequency type.

Control Dolly and Operation

The camera-control dolly is a lightweight frame on 10-inch pneumatic wheels occupying a floor space of 64×28½ inches for a dual chain. The height of the control desk is 30 inches and the operating desk slides into the unit when not in use. Single or dual equipment is controlled from the camera-control dolly by the camera-control operator. He has control over the electrical performance of the video system, including the synchronizing generator. Each camera is operated by a cameraman who, with the aid of the electronic view

TABLE II

Statistics, Mobile Camera Equipment

Unit	Weight Pounds	Case Size Inches	Power Watts	No. Tubes
Synchronizing generator	34	8×20×16	—	25
Blanking-sweep power unit	43	8×20×16	280	22
Camera	45	10×20×17½	—	13
Camera power supply	55*	8½×17½×15	225	19
Electronic view finder	14½	6¼×18×8	—	7
View-finder power supply	42*	8¼×17¼×10	230	8
Camera control	28	8×20×16	—	19
Shading generator	43	8×20×16	180	20
Camera monitor	12½	6¼×18×8	—	7
Camera-monitor power supply	42*	8½×17½×10	205	8
Camera-control power supply	49*	8½×17½×10	345	11
Line amplifier	31	8×20×16	—	16
Line-amplifier power supply	49*	8½×17½×10	325	11
Line monitor	12½	6¼×18×8	—	7
Line-monitor power supply	59*	8½×17½×16	230	16
	559.5		2025	209
Equipment dolly	115			
Total weight	674.5			

* Partly steel construction due to unavailability of lightweight metals.

finder, follows the action, maintains the focus, and is in general control of the pictorial value of the subject matter being picked up by his camera. There are provisions for interphone connections by which this operator is in communication with the two cameramen and is also in communication with the terminal point to which the video signal is being supplied. A sound-program-control unit is sometimes mounted on the camera-control dolly. When sound facilities are controlled here, some of the duties of the video-control operator can be taken over by the sound man.

Acknowledgment

The authors of this paper wish to acknowledge the capable direction of Mr. Allen B. DuMont resulting in the speedy accomplishment of this project; the many helpful suggestions of Mr. Paul Raibourn of Paramount Pictures, particularly regarding the camera-control dolly; the helpful suggestions and criticisms of

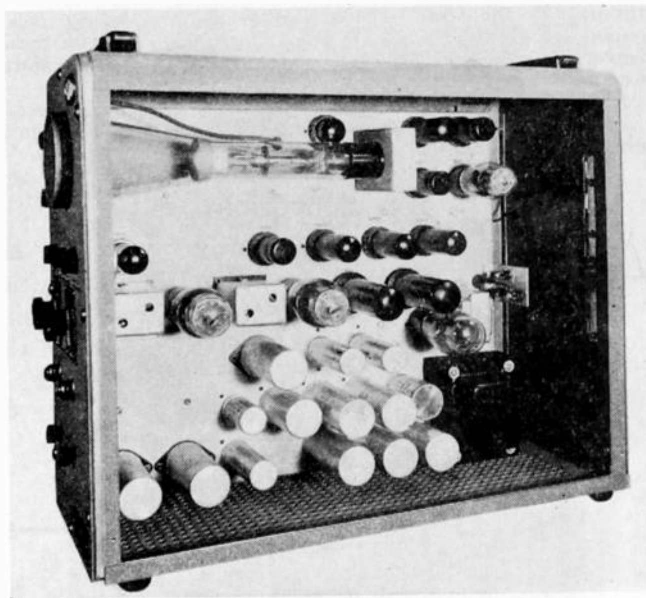

Fig. 14—Line amplifier, tube side.

Lieut. William C. Eddy and staff of Balaban and Katz Management, Chicago; and the valuable work of Messrs. M. A. Sanders, J. Winter, and others of the Allen B. DuMont Laboratories. The camera-control unit as described above was designed by Mr. W. H. Sayer.

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A Simple Television Demonstration System*

JESSE B. SHERMAN†, ASSOCIATE, I.R.E.

Summary—*The circuits and apparatus of a simple 150-line television demonstration system are described. The equipment employs the small type 1847 iconoscope as camera tube, remotely operated from the transmitter control panel. The transmitter is connected by cable to receiving equipment of which the principal item is a modified conventional oscilloscope. Some photographs taken of pictures on the kinescope screen are shown.*

It is shown that the low-frequency characteristic of a video cable driven from an amplifier plate circuit differs considerably from the characteristic of the usual interstage coupling.

I. Introduction

A SMALL iconoscope, recently available, has made it possible to develop medium-definition electronic television systems of relative simplicity. A previous paper has discussed the use of this iconoscope in amateur communication.[1] It is the purpose of the present paper to describe an application to apparatus for general demonstration and school-laboratory use, the total cost of which is less than the cost of a commercial-type iconoscope alone.

II. General

The transmitting end of the equipment consists of two units, the camera and the control cabinet. The camera contains the type 1847 iconoscope, lens, video pre-amplifier, scanning amplifiers, and remote-control focusing motor. The control cabinet contains the remainder of the video amplifier, the power supplies, scanning and pulse generators and amplifiers, monitor kinescope, and monitor oscilloscope.

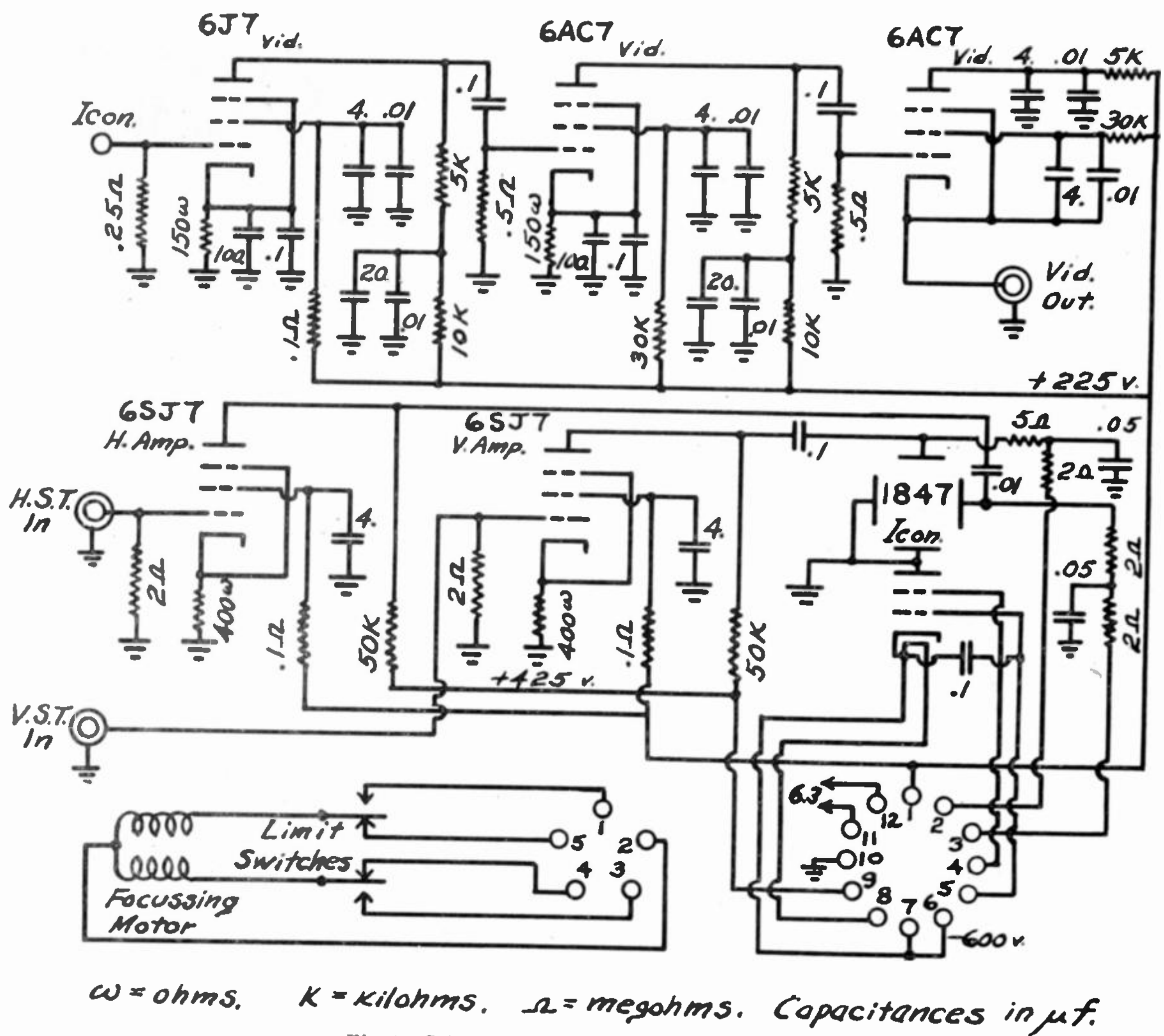

Fig. 1—Schematic wiring diagram of camera.

The receiving equipment consists of two units. The synchronizing pulse separator and amplifier, scanning oscillators, and power supply comprise one unit, while the other is made up of the kinescope and its associated equipment.

In the interest of simplicity and because the frequency band is of moderate width, interlacing is not

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[1] J. B. Sherman, "A new electronic television transmitting system for the amateur," *QST*, vol. 24, pp. 30–36; May, 1940.

used. The frame frequency is 60 cycles per second and the line frequency is 9 kilocycles, giving 150 lines. This was chosen as being at least the limit of the iconoscope definition. The frequency band produced, in the absence of iconoscope limitation, is approximately 600 kilocycles.

III. Camera

The camera is housed in a steel box $5 \times 6 \times 9$ inches with a separate enclosure at the rear for the focusing motor. As shown in Fig. 1, the video pre-amplifier contains three stages, the last cathode-coupled to the cable. The iconoscope is of the single-ended electrostatic-deflection type, the horizontal and vertical saw-tooth voltages being brought to the camera at a level of 5 to 10 volts and amplified in single-stage amplifiers.

The lens used has a focal length of 2 inches, $f/1.9$, and is of the 35-millimeter camera variety. This covers the usable area of the mosaic very adequately. In order to make it unnecessary for the camera to be within reach of the monitoring position, focusing is accomplished by means of a small reversible motor similar to

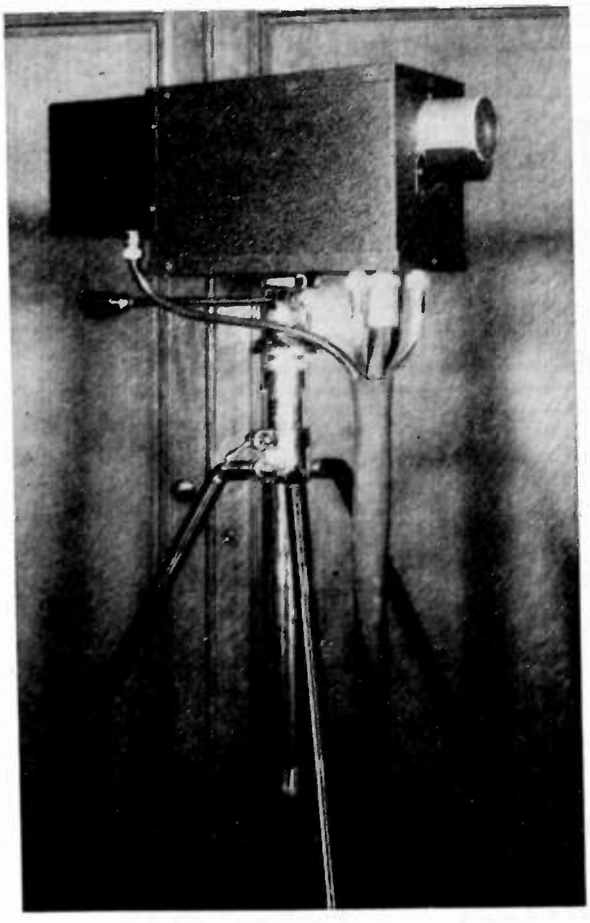


Fig. 2—External view of camera.

the kind commonly used for remote tuning of broadcast receivers. The rotor clutches and declutches automatically on starting and stopping, and the gearing is such that the shaft speed is about 60 revolutions per minute. The lens is driven on a screw having 32 threads per inch; thus it requires about a half minute to move the lens 1 inch. This speed is more than sufficiently slow for easy focusing; it is also such that only a very few seconds are usually consumed in changing focus. Limit switches are provided to prevent overtravel. The

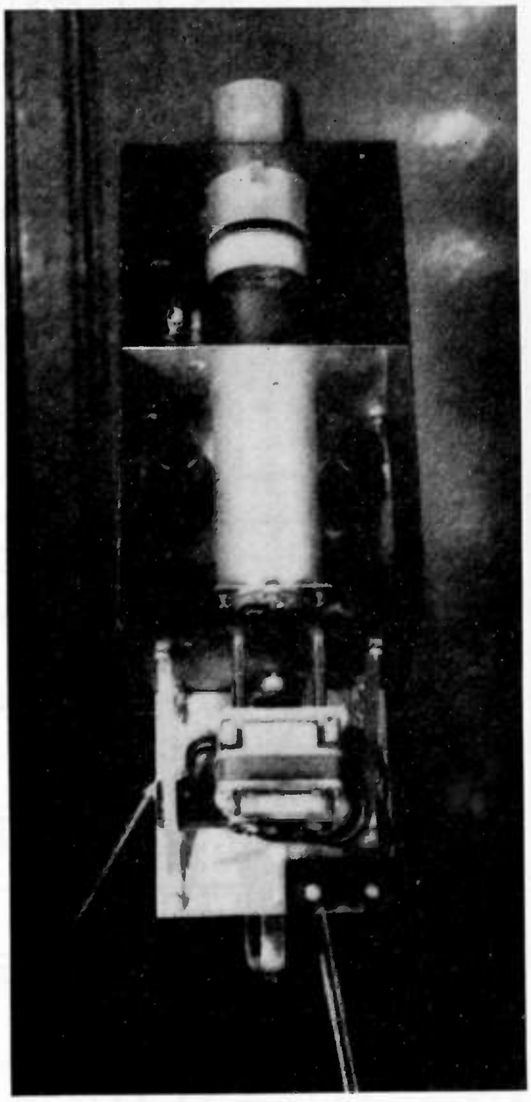


Fig. 3—Camera with top and motor housing removed.

actual focusing control on the control panel consists of a key-switch-type handle which actuates short-action snap switches, so that movement of the handle about $\frac{1}{8}$ inch in either direction advances or retracts the lens. The focusing method has proved very convenient, and the required equipment is both economical and compact.

The camera is connected to the control cabinet by the video and scanning cables, the motor cable, and the power-supply cable, all of which are enclosed in a single rubber tube. Fig. 2 shows the outer appearance of the camera. In Fig. 3 the top and the motor housing have been removed.

IV. Main Chassis

The control cabinet contains three chassis: the main chassis, containing the video amplifier, scanning and

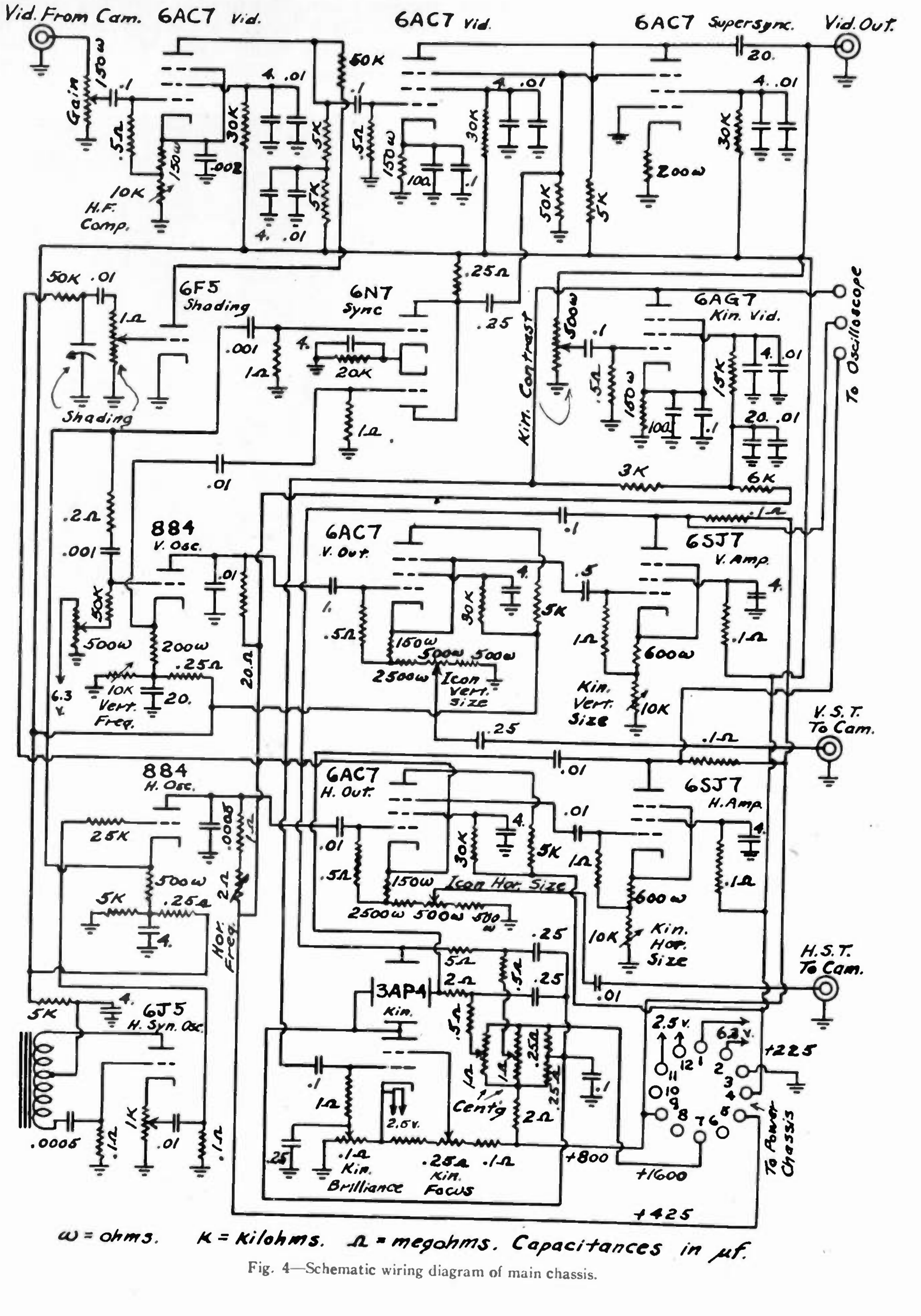


Fig. 4—Schematic wiring diagram of main chassis.

pulse generators and amplifiers, and monitor kinescope; the oscilloscope; and the power supply.

Fig. 4 is the schematic wiring diagram of the main chassis. The horizontal and vertical scanning oscillators employ gas triodes. The vertical oscillator is these horizontal lines to appear on the kinescope screen. However, in using gas-discharge tubes, the vertical return time can readily be made only a few per cent of the horizontal scanning time, so that in the absence of vertical blanking a single, nearly vertical,

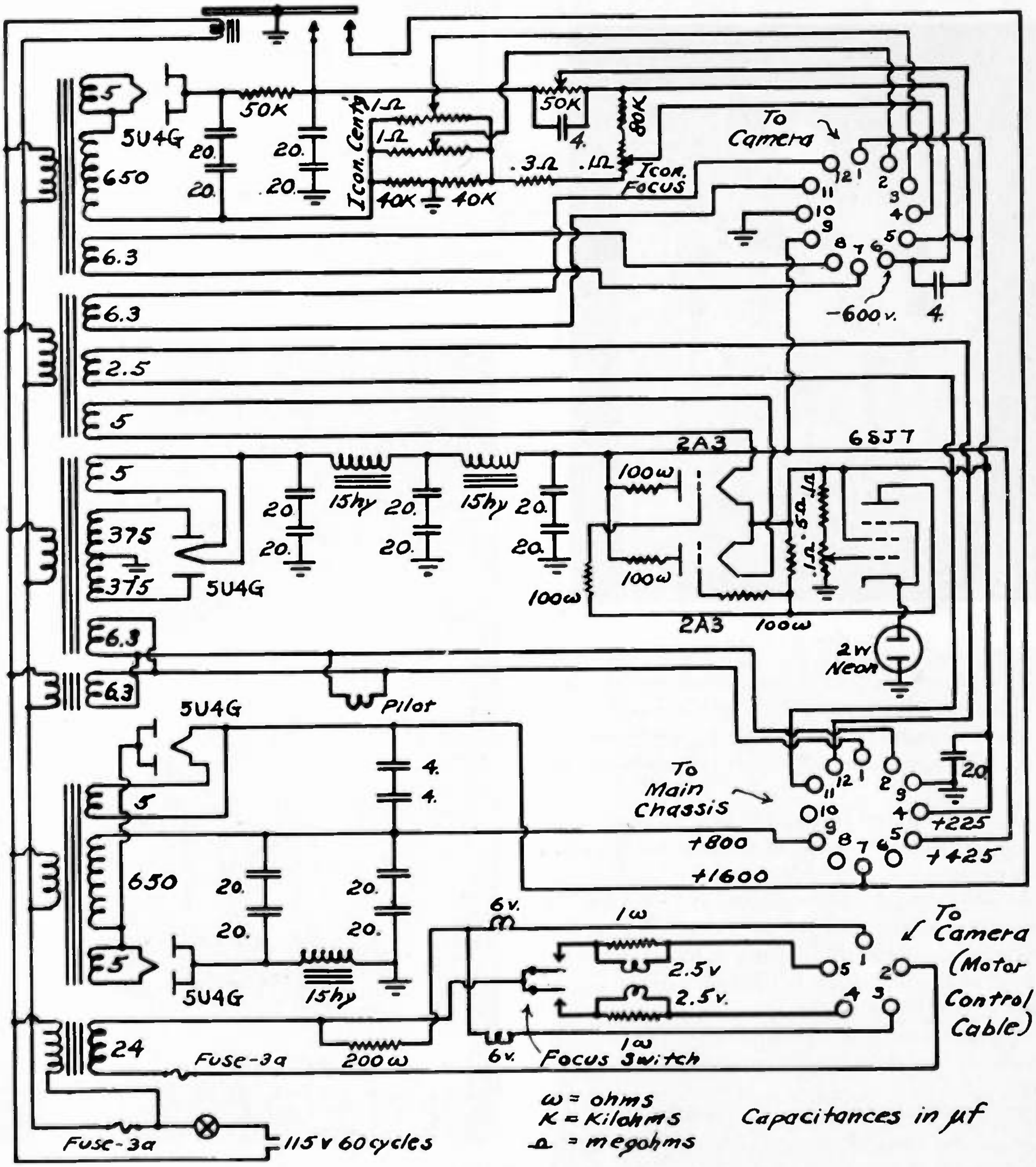


Fig. 5.—Schematic wiring diagram of power-supply chassis.

locked to the 60-cycle power supply, while the horizontal oscillator is locked to a stable 9-kilocycle oscillator. In using high-vacuum tubes as scanning oscillators and discharge tubes, it usually turns out that the vertical return time occupies perhaps ten horizontal lines, which in the absence of vertical iconoscope blanking results in the generation of signals causing line appears on the screen. If the horizontal and vertical oscillators have no rigid frequency relation to each other, this line will occupy different, random positions in successive frames. However, if a small pulse from the horizontal oscillator is injected into the vertical oscillator, the vertical discharge can be made to occur at the same time as a horizontal discharge, even though the vertical oscillator is at the same time synchronized with the 60-cycle supply. The effect is to make the single vertical line occur always at the edge of the scanning raster, leaving the balance free of return trace. It will be noted from Fig. 4 that this scheme has been employed and for this reason no vertical iconoscope blanking is used. (There is no horizontal iconoscope blanking, which is usual.) The kinescope return trace, which appears independently of the iconoscope trace, occupies approximately the same position on the screen.

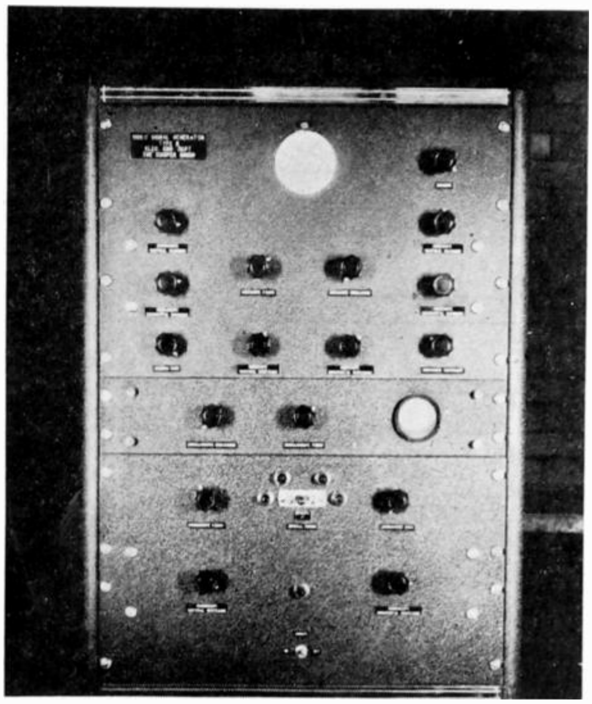


Fig. 6—Front view of control cabinet.

The discharge pulses of the scanning oscillators are amplified and made to serve as synchronizing signals by injection into the output stage of the video amplifier. The injection is accomplished by applying the signals with negative polarity to the suppressor grid and driving beyond cutoff. In order to insure a "supersync" beyond the maximum black video signal, a tube operated in parallel with the video tube is so biased that the synchronizing signals produce a further positive excursion of plate voltage than do the video signals at cutoff.

The scanning oscillators deliver saw-tooth deflecting voltages to the camera via cathode-coupled output stages, at a level of 5 to 10 volts as already mentioned. Single-stage amplifiers deliver the same deflecting voltages to the 3-inch monitoring kinescope, the video signal for which is taken from the outgoing cable through one video stage.

Some experimenting with shading was performed to determine whether any simple arrangement could be used to advantage. It was found that a horizontal shading signal of approximately half-sine shape, with the peak about the center of the picture, appreciably increased the permissible iconoscope beam current. This signal was readily derived from the horizontal deflecting voltage, and involves only a single panel control for amplitude. A variable capacitance, not on the panel, permits changing the shape of the shading waveform.

V. Power-Supply Chassis

The circuit diagram of the power-supply chassis is shown in Fig. 5. The video amplifier and other low-voltage requirements are furnished by a regulated supply of the series-regulation type. The monitor kinescope requires a total direct voltage of 1600, which is obtained from a voltage-doubling rectifier. The same supply furnishes 800 volts to the kinescope deflecting amplifier tubes. The iconoscope requires 600 volts, negative with respect to ground, obtained from a third supply. There are, of course, in addition, the various alternating filament voltages.

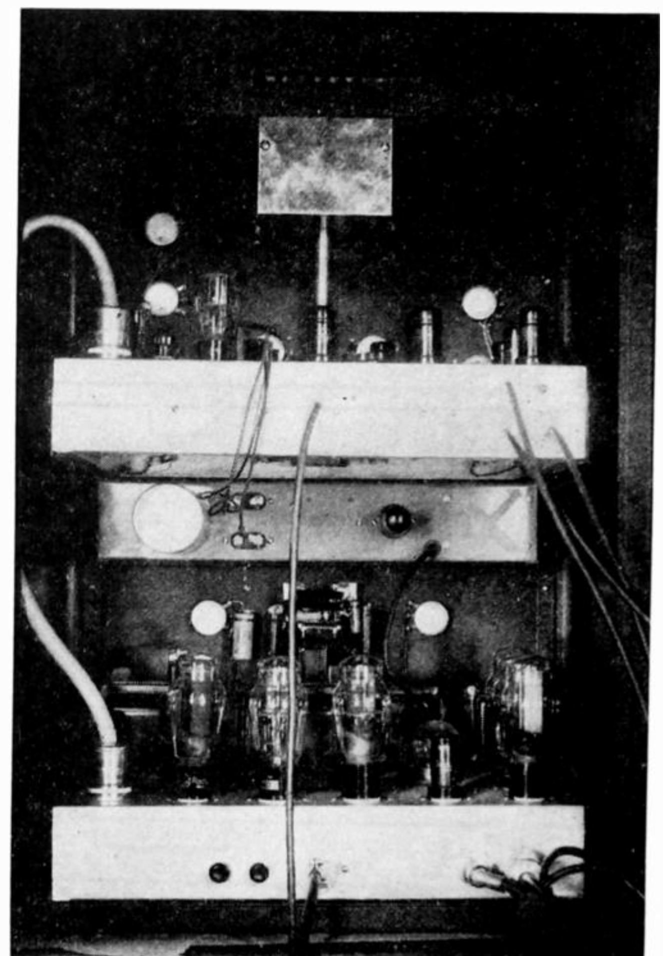


Fig. 7—Rear view of control cabinet with door open.

The power-supply chassis is connected to the main chassis by a 12-conductor plug and cable, and to the camera by a similar plug and cable plus a 5-conductor motor cable.

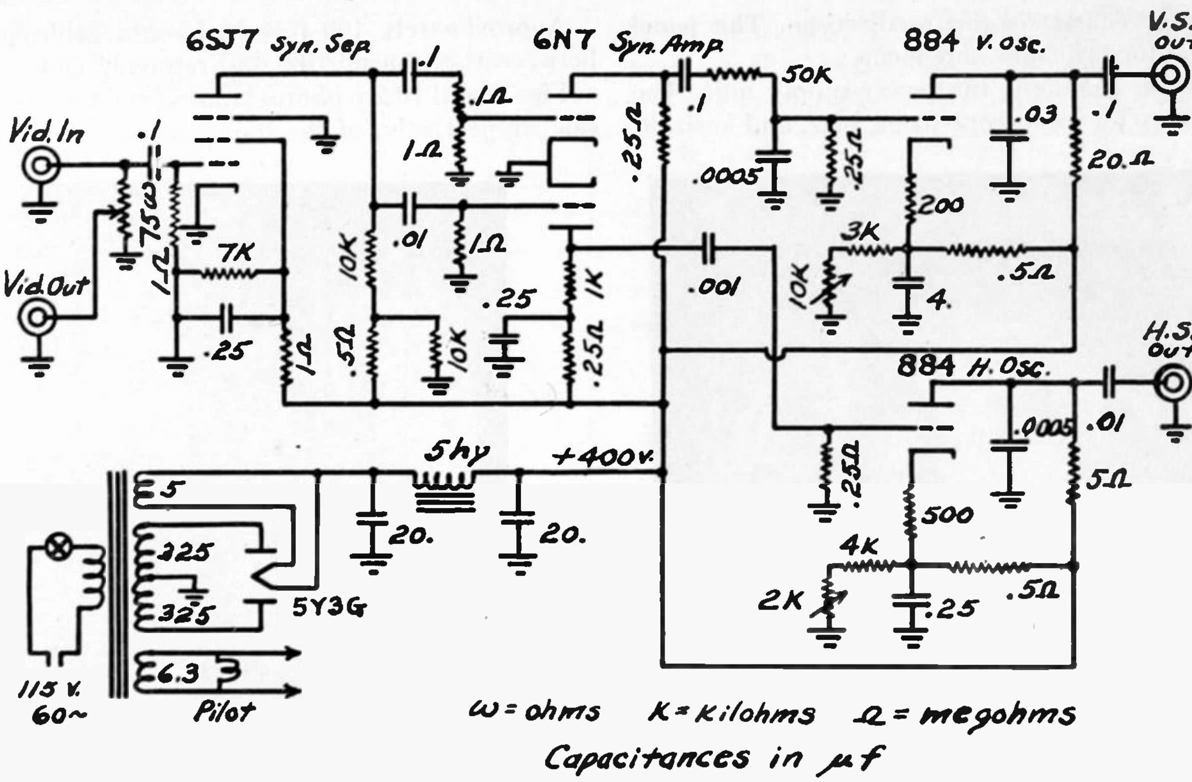


Fig. 8—Schematic wiring diagram of receiving unit.

It will be noted that the motor circuit includes pilot lights indicating when the motor is running and also when the limit switches have stopped the motor at either extreme of travel.

VI. Oscilloscope

The third chassis of the control unit consists of a 2-inch oscilloscope for observing the composite video signal. By using either the frame or line-deflecting voltage as horizontal sweep, either frame or line content may be observed. The oscilloscope circuit is conventional and is therefore not shown.

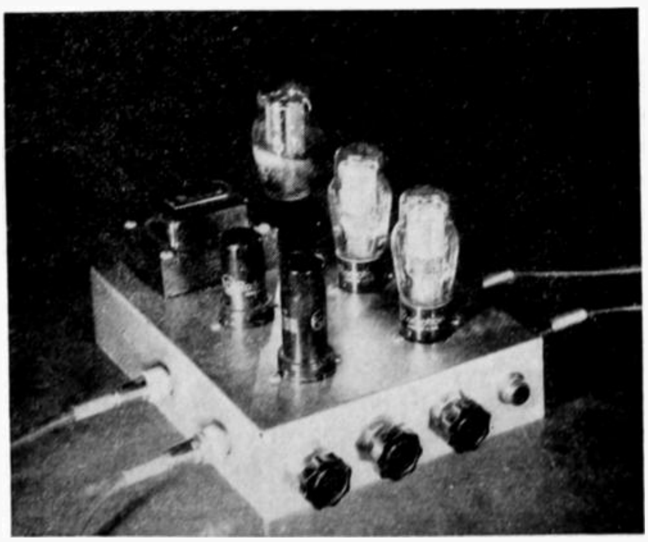


Fig. 9—Top view of receiving unit.

VII. Complete Control Equipment

Figs. 6 and 7 show, respectively, the front and rear of the control cabinet. The top chassis is the main

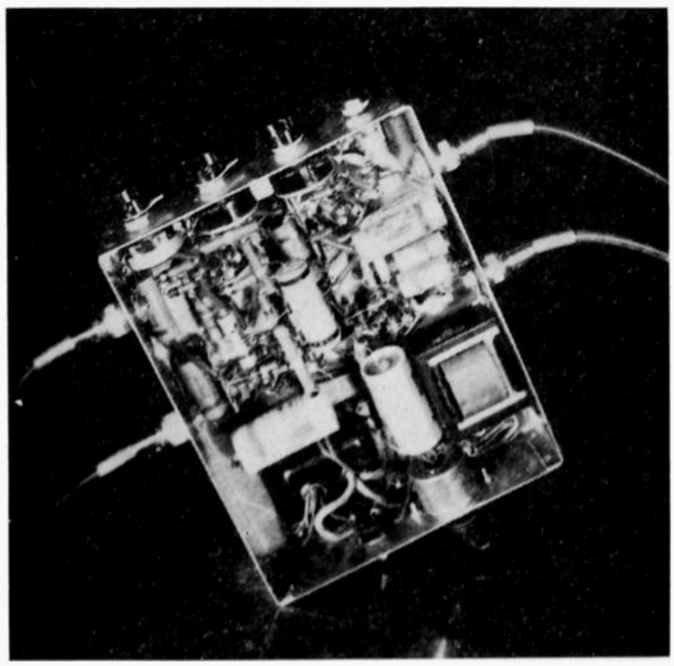


Fig. 10—Underside view of receiving unit.

unit. The controls on this panel are iconoscope vertical and horizontal size; iconoscope shading; camera video gain; kinescope horizontal and vertical size; kinescope horizontal and vertical centering; and kinescope brilliance, focus, and contrast. A detachable light shield may be slipped over the kinescope screen.

The center chassis is the oscilloscope. The panel controls are for brilliance and focus.

The bottom chassis is the power-supply unit. The four knobs are for iconoscope focus, bias, and horizontal and vertical centering. The optical focusing switch and pilots are in the upper center of the panel. The power switch and pilot are in the lower center.

Fig. 11—Photograph taken on monitor screen (Dr. Burdell).

The cabinet measures $28 \times 21 \times 13\frac{1}{2}$ inches.

VIII. Receiving Unit

It is intended that a conventional 5- or 9-inch oscilloscope shall be used as the receiving-end kinescope for the equipment, with the addition of a single-stage video amplifier, and the unit is shown schematically in Fig. 8. This unit comprises a synchronizing-signal separator and amplifier, scanning oscillators like those used at the transmitting end, and a 400-volt direct-current power supply. The oscillators are connected to the respective deflecting amplifiers of the oscilloscope.

Fig. 12—Photograph taken on monitor screen (Students).

Approximately 100 feet of 75-ohm cable are used between the transmitting and receiving ends.

Figs. 9 and 10 are photographs of the top and underside, respectively, of the receiving unit. The controls are the video gain, horizontal hold, and vertical hold. The chassis measures $7 \times 9 \times 2$ inches.

Fig. 13—Photograph taken on monitor screen. (Foundation Building).

IX. Results

Some photographs taken of pictures on the monitor screen are shown. The subject in Fig. 11 was a photograph of Dr. Edwin Sharp Burdell, Director of The Cooper Union. The subjects in Fig. 12 were two obliging students; this was performed in daylight in the laboratory. In Fig. 13, the iconoscope was pointed out the window at the Foundation Building across the street; the Third Avenue elevated structure appears in the foreground.

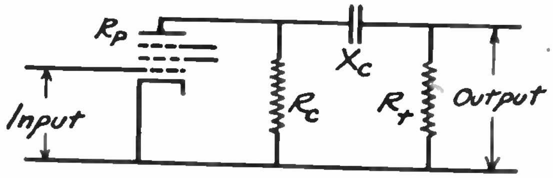


Fig. 14—Pertaining to low-frequency response.

Appendix I

Note on the Low-Frequency Characteristic of a Video Cable

It is of interest to observe that the low-frequency characteristic of a video cable driven from an amplifier plate circuit, as in Fig. 4, differs considerably from the characteristic of the usual interstage coupling.

Referring to Fig. 14, it is readily shown[2] that

[2] F. E. Terman, "Radio Engineering," McGraw-Hill Book Company, New York, N. Y., second edition, 1937, p. 179.

$$\frac{\text{amplification at low frequencies}}{\text{amplification in middle range}} = \frac{1}{\sqrt{1 + (X_c/R)^2}}$$

where

$$R = R_t + \frac{R_c R_p}{R_c + R_p}.$$

In the usual video interstage coupling, R_t is very large compared with R_c, and R_p (with pentodes) is also large compared with R_c. Hence the low-frequency characteristic is determined by the ratio of X_c to R_t. However, when the amplifier is followed by a low-impedance cable, R_t becomes small in comparison with R_c and R_p. The circuit assumes a constant-current aspect, and the relative low-frequency response is much improved.

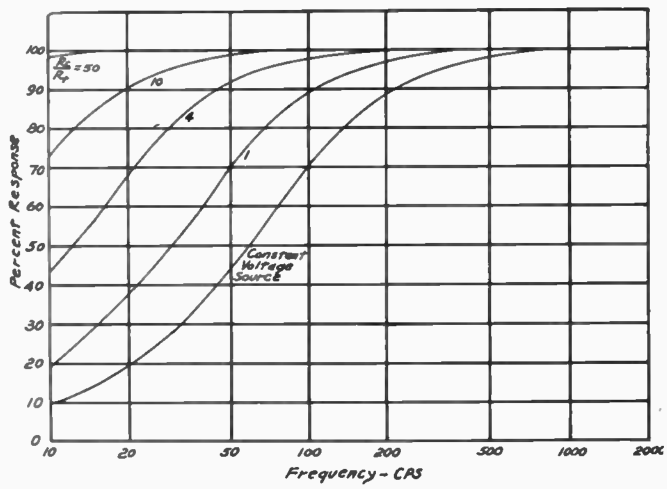


Fig. 15—Variation of low-frequency response with ratio of R_c to R_t (R_t = 75 ohms).

This is brought out by Figs. 15 and 16. In Fig. 16 a 75-ohm cable is assumed with a coupling capacitance of 20 microfarads, and the relative response is plotted for various ratios of R_c to R_t, and for a constant-voltage source. In Fig. 16 a 75-ohm cable is also assumed, and two values of capacitance chosen. The dotted curves give the relative response for a constant-voltage source, and the solid curves for $R_c/R_t = 50$.

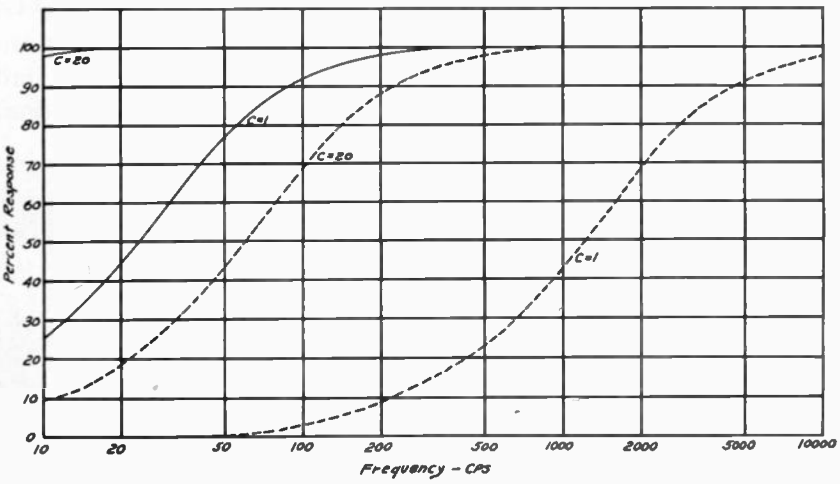


Fig. 16—Comparison of low-frequency response: constant-voltage source (dotted curves) and $R_c/R_t = 50$ (solid curves).

Appendix II

Note on the Low-Frequency Response of the Type 1847 Iconoscope

This iconoscope does not have a direct connection to the signal plate but instead is dependent for signal connection on a capacitance of 50 micromicrofarads through the wall of the tube. This immediately introduces a fundamental low-frequency deficiency, which however is not nearly as severe as might appear, for it is the high internal resistance of the iconoscope signal circuit which determines the low-frequency response, rather than the relatively low external resistance. The internal resistance is the dynamic resistance as determined from the secondary-emission characteristic; hence it may be expected that the low-frequency response will vary to some extent with beam current.

Orthicon Portable Television Equipment*

M. A. TRAINER†, ASSOCIATE, I.R.E.

Summary—Recently developed orthicon portable television equipment is described. Because of the greater light sensitivity of the orthicon, it is expected that the equipment will fill a need for lightweight equipment to be used under adverse light conditions. Among the novel features of the design are the use of a forced-air-cooled transformer in the regulated power-supply unit, an all-electronic synchronizing generator, gamma control, and keyed diodes for black-level setting.

The units required for a one-camera setup are: camera and tripod, camera control unit, power-supply unit, pulse unit, and shaping unit. The total weight, exclusive of camera cable, is 370 pounds and the power required is 1250 watts.

CONSIDERABLE field experience with iconoscope portable television equipment indicated a need for camera equipment that would produce satisfactory pictures under very unfavorable lighting conditions. It was felt that if the orthicon, with its superior sensitivity, could be used as a camera tube in portable equipment, the desired results would be obtained. Many construction problems had to be solved, particularly in the design of the camera, so that the finished equipment would be sufficiently portable for field use.

Experience has also shown that the camera must be equipped with a view finder and focusing system that would not unduly tire the camera operator and

* Decimal classification: R583. Original manuscript received by the Institute, July 2, 1941. Presented, Summer Convention, Detroit, Michigan, June 25, 1941.

† RCA Manufacturing Company, Victor Division, Camden, New Jersey.

the power-supply unit should be regulated to provide satisfactory operation on the power-supply lines encountered in field work.

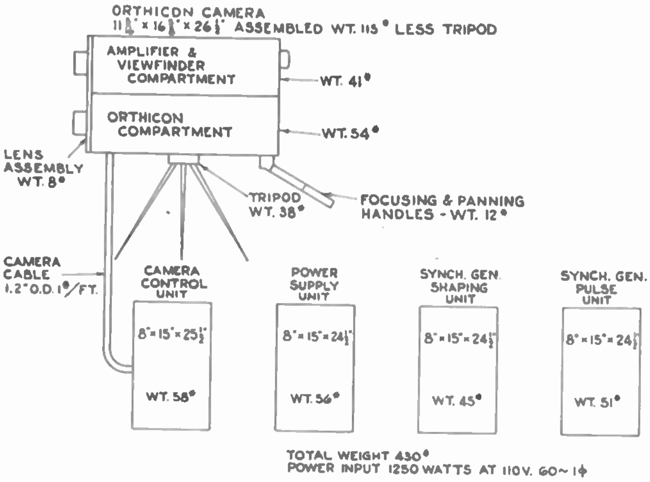


Fig. 1—Schematic diagram of orthicon television equipment.

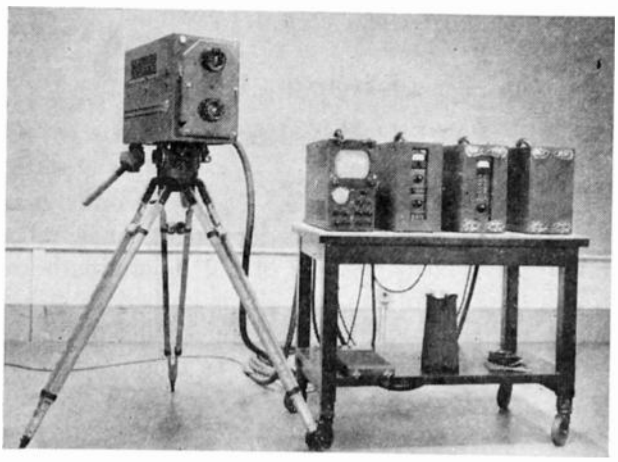


Fig. 2—Orthicon portable television equipment.

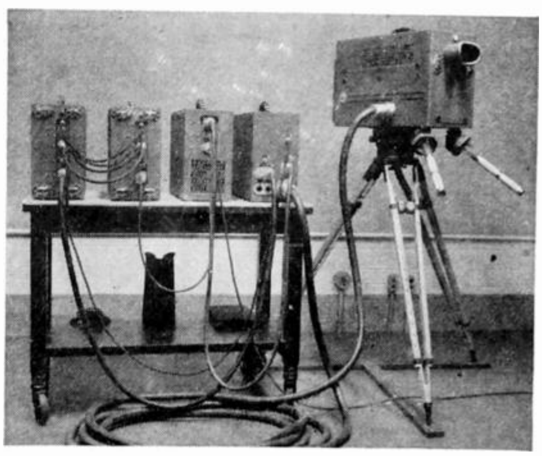


Fig. 3

The diagram in Fig. 1 shows the number of units required for a one-camera setup. The weights and power requirements of the various units are indicated. Fig. 2 is a photograph of the actual equipment. Fig. 3 is a rear view of the equipment showing the plug and interconnecting cable arrangement for the several units. Fig. 4 is a close-up view of the four suitcase-type units. A detachable cover is provided for the controls and cathode-ray tubes in the camera control unit. The knobs and meters on the power-supply unit and the pulse unit are recessed to avoid injury during transportation.

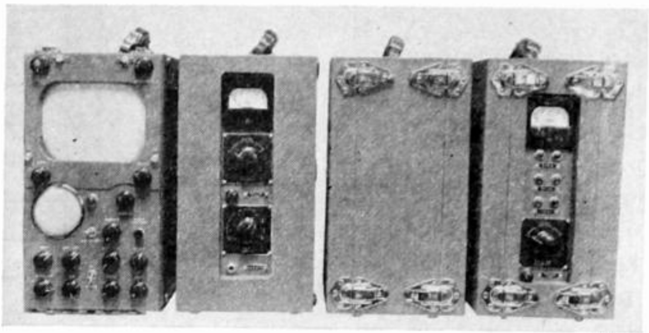


Fig. 4—Camera control, power-supply, shaping and pulse units. The last two constitute the synchronizing generator.

Camera

Fig. 5 shows the orthicon camera. In order to facilitate handling, the camera was made in two main sections. The upper half contains the optical view finder and the amplifiers. The lower half houses the orthicon, the orthicon focusing coil, and the lens-carriage mechanism.

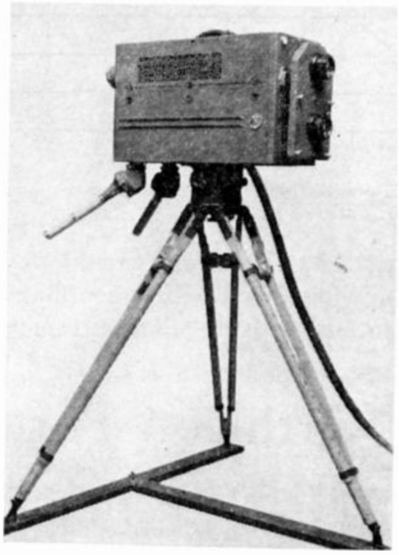


Fig. 5—Orthicon television camera.

A lens-adapter plate on which both the view-finding and the picture-taking lenses are mounted is fastened, by means of thumbscrews, to the lens-carriage mechanism. When it is desired to change to lenses of a different focal length, a second adapter plate with the desired lenses is substituted.

The two handles at the rear of the camera provide convenient means for tilting, panning, and focusing the camera. The upper section of the right handle provides for coarse focusing while the lower section allows for fine focusing. Both handles are readily detachable from the camera.

The optical view finder produces an image of the scene that is correct left to right and top to bottom. Parallax correction is provided and is automatically adjusted when lenses of different focal lengths are used.

A field of view approximately 30 per cent higher and 15 per cent wider than that included by the picture-taking lens is provided by the view finder. This is a distinct advantage because it allows the camera operator to avoid panning into unwanted parts of a scene.

Fig. 6 shows the camera with the compartment covers open. The deflection and blanking amplifier may be seen in the upper compartment. Electrostatic horizontal deflection and magnetic vertical deflection are provided for the orthicon. The focusing coil and the lens-carriage drive mechanism can be seen in the lower compartment. The focusing coil is mounted in a trunion in order that the magnetic field may be aligned with the axis of the orthicon.

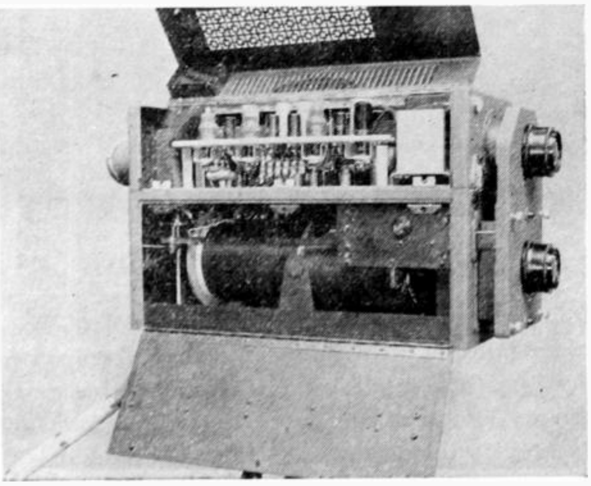


Fig. 6

Fig. 7 shows the 5-stage camera video amplifier in the upper compartment. The amplifier is shock-mounted to minimize microphonics. The receptacle for the camera-cable plug is visible in the lower compartment. All electrical connections between the upper and lower compartments are made by means of self-aligning plugs and receptacles. When the camera is being set up, the upper half is literally plugged into the lower half. Receptacles are provided at the rear of the lower compartment for the camera-operator's headphones and breastset by means of which he may communicate with the other personnel. Camera cables up to 500 feet in length may be used between the camera and the camera control unit.

Camera Control Unit

Fig. 8 shows the circuit side of the camera control unit. The 7-inch kinescope and the 3-inch oscilloscope can be seen in their magnetic shields. All cable connections are made at the rear of the unit. The controls that are used frequently for adjusting the video amplifier, the orthicon, and the kinescope are located on the front of the unit. Additional controls that require infrequent adjustment are available at the top front of the unit. Three unusual features of the video amplifier

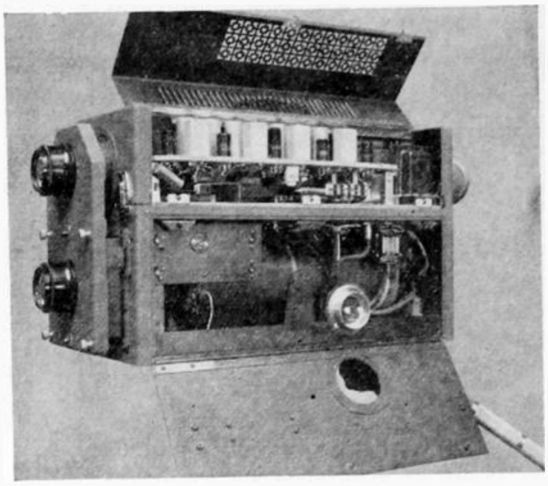


Fig. 7

are the black-level setter or "clamp circuit," the linear clipper, and the gamma control.

The black-level-setter tube is a double diode which is keyed at the end of every horizontal line. By this means, the top of every horizontal blanking impulse is returned to a fixed voltage level. Hum and low-frequency microphonics are very effectively removed from the video signal. An adjustment is provided to allow any desired amount of "setup" between the picture blacks and the top of blanking impulses.

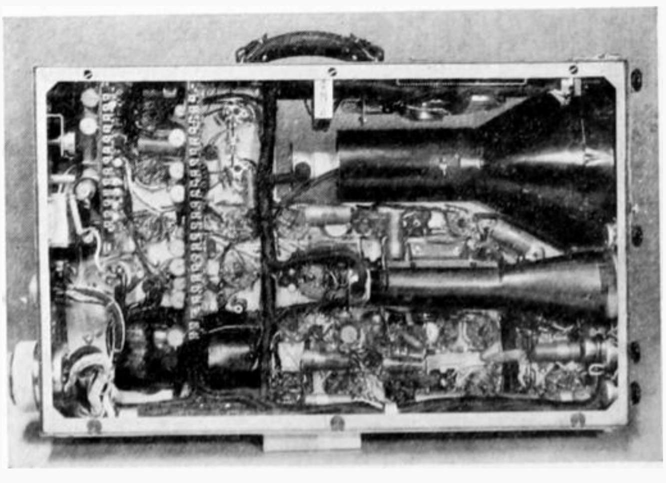


Fig. 8—Orthicon camera control unit—circuit side.

The clipper tube is direct-current connected to the black-level-setter tube and the clipping is done in the plate circuit by means of a diode which is adjusted to open the circuit at a predetermined voltage level. The clipping point is well defined and the black portion of the video signal is not compressed as it is in the usual type of clipper.

When an orthicon picture is viewed on a kinescope with its curved grid-voltage versus light-output characteristic, the black portions of the picture are compressed. To provide a more pleasing picture, two kinds of gamma correction are incorporated in this unit. When gamma *A* is used, the video signal is amplified by three tubes in parallel, two of which amplify only the black portion of the signal. When gamma *B* is used, the plate impedance for one of the video amplifier stages is increased for the black portion of the video signal. The disadvantage of this method is that the high-frequency components of the black portion of the video signal are attenuated. However, the high-frequency noise that is present in the black portion of the signal is also attenuated. It is felt that gamma-*B* correction may prove useful when scenes are being televised under unfavorable lighting conditions. The amount of gamma-*A* or gamma-*B* correction used can be varied by the operator.

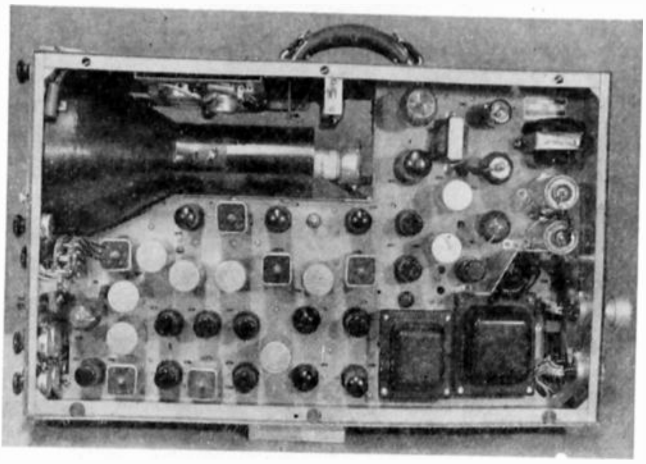


Fig. 9—Camera control unit—tube side.

The deflection and video amplifier circuits for the kinescope and the oscilloscope are located in this unit. The video signal level at the output is 1.5 volts peak to peak across 76 ohms. For a single-camera setup, the synchronizing signal is added to the video signal in this unit.

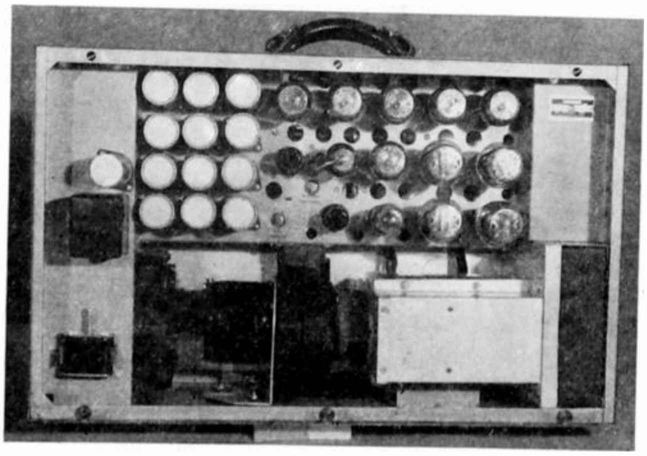


Fig. 10—Power-supply unit—tube side.

Fig. 9 shows the tube side of the camera control unit. The high-voltage rectifier that supplies second-anode voltage for the kinescope is visible in the lower rear corner of the unit.

Power-Supply Unit

Fig. 10 shows the tube side of the power-supply unit. This unit supplies 600 milliamperes at 300 volts direct current. The voltage is regulated and is maintained substantially constant despite line-voltage or load-current fluctuations. An additional 200 milliamperes is supplied for the orthicon focusing coil. This current is regulated and is maintained constant although the resistance of the coil changes considerably with temperature.

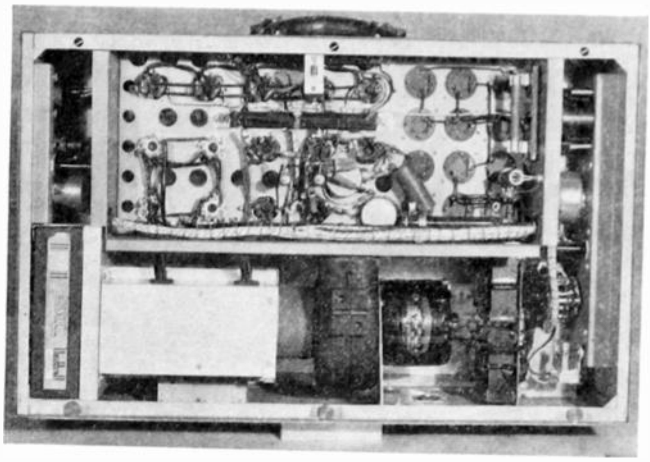


Fig. 11—Power-supply unit—circuit side.

The power transformer is forced-air-cooled. The air, after being drawn through the air filter and transformer, is forced past the regulator and rectifier tubes and provides very effective cooling. The combined weight of the blower and special transformer is about one third the weight of a transformer of conventional design having the same power rating.

Fig. 11 shows the circuit side of the same unit. A line voltmeter is mounted on the front panel. By means of a plug-in meter, the plate current of each individual regulator tube can be read.

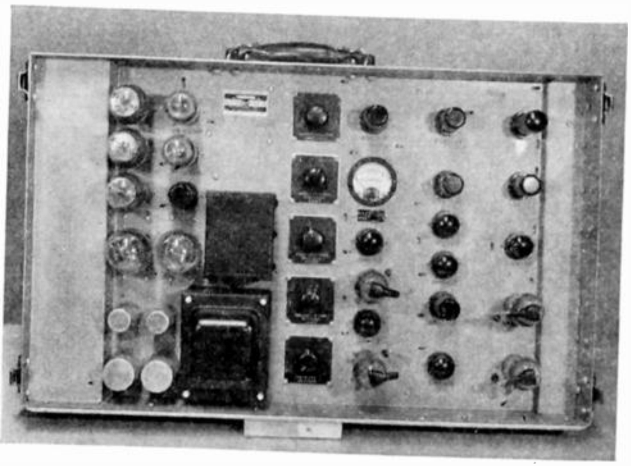


Fig. 12—Orthicon pulse unit—tube side.

Synchronizing Generator

The synchronizing generator is comprised of two units, the pulse unit and the shaping unit. The Radio Manufacturers Association type of synchronizing pulses for 525-line, 30-frame, 60-field pictures, and all the necessary driving and blanking pulses are provided. The generator is all electronic and has a self-contained regulated power supply.

Fig. 12 shows the tube side of the pulse unit. The particular unit illustrated was designed for 441-line operation as indicated by the dial markings.

A master oscillator operating at twice line frequency

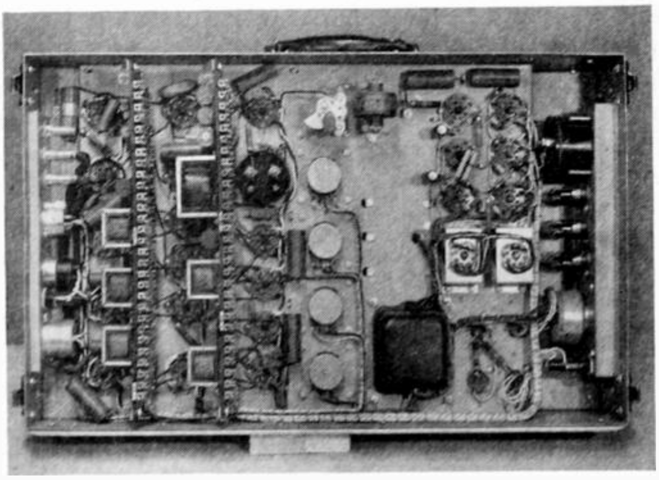


Fig. 13—Orthicon pulse unit—circuit side.

is stepped down by means of counter circuits to field frequency. Provision is made for locking the generator with the 60-cycle power supply. The regulated power supply is visible at the left. A line voltmeter and 6 test buttons are mounted on the front panel. The test buttons provide an easy means for checking the adjustment of the counter circuits without opening the side covers. Fig. 13 shows the circuit side of the same unit.

The tube side of the shaping unit is shown in Fig. 14. As the name implies, the function of this unit is to shape properly the pulses received from the pulse unit. Controls are provided to adjust the width of the various pulses.

Fig. 15 is a view of the circuit side of the shaping unit. Although this unit contains 26 tubes with their associated circuit components, all parts are readily accessible. No delay unit is required for camera cables up to 500 feet in length.

Master Control Unit

A master control unit for use with two or three orthicon cameras is being constructed. The video signal output of any camera control unit is switched at the input to the master control unit. A pilot light indicates which camera is connected and at the same time a light on the selected camera warns the operator that the camera is "on the air." A 7-inch monitoring kinescope and a 5-inch oscilloscope tube are included in this unit. Operating tests have shown that a large oscilloscope tube greatly facilitates setting and maintaining the proper signal levels at the output of the master control unit. Synchronizing signal is added to the video signal in this unit. The video signal level at the output

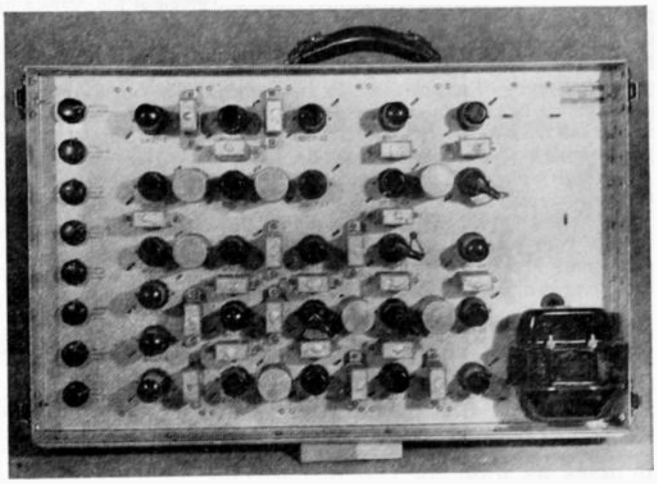


Fig. 14—Orthicon shaping unit—tube side.

is 5 volts peak to peak across 76 ohms. A power-supply unit similar to the one previously described supplies power to the master control unit.

Conclusion

Orthicon portable television equipment is expected to fill a need for equipment to be used where the artificial or natural light is insufficient for satisfactory

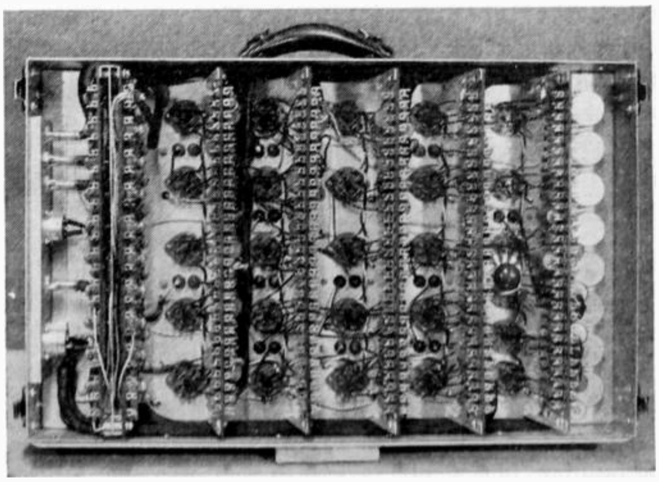


Fig. 15—Orthicon shaping unit—circuit side.

iconoscope pictures. This condition is frequently encountered when televising indoor sporting events or outdoor scenes in the late afternoon.

Acknowledgment

The design and development of this equipment has been made possible by the co-operative efforts of many fellow engineers, to whom I express my appreciation.

The Design and Development of Three New Ultra-High-Frequency Transmitting Tubes*

CECIL E. HALLER†, MEMBER, I.R.E.,

Summary—*A discussion and review are given of the service and design requirements of transmitting tubes intended for application in the ultra-high-frequency spectrum. These requirements fall in two classes: (1) those imposed by service conditions and (2) those imposed by the frequency at which the tube is operated. The fulfilling of these requirements has led to certain design and manufacturing problems, such as the reduction of grid emission, choice of anode material, choice of a suitable mechanical structure, etc.*

A description of the novel features of construction and the operation of three new ultra-high-frequency transmitting tubes are also given. Two of these tubes are the RCA-815 and RCA-829 which are push-pull beam tetrodes while the third tube, the RCA-826, is a triode. Some precautions necessary for obtaining satisfactory operation with these tubes are given.

THE recent activity in the development of ultra-high-frequency communication has created a demand for transmitting tubes having carrier power outputs of the order of 30 to 60 watts at frequencies as high as 150 megacycles and in some instances as high as 250 megacycles. In order to clarify the tube-design requirements it is well to enumerate the more common uses to which tubes are applied in ultra-high-frequency communication. These uses include police, aviation, marine, television, amateur, and point-to-point communication and employ continuous-wave transmission, amplitude modulation, or frequency modulation. A survey of these uses yields some interesting requirements which can be divided in two classes: (1) those imposed by service conditions; and (2) those imposed by the frequency spectrum in which the tubes are operated. The requirements of tubes suitable for operation in this frequency spectrum have been discussed at length by others[1–6] but are summarized in order that they can be considered in connection with the design and development of the three new ultra-high-frequency transmitting tubes described in this paper.

Tube-Design Requirements Imposed by Service Conditions

(1) The tube must be strongly constructed since it must withstand the mechanical shocks incident to all types of services in which it is applied, such as, for example, those which might be experienced when a police car drives over a curb or rugged terrain. It should not produce objectionable microphonics under such conditions.

(2) The tube must usually be designed for a plate voltage of less than 1000 volts and often as low as 300 volts. The use of low voltages offers many advantages to the equipment designer in insulation and weight requirements. In applications where power economy is a factor, it is desirable that all tubes of the transmitter operate at a common plate voltage in order to eliminate the need for power-consuming, voltage-dropping resistors.

(3) Air cooling must be used as the size and weight of liquid-cooling equipment is prohibitive in most cases.

(4) A tube is required which gives the highest efficiency with as low value of driving power as possible in order to reduce the size, weight, and cost of component parts of the transmitter and associated power supplies. In terms of tube design, a high-perveance tube is indicated, preferably of the screen-grid type.[7]

(5) A filament or heater operating at 6.3 or 12.6 volts should be used in those tubes intended for mobile services in order to avoid series dropping resistors and attendant loss of power.

(6) Police, marine, and aviation services usually require that the tube occupy a minimum of space and be of as small over-all length as possible.

(7) The construction must be such that the tube, in case of failure, can be replaced readily and quickly in the transmitter.

Design Requirements Imposed by Frequency of Operation

(1) Electrically the tube should lend itself readily to circuit design and what is, perhaps, more important if transmission lines are used, should not excessively shorten either the input or output lines because of its input and output capacitance. An appreciable part of the circuit must appear outside the tube.

(2) The dimensions and material of the electrode leads must be such that they will safely carry the ultra-high-frequency currents without overheating or introducing any appreciable amount of impedance.

(3) Transit-time losses must be minimized.[5,8] This

* Decimal classification: R339. Original manuscript received by the Institute, June 5, 1941. Presented, Summer Convention, Detroit, Michigan, June 24, 1941.

† Research and Engineering Department, RCA Manufacturing Company, Inc., Harrison, New Jersey.

[1] C. E. Fay and A. L. Samuel, "Vacuum tubes for generating frequencies above one hundred megacycles," Proc. I.R.E., vol. 23, pp. 199–212; March, 1935.

[2] M. J. Kelly and A. L. Samuel, "Vacuum tubes as high frequency oscillators," *Trans. A. I. E. E.* (*Elec. Eng.*, November 1934), vol. 53, pp. 1501–1517; 1934; *Bell Sys. Tech. Jour.*, vol. 14, pp. 97–134; January, 1935.

[3] W. G. Wagener, "The developmental problems and operating characteristics of two new ultra-high-frequency triodes," Proc. I.R.E., vol. 26, pp. 401–414; April, 1938.

[4] A. L. Samuel and N. E. Sowers, "A power amplifier for ultra-high frequencies," Proc. I.R.E., vol. 24, pp. 1464–1483; November, 1936; *Bell Sys. Tech. Jour.*, vol. 16, pp. 10–34; January, 1937.

[5] A. K. Wing, Jr., "A push-pull ultra-high frequency beam tetrode," *RCA Rev.*, vol. 4, pp. 62–72; July, 1939.

[6] W. G. Wagener, "The requirements of a new ultra-high-frequency power tube," *RCA Rev.*, vol. 2, pp. 258–265; October, 1937.

[7] O. H. Schade, "Beam power tubes," Proc. I.R.E., vol. 26, pp. 137–181; February, 1938.

[8] A. V. Haeff, "Effect of electron transit time on efficiency of a power amplifier," *RCA Rev.*, vol. 4, pp. 114–122; July, 1939.

implies close-spaced electrodes and, consequently, high dissipation densities per unit area.[9]

(4) The insulation between electrodes must not introduce any appreciable losses and also must not disintegrate at high frequencies or at elevated temperatures.

(5) If the tube is of screen-grid type, it must perform as a stable amplifier without neutralization. It must, therefore, have a low feedback capacitance and also must not have other characteristics, such as common impedances, which provide unwanted energy interchange. If the tube is a triode, it must be capable of satisfactory neutralization.

(6) Provisions must be made to minimize random and stray electrons which may, due to their long time of flight, cause poor efficiency. Stray electrons striking a glass envelope may lead to gas evolution directly or in more severe cases to gas evolution indirectly through glass decomposition. Furthermore, under certain conditions they can give rise to secondary emission and can, thus, cause a portion of the glass surface to assume a positive potential. This charged portion may attract more electrons and cause localized bulb heating and consequent softening of the bulb to the point where failure occurs.

Special Problems of Design and Manufacture

Departures from conventional design procedures such as reduction in size, decrease in interelectrode spacings, and the consequent increase in dissipation densities made in order to secure better high-frequency performance have given rise to some fundamental design problems. Probably the most important of these is the elimination of grid emission. Grid emission results from operation of the grid at elevated temperatures produced by the inherently high dissipation densities and may be intensified by a deposit of active material from the emitter either during manufacture or during operation. Some basic means must be provided which will make the grid operate at temperatures below that at which it emits objectionably. This reduction in temperature may be obtained through removing heat more rapidly from the grid either by (a) conduction, through the use of copper side rods for the control grid and the use of grid wire of high thermal conductivity or (b) radiation, by carbonizing any part from which heat can be radiated so that the resultant temperatures of electrodes adjacent to the grid are reduced.

Considerable improvement may be had by operation of the cathode at a temperature low enough to provide satisfactory emission but yet not high enough to cause excessive evaporation of active emitting material to the grid. Gains can also be obtained in screen-grid types of tubes by designing the screen so that its dissipation and consequently its temperature is reduced. The beam type of tube offers much advantage in this respect.[7] Finally, a strategic choice of materials has important benefits in the reduction of grid emission. It is well known in the case of oxide-coated emitters that certain nickel-base alloys behave much better than others so far as causing grid emission is concerned and it is also well known that certain grid materials are apt to emit more copiously than others.

The increase in safe anode-dissipation density can be accomplished by using an anode whose effective radiating area is increased by the use of fins or by the roughening of the surface. It may also be increased by using an auxiliary material on the surface which increases the heat-radiating properties by either producing an increase in thermal emissivity or in surface area. Although carbon has a high emissivity, its use presents some manufacturing problems in small high-frequency tubes. It is difficult to obtain a good low-resistance contact with the carbon, a very important requirement at high frequencies. In the case of oxide-coated emitters, the degassing temperature of carbon lies near or above that to which the emitter can be heated without impairing its emission properties. These disadvantages of carbon make the choice of a metal anode material preferable. For one of the new tubes described in this paper, a carbonized nickel anode is used with a fin structure which effectively doubles the power it can radiate. In another tube type, a zirconium-coated molybdenum anode is used which has a fin structure so constructed that it has a power-handling capability that approaches that of carbon.[10,11]

Operation of electrodes at higher temperatures and the larger ratio of mass of metal to volume of the envelope of the tube requires a most active and efficient getter. Batalum getter[12,13] is quite satisfactory and in the case of zirconium-coated anodes, zirconium itself exhibits a remarkable getter action.

At the higher frequencies the choice of material and the location of insulators is very important in order to avoid localized heating which in addition to the attendant loss of power may cause the release of gases and the consequent impairment of the emission characteristic.

Because of space requirements and the necessity of short direct leads, the volume of the envelope must be minimized. This requires that the glass of envelope and stem be very stable and that it also be carefully processed in order to remove surface contaminations and gases so that little or no gas be released during the life of the tube. The release of gas from glass proceeds quite rapidly at elevated temperatures or undue electron bombardment; hence, it is imperative that few

[9] B. J. Thompson, "Review of ultra-high-frequency vacuum tube problems," *RCA Rev.*, vol. 3, pp. 146–155; October, 1938.

[10] J. D. Fast, "Zirconium," *Foote-Prints*, vol. 10, pp. 1–24; December, 1937.

[11] J. D. Fast, "Zirconium as a getter," *Foote-Prints*, vol. 14, pp. 22–30; June, 1940.

[12] E. A. Lederer, "Recent advances in barium getter technique," *RCA Rev.*, vol. 4, pp. 310–318; January, 1940.

[13] E. A. Lederer and D. H. Walmsley, "Batalum—A barium getter for metal tubes," *RCA Rev.*, vol. 2, pp. 117–123; July, 1937.

stray electrons strike the glass and that it operate at not too high temperature. Forced-air cooling may be used to advantage in some instances to reduce the bulb temperature. As the mount size is reduced and leads are shortened, the problem of sealing the glass envelope to the stem without excessive oxidation or burning of the tube electrodes becomes a major problem.

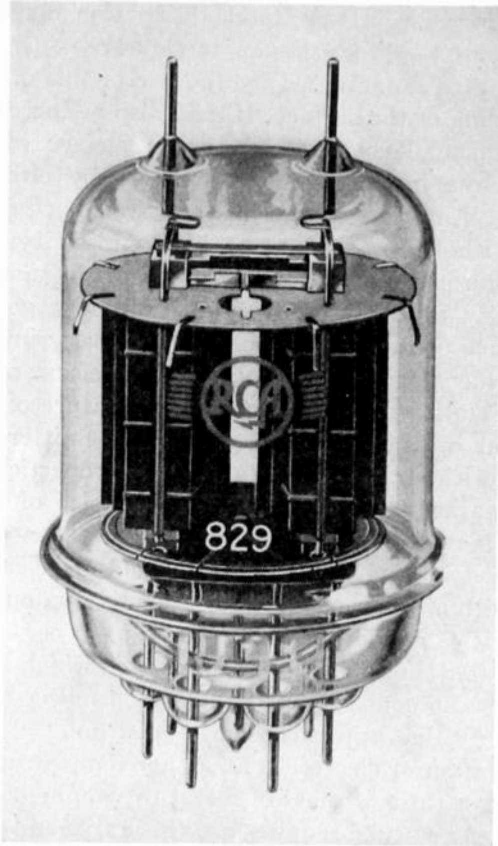


Fig. 1—A push-pull beam transmitting tube having oxide-coated cathodes and a total plate dissipation of 40 watts. It may be operated at full rating to frequencies as high as 200 megacycles.

The Three New Ultra-High-Frequency Transmitting Tubes

Three new ultra-high-frequency transmitting tubes were designed and developed in accordance with these service and frequency requirements to fulfill specific commercial needs. Two of these tubes, the RCA-829 and the RCA-815 are push-pull beam power tetrodes, and the other, the RCA-826, is a triode.

RCA-829

The RCA-829 is a push-pull beam power tetrode of the heater-cathode type. (Fig. 1.) It contains two beam power units within one envelope. The total maximum plate dissipation is 40 watts. It is capable of handling power inputs of 120 watts with very low driving power at frequencies as high as 200 megacycles and may be operated at reduced ratings at frequencies as high as 250 megacycles. The RCA-829 was developed to meet the need for a tube that would be suitable for mobile operation and would deliver a carrier power output of 50 watts at 200 megacycles.

A consideration of the fundamental requirements and experience on the type RCA-832 (a push-pull tetrode)[6] led to the choice of a structure similar in many ways to the RCA-832 but with greater power capabilities. Inasmuch as it was not desirable to raise the plate voltage above 500 volts, a longer mount structure was used in order to obtain an input current which would fulfill the output power requirements. This change led to several interesting mechanical and electrical problems in the design of the RCA-829.

In the early developmental tubes, considerable energy interchange appeared between the plate circuit and grid circuit, so much that operation of the tube as a power amplifier at 200 megacycles was very difficult. When the plate circuit was tuned through resonance, with the grid circuit excited in a normal manner and no voltage applied to the anodes of the tube, measurements showed more than 1 watt of power being fed through the tube from the grid circuit into the plate circuit. This large amount of power did not appear to be due entirely to the direct feedback capacitance of the 829. All attempts at internal neutralization of the tube were unsuccessful. Furthermore, it also appeared that neutralization in any form would be extremely difficult if not practically impossible at these ultra-high frequencies. This feed-through of power or apparent increase in feedback capacitance was found to be due to the series inductance of the tube electrodes produced by their respective lengths. This inductance caused a radio-frequency potential to exist between the top and bottom of each respective tube electrode. Such a condition was particularly objectionable in the case of the beam-plate assemblies and screens as it gave rise to an indirect feedback which produced energy interchange between the control-grid circuit and the plate circuit and prevented stable amplifier operation. The magnitude of this series inductance was very much reduced by cross-connecting the screens of each unit at both the top and bottom of the mount.[14] Similarly, the beam-plate assemblies and cathodes of each unit were cross-connected. Fig. 1 illustrates the method in which these connections are made at the top of the tube. Tubes made in this manner are quite stable when operated as power amplifiers at a frequency of 200 megacycles. After this change was made only a small fraction of a watt was fed through the tube.

The output capacitance is approximately 7.0 micromicrofarads. This value is neither excessive nor objectionable as it permits an external circuit of about 6 inches in length for operation at 200 megacycles with a quarter-wave parallel line having a surge impedance of approximately 180 ohms. The input-circuit capacitance is approximately 15 micromicrofarads. This

[14] The writer is indebted to Mr. Bernard Salzberg, formerly of the RCA Manufacturing Company (Harrison, N. J.), now of the Naval Research Laboratories (Anacostia, D. C.), for this method of reducing electrode inductance.

value at a frequency of 200 megacycles may or may not permit the use of a quarter-wavelength line depending upon the surge impedance of the line. Consideration of the factors contributing to input capacitance such as grid-to-cathode spacing, grid-to-screen spacing, cathode width, cathode length, etc., indicated that this value could not be materially reduced without seriously impairing the perveance, mechanical stability, and high-frequency characteristics.

Some means had to be provided for bracing the mount when the length was increased, which would not appreciably impair the high-frequency characteristics of the tube. This was accomplished by using staples in the top mica bearing against the inner bulb surface. The two large openings in the top mica which are spaced between the two units are provided to remove the mica from a position of high field gradient. With the tube under much higher than rated anode voltage and at a frequency of 200 megacycles, it was discovered that the mica was heated to a point of showing color in the position where the holes are now located.

Nickel side rods on the No. 1 grid which had been used satisfactorily in the type 832 were found to be

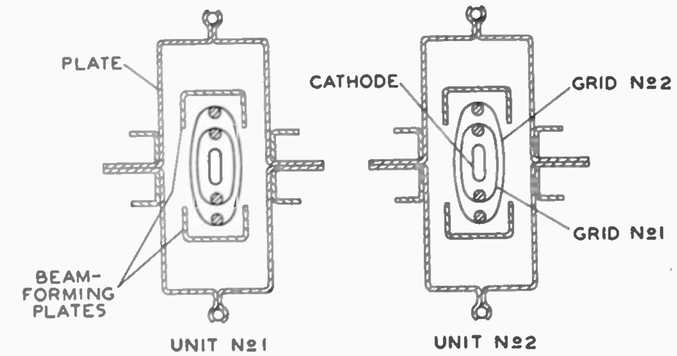


Fig. 2—Horizontal cross section of the RCA-829 electrodes.

unsuitable because of grid emission in the type RCA-829. This was due to the increased length of the grid. The center section operated at a higher temperature due to the proportionally lower amount of heat conducted away by the side rods to the grid terminals. The use of a silver-copper-alloy side rod has satisfactorily corrected this trouble. The silver-copper alloy has practically all the thermal conductivity of copper but is materially stiffened by the presence of a small amount of silver. It was found necessary to use a copper connector and radiator on the bottom of the control grid in order to permit heat flow from both side rods to the stem lead. An anode constructed with fins in a manner shown in Fig. 2 aids in reducing the anode temperature and, consequently, the operating temperature of the control grid. These fins increase the effective radiating area of the anode. Although the anode is made of carbonized nickel, great care must be exercised in order that no loose particles of carbon are left on the anode. These, under electrostatic fields, tend to migrate to the cathode and impair its emitting properties. Therefore, the anode must be brushed thoroughly to remove all loose carbon particles. The ends of the beam plates above the anode are closed over the grid-cathode structure in order to prevent stray electrons from escaping and striking the glass envelope.

Fig. 3 shows the stem used on both the RCA-829 and RCA-826. This type of stem permits of short, heavy, and low-loss leads which are quite essential at ultra-high-frequency. The 0.060-inch diameter tungsten

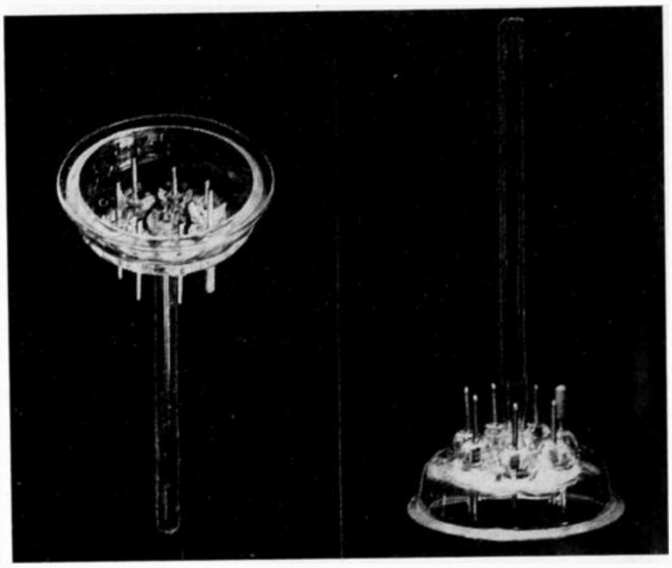


Fig. 3—Molded-flare stem as used on the RCA-826 and RCA-829.

leads of the stem afford an excellent means of transferring heat away from the control grid by conduction to the socket terminals.

The maximum ratings of the RCA-829 are given in Table I. Forced-air cooling must be used when the

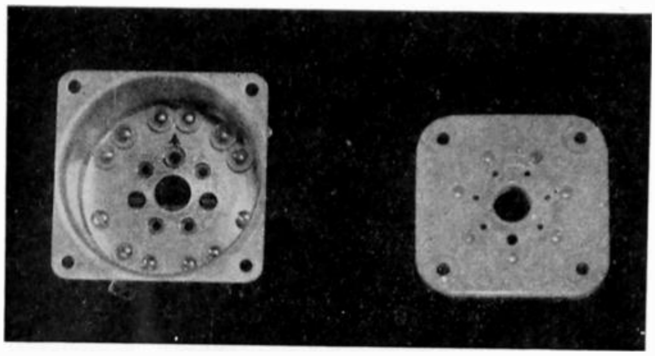
TABLE I

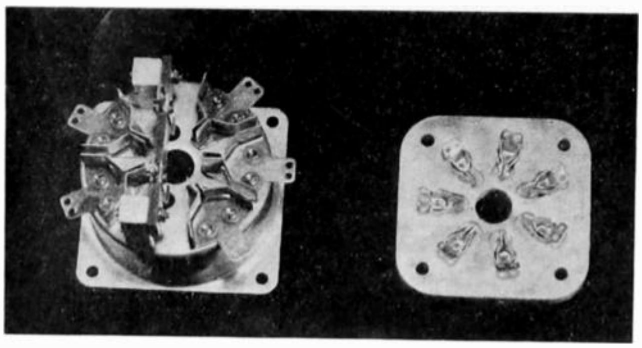
MAXIMUM RATINGS

	RCA-829	*RCA-815*	*RCA-826*
Class C Telephony (Carrier Condition)			
Direct plate voltage	425	325	800 volts
Direct screen voltage	225	200	— volts
Direct grid voltage	−175	−175	−500 volts
Direct plate current	212	125	95 milliamperes
Direct grid current	15	6	40 milliamperes
Plate input	90	40	75 watts
Screen input	7	2.7	— watts
Plate dissipation	28	13.5	40 watts
Class C Telegraphy			
Direct plate voltage	500	400	1000 volts
Direct screen voltage	225	200	— volts
Direct grid voltage	−175	−175	−500 volts
Direct plate current	240	150	125 milliamperes
Direct grid current	15	6	35 milliamperes
Plate input	120	60	125 watts
Screen input	7	4	— watts
Plate dissipation	40	20	60 watts

tube is operated at full rated input. With a circuit employing grid and plate transmission lines, a carrier power output of 53 watts (as measured in a lamp load) was obtained at 200 megacycles for a plate voltage of 425 volts and a current of 210 milliamperes. Measurements have indicated that the output circuit loss was approximately 10 watts. The circuit loss was estimated by attaching a thermocouple to the tube envelope opposite one anode. The temperature was noted under normal operating conditions in order to obtain an index of anode dissipation. The load was then removed and with the plate circuit tuned and with normal plate voltage maintained the input current was varied by adjusting the screen voltage until the same bulb temperature as previously noted was obtained. The plate

(a)





(b)

Fig. 4—(a) Top view of the UT-106 and UT-107 sockets.
(b) Bottom view of the UT-106 and UT-107 sockets.

circuit was then detuned and the same process repeated. The difference in the input power required to give the same envelope temperature with the unloaded circuit tuned and detuned gives an approximate measure of circuit loss. Since the plate-circuit losses are proportional to the square of the anode voltage, the plate-circuit efficiency should not rise as fast as predicted from voltage considerations. That such is the case has been verified in that little or no improvement in efficiency was obtained by increasing the anode operating voltage on the RCA-829 from 400 to 500 volts.

In order to realize the full capabilities of the RCA-829 it is necessary that all radio-frequency by-passing be as near the tube terminals as possible. It is desirable to have the by-pass condensers made an integral part of the socket assembly. Fig. 4 illustrates a socket (UT-107) that was developed commercially for this tube. For stable amplifier operation it is essential that the input circuit be well shielded from the output circuit. It is also important at ultra-high-frequencies to provide low-loss terminal connections to the tube.

RCA-815

The good performance and exceptional operating characteristics of the RCA-829 led to the consideration of a new tube that would be more adaptable to quantity production and that could be produced at a cost below that of the RCA-829. Such a tube should fulfill a need existing in the radio amateur field for a small tube suitable for 112-megacycle operation.

The RCA-815 is a twin beam tetrode designed for a maximum input of 60 watts at frequencies as high as 150 megacycles. (Table I.) The RCA-815 like the RCA-829 contains two beam power units except that the envelope is of soft glass instead of hard glass and the units are somewhat smaller than for the RCA-829 electrode assembly (Fig. 5). The total maximum plate dissipation is 20 watts instead of 40 watts but like its prototype it has a heater arrangement permitting either 12.6- or 6.3-volt operation. The two units may be operated either in parallel or push-pull as a modulator, oscillator, or radio-frequency amplifier. This tube is well suited for use as a frequency multiplier or driver for another RCA-815, or for the RCA-829.

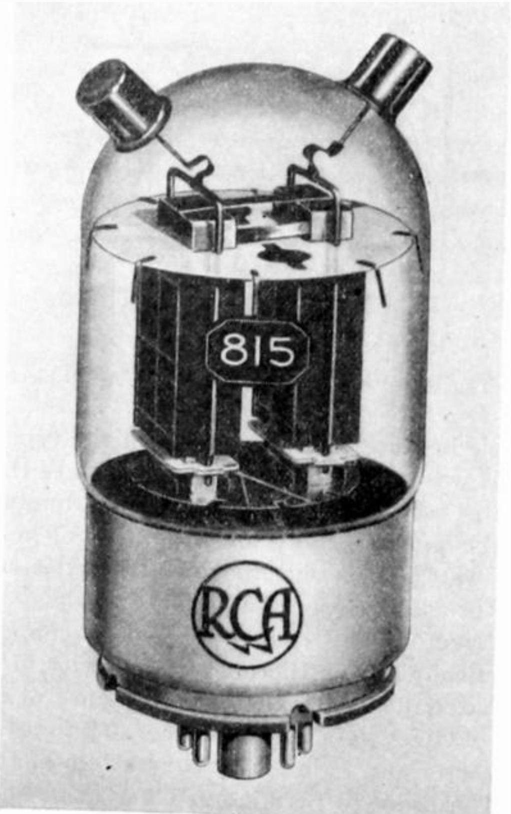


Fig. 5—A push-pull beam transmitting tube having oxide-coated cathodes and a total plate dissipation of 20 watts. It may be operated at full rating to frequencies as high as 150 megacycles.

The button stem shown in Fig. 6 is used to provide short lead length. The RCA-815 is made with a metal-shell micanol-wafer base which permits the use of conventional low-loss octal sockets since the maximum frequency is but 150 megacycles. Plate connections are made through conventional top caps. In order to minimize the inductance of the electrodes, use of a structure quite similar to that of the RCA-829 was obvious. The RCA-815 has substantially one half of the power-handling capabilities of the RCA-829.

In properly designed circuits a carrier power output of 30 watts may be obtained at a frequency of 150 megacycles. Forced-air cooling is not required.

RCA-826

In some fixed-frequency ultra-high-frequency transmitters the simplicity of triodes and their associated circuits offer an advantage over screen-grid types. The RCA-826 (Fig. 7) was developed to serve this field. It may be operated at maximum ratings at frequencies as high as 250 megacycles. (See Table I.)

This tube has a double-helix thoriated-tungsten filament, a zirconium-coated molybdenum anode, and a molybdenum grid. The anode has 8 fins to increase its total effective radiating area and is assembled by weld-

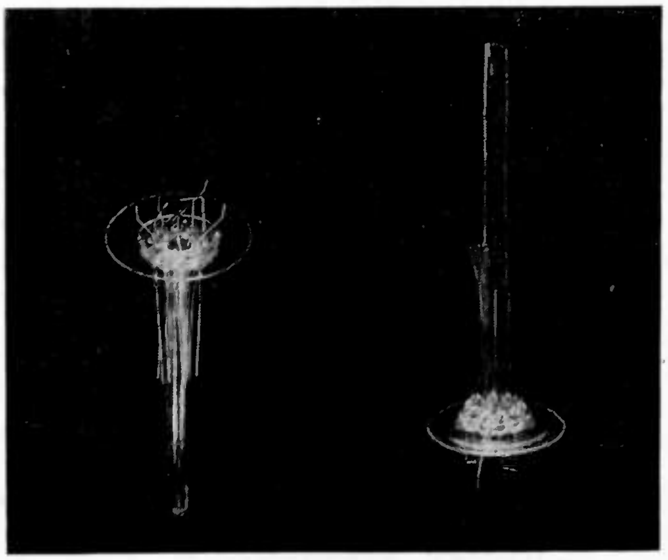


Fig. 6—Glass-button stem as used on the RCA-815.

ing 8 preformed molybdenum channels together. Sides of two adjacent channels form one fin, thus giving good heat conductivity from the barrel of the anode to the fin. The molybdenum is coated with zirconium for the twofold purpose of increasing the thermal emissivity over that of the conventional sandblasted molybdenum and to obtain the getter action of zirconium. Although this tube also has a conventional getter, developmental tubes without any getter other than the zirconium have been made, which were satisfactory on life test. This anode will safely withstand a momentary dissipation equal to several times rated plate input, without injuring the zirconium surface or impairing the quality of the tube. No insulation is used between the electrodes other than that of the molded-glass stem. Shields are placed on each of the structures to reduce stray electrons. These shields are mounted on the filament center support so that they will be at filament potential and will not collect any current during operation of the tube. A dual set of leads is brought

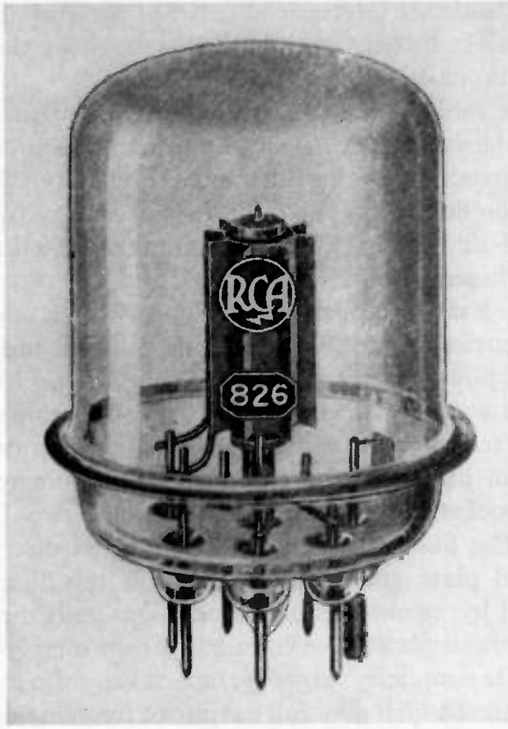


Fig. 7—A transmitting triode having a thoriated-tungsten filament and a total plate dissipation of 60 watts. This tube may be operated at full rating to frequencies as high as 250 megacycles.

out from the grid and plate, these leads ordinarily being arranged in parallel to reduce lead inductance and improve the ease of high-frequency operation. If desired, one set of leads may be used for neutralizing

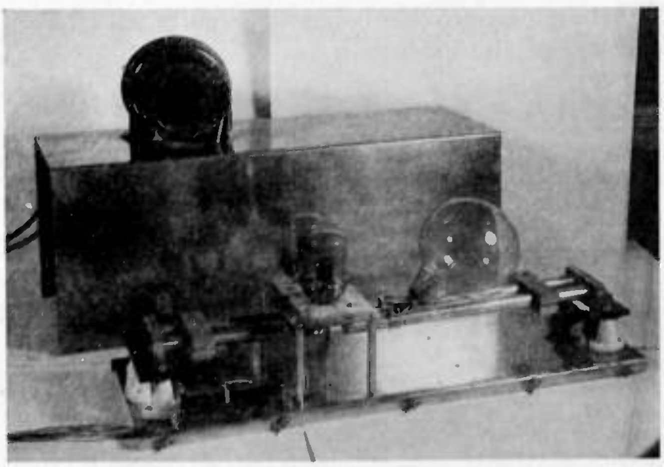


Fig. 8—A push-pull oscillator using RCA-826's.

and the other for handling power. This arrangement tends to minimize the trouble sometimes experienced in neutralizing where a considerable portion of the neutralizing circuit is common to the power-handling or controlling circuit. These dual leads are symmetrical about a diametrical plane through the filament center tap. This arrangement fulfills the frequent request for a right- and left-handed tube for push-pull operation in order to obtain short symmetrical leads. The filament leads are interspaced between the grid and anode leads since this arrangement helps to reduce the mutual inductance of these two latter leads.

The molded-glass flare used in this stem helps to provide short direct electrode leads. Due to the compact structure of the tube and the relatively high dissipation density of the bulb, it is necessary to provide forced-air cooling, when the tube is operated at maximum input rating.

Fig. 8 shows a push-pull oscillator which operates at frequencies from 140 to 200 megacycles and gives a useful power output as measured in a lamp load of 90 to 125 watts. This is essentially a conventional tuned-plate tuned-grid oscillator using quarter-wave transmission lines with no tuning of the filament circuit. The sockets used in this instance are two UT-106's with the filament center tap connected directly to a ground plate and with each side of the filament by-passed by means of small condensers built directly on the ground plate. The circuit is shown merely to illustrate its simplicity; however, special circuits have been developed which give full output to frequencies as high as 250 megacycles.

Application

In the application of these and other ultra-high-frequency tubes, it may be advisable to enumerate several precautions which are necessary to obtain satisfactory operation. These are:

1. Most push-pull amplifiers or oscillators are apt to have parallel oscillations unless special precautions are taken to provide balanced circuits with balanced excitation and loading.

2. Circuits used at ultra-high-frequencies should be silver-plated in order to minimize circuit losses.

3. Radiation should be minimized. This can be accomplished by a proper choice of shielding and the use of symmetrical circuits. Instances have been noted where full power output was not obtained because of resonance of the transmitter structure.

4. Adequate by-passing should be provided. The RCA UT-107, or its equivalent socket, should be used with the RCA-829. The UT-106 with postage-stamp mica condensers for by-passing should not be used with the 829 at ultra-high-frequencies.

5. Sockets should have large nonoxidizing contacts since a corroded or oxidized contact not only causes a reduction of output at ultra-high frequency but may cause severe circuit unbalance.

6. Since only one screen terminal is provided it is recommended that the individual units of either the RCA-815 or RCA-829 not be used as separate amplifiers, since it is quite difficult to determine the actual division of the screen input for other than push-pull or parallel arrangements of the units.

7. A vacuum-type incandescent lamp should be used in preference to the gaseous type for power-output measurements in order to avoid losses due to the ionization and arcing of the gas at high frequency. It may also be necessary to try a variety of sizes and structures of lamps before a satisfactory load is obtained.

Acknowledgment

In conclusion the author wishes to express his appreciation to his fellow members of the RCA Manufacturing Company, Inc., for their aid in the development of these tubes and in particular to Messrs. J. C. Hapgood, H. R. Nelson, C. F. Nesslage, and R. B. Vandegrift for their aid in the solution of many design and manufacturing problems.

Radiating System for 75-Megacycle Cone-of-Silence Marker*

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Summary—A brief description is given of a new 75-megacycle cone-of-silence marker radiating system developed for use on the Canadian airways. This system provides a sharper marker beam and reduces orientation error over previous marker radiating systems.

Accumulated civil-aviation experience with early types of cone-of-silence markers[1] has revealed certain deficiencies which pilots felt were compromising their usefulness somewhat. It was desirable to reduce these deficiencies by further development of the radiating system. A sharper vertical beam and less orientation effect with the typical horizontal linear marker-receiver antenna were the objectives sought.

A radiating system having the theoretical radiation characteristics shown in Fig. 1 appeared to offer the desired improvements. The mechanical form appeared to have acceptable cost and practicability characteristics. The authors undertook a collaborative development of the system in behalf of their respective organizations at the St. Hubert radio range station near Montreal in June, 1940. The development effort

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‡ Department of Transport, Ottawa, Ontario, Canada.
[1] Department of Commerce, Bureau of Air Commerce, Safety and Planning Division Report No. 16.

centered principally around the feeder problem. Following extended aerial field-intensity measurements for proof of performance, the St. Hubert marker was placed in service in September, 1940. Other similar installations are in service or under construction on the civil airways across Canada.

The finished form of this radiating system is pictured stereoscopically[2] in Fig. 2. It consists of four 2-layer turnstile radiators located at the corners of a 190-degree square above a 30-foot square reflecting screen. This screen is similar to that used with earlier-type markers. The lower dipole of each turnstile is 90 degrees above the screen, and the upper dipole 180 degrees above the lower. The turnstile form of radiator permits full metallic construction and in this case was built of solder-type copper pipe and standard fittings. All dipole currents are substantially equal. The current in the upper radiators is antiphased with respect to that in the lower to produce upwardly dominant radiation. This is accomplished by shunt-exciting the dipoles with vertical balanced feeders which are not transposed. All the dipoles of one orientation at the same level are cophased, but the phasing of the two orientations is in quadrature to produce a rotating field.

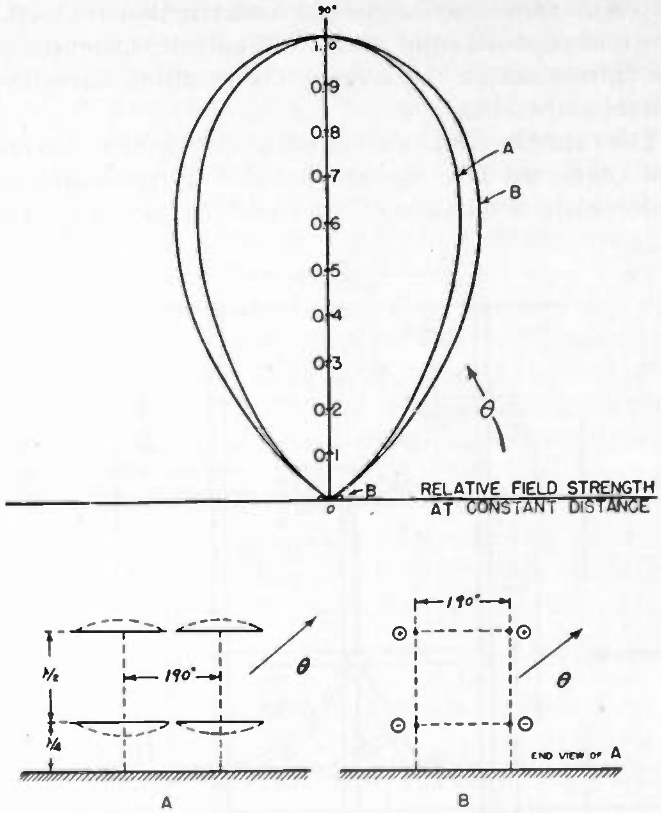

Fig. 1—Vertical radiation patterns for Department of Transport cone-of-silence marker antenna.

The radiators were composed of $\frac{7}{8}$-inch outside-diameter copper pipe with end plugs which had the same diameter. The measured resonant length of such radiators is 92 per per cent of theoretical. The central turnstile supports are 2-inch pipe, and the vertical feeders attached to the radiators are 3/8-inch outside diameter. To brace the turnstiles mutually and mechanically, a cross bracing of treated hardwood poles equipped with copper-tube ends entering

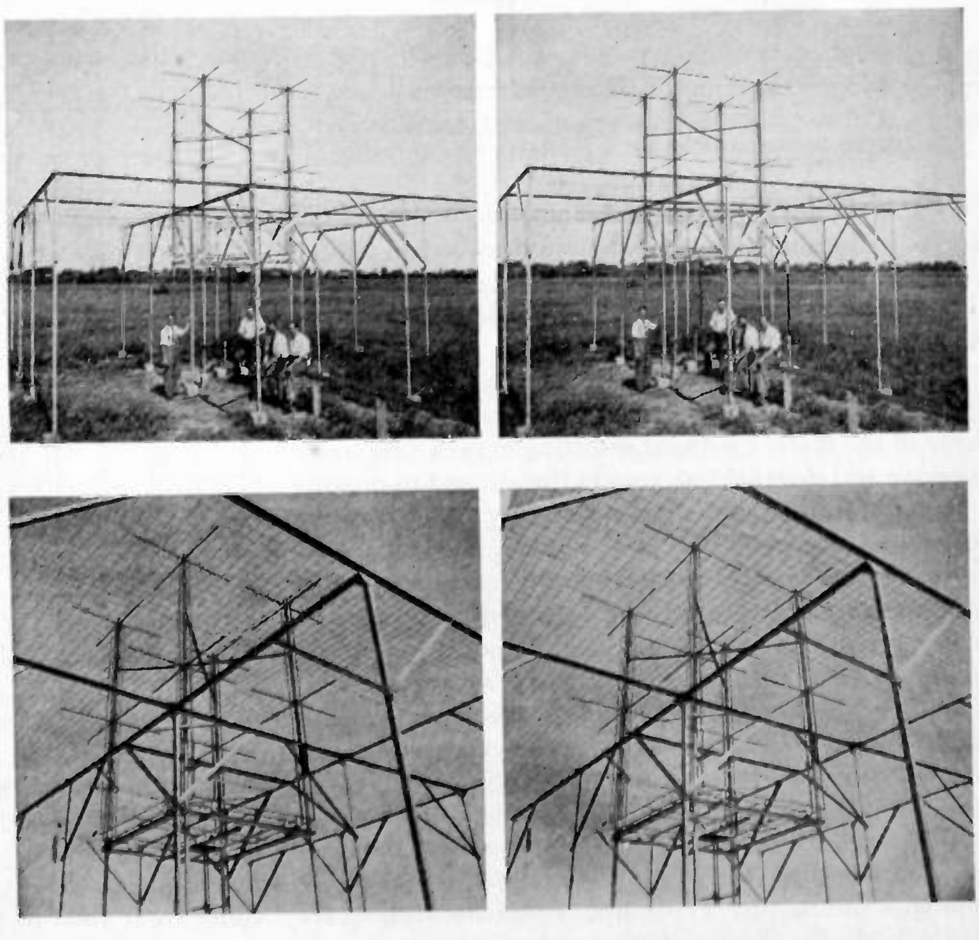

Fig. 2—Stereoscopic views of cone-of-silence marker antenna and counterpoise erected and tested at St. Hubert radio range station near Montreal.

[2] An unmounted pair of stereoscopic lenses provides the easiest method of viewing, it being necessary to adjust only the distance to the paper and the optical alignment to render the image in full natural perspective.

Without a stereoscope the following method may be used by most people with normal eyes and muscular accommodation: Use a piece of cardboard about 3 by 7 inches normal to the paper placed between the two images so that each eye sees only its own half of the stereogram. Adjust the eyes until the two images merge into one. Then as the eyes become accustomed to this accommodation gradually focus the eyes, increasing the eye distance from the paper if necessary, until the image is distinct, meanwhile holding the eye line parallel to the center line of the photographs. When in focus, with the accommodation adjusted to see the two images in parallel vision, three-dimensional effects will result.

It is necessary to have both images well illuminated and free of shadow when viewing. With brief practice the full advantage of stereo detail can be obtained.

the regular soldered copper-pipe construction are used. The four turnstile supports are individually fastened to the framework for the screen. The resulting assembly is rigid and stable.

The turnstile feeders are brought through the screen and connected into the open feeder system which is underneath. A diagram of the feeder circuits is shown in Fig. 3. Symmetry is depended upon to maintain uniformity of radiator currents in one orientation, and no attempt is made to match impedances except near the junction with the main balanced shielded transmission line from the transmitter.

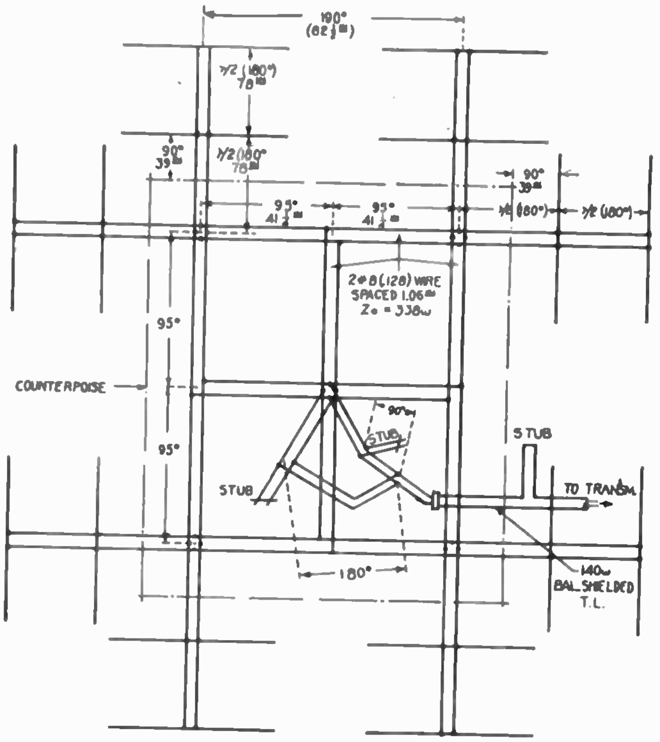

Fig. 3—Developed circuits for experimental cone-of-silence marker antenna system for St. Hubert range station.

During this development it was desirable to find a practical means for determining the correct performance of the marker without resorting in each case to expensive and slow field surveys in the air, and to provide means for adjusting the system quickly and accurately to obtain the requisite performance. This effort resulted in a method by which, after construction of the radiating and feeder system to strict dimensions, the correct phasing and power balance between the radiators of the two orientations was separately controllable by adjustment of the length only of two stub lines. In this regard, results exceeded expectations, for it made possible the checking or adjusting of an installation by the regular operating personnel.

The radiators of each orientation are brought back to common feeders where an approximate impedance match is effected by a stub line. From one stub to the main feeder, a length of 90 degrees is inserted, and from the other 180 degrees is used. The power in the two sets of radiators can be accurately equalized by adjusting the length of the stub in the 180-degree side. The phase difference between the two sets is adjustable by changing the length of the stub in the 90-degree side. The two adjustments thus give substantially independent control of these two effects which control the radiation pattern, and its rotation. To test this, a horizontal dipole with a central radio-frequency milliammeter is held aloft on a long pole above the center of the radiating system and its indicated current plotted as the dipole is rotated horizontally. An elongation of the plot in the direction of orientation of one of the sets of radiators indicates unequal power in the two halves of the system, and a correction in the stub length in the 180-degree feeder line is made. If the elongation of the plot occurs on a diagonal, the phase difference is off and a correction is made in the length of the stub in the 90-degree feeder. The final correct result is very nearly a circular plot.

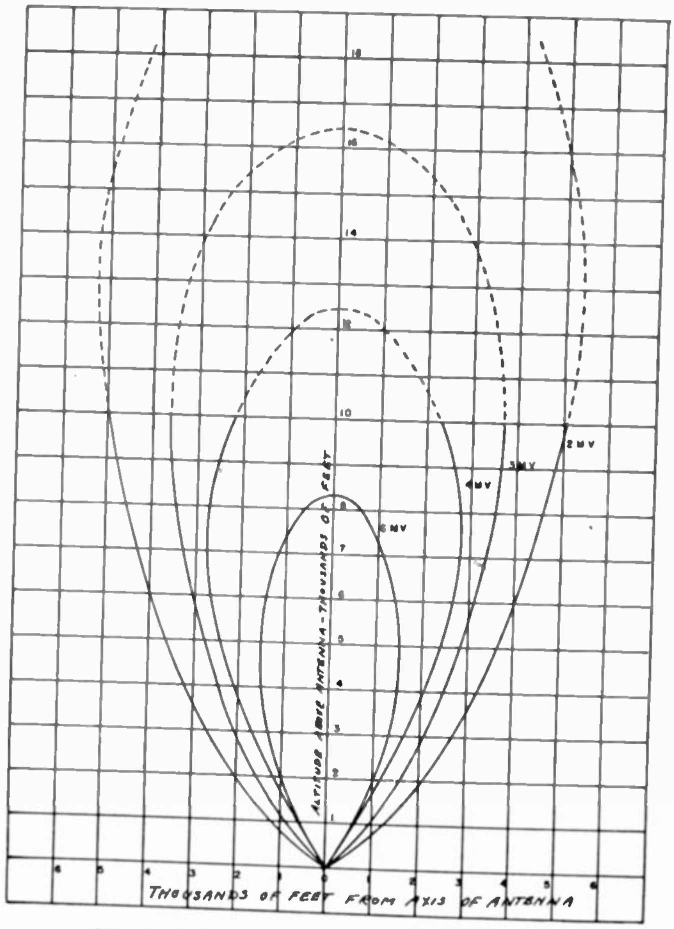

Fig. 4—Measured signal potentials induced in plane's marker antenna.

The actual pattern of the radiating system is shown in Fig. 4. This was deduced from flight measurements of signal potentials induced in the ship's receiving antenna taken on eight courses across the marker at altitudes from 1000 to 10,000 feet. As such they include the effect of directivity response of the plane's receiving antenna. The power finally used was approximately

3 watts and these measurements are on this value. Measurements made in direct comparison on previous types of marker antennas for orientation effects of the receiving antenna showed the present system to have reduced this to virtual nonexistence.

The development of this radiating system included various attempts to measure currents and potentials at various points in the radiating and feeder sections with the usual indifferent results. The ultimate test is the measurement of the radiation pattern. If this is correct, it is reasonable to tolerate some imperfections of terminations or line balance provided the conditions are stable. End effect between adjacent turnstile radiators undoubtedly causes a slight unbalance in the radiator impedance.

An initial source of electrical instability was found to be moisture on the end-seal insulators for the enclosed lines. It was necessary to protect these in a weatherproof housing. The only other source of electrical instability disclosed during its subsequent operation was ice accumulation on the radiators, but it is doubtful if any system would be immune from this. In the original design provisions were made for the use of asbestos-covered resistance wires to be run inside the copper pipes of the radiators and supports for sleet-melting, but so far this has not been undertaken. Housing the entire radiating system in a simple wood enclosure has been considered.

The operating characteristics of this radiating system, as determined from several months of use, appear to be suitable for marking by indicator lamp or aural means at high altitudes as well as providing an accurate fix at low altitudes for instrument approaches.

ACKNOWLEDGMENT

The authors wish to acknowledge the able assistance of Messrs. J. McKay and D. McDougal, of the Department of Transport staff, during this work.

A Quartz Plate with Coupled Liquid Column as a Variable Resonator*

FRANCIS E. FOX†, NONMEMBER, I.R.E., AND GEORGE D. ROCK†, NONMEMBER, I.R.E.

Summary—*The equivalent electrical constants of a combination resonator consisting of a piezoelectric X-cut quartz plate and a closely coupled liquid column have been experimentally investigated. The equivalent electrical constants of such a resonator were greatly changed and the resonant or oscillating frequency was shifted over a wide range (2.2 to 3.0 megacycles) by adjusting the length of the liquid column. In particular, the sharpness of resonance was found to vary from 400 for the plate, cemented to a holder but without the liquid column, to 4000 for the combination resonator.*

IT IS a well-known fact that a quartz plate, such as used in the stabilization of electrical oscillating circuits or in crystal filters, may have its oscillating or resonating frequency shifted slightly when used in a holder with an air gap between the plate and one of the electrodes. The frequency variation obtainable by adjusting the air gap is, however, very small. In a former paper upon ultrasonic interferometry in liquid media,[1] one of the present authors discussed briefly in a footnote the effect of a liquid column coupled to a quartz plate in changing the resonating frequency of the combined resonating system composed of the plate plus the liquid column. A liquid column coupled to the quartz might, therefore, be utilized to produce a relatively large shift in the resonating frequency of the system. Since this is an obviously desirable feature in some types of electrical equipment, the present paper presents the results of investigations made to determine the modifications that can be produced in the equivalent electrical constants of a quartz resonator or oscillator by means of a closely coupled liquid column.

In the theory of the acoustic interferometer as developed by Hubbard,[2,3] and applied to the case of liquid media by one of the authors,[1] the vibrating quartz is assumed to act as a source of plane sound waves. These are reflected and react upon the quartz. It is possible to express this reaction in terms of "equivalent" electrical resistance and reactance, which, together with the equivalent electrical constants of the quartz plate, determine the electrical characteristics of the combined "plate-plus-column" resonating circuit. This will not be treated in detail in the present paper. Here it is proposed to adopt the viewpoint that the effect of the liquid column is simply to change or modify the equivalent electrical constants of the vibrating system of "plate-plus-column" as a single unit. Although it is realized that it is perhaps an oversimplification to assume that these constants represent a simple series circuit, it is felt that the coefficients are constant over a sufficiently wide frequency range to justify the method of measurement and that this offers

* Decimal classification: 537.65. Original manuscript received by the Institute, April 25, 1941; revised manuscript received, October 7, 1941.

† Catholic University of America, Washington, D. C.

[1] F. E. Fox, "Ultrasonic interferometry for liquid media," *Phys. Rev.*, vol. 52, pp. 973–981; November, 1937.

[2] J. C. Hubbard, "The acoustic resonator interferometer: I. The acoustic system and its equivalent network," *Phys. Rev.*, vol. 38, pp. 1011–1019; September, 1931.

[3] J. C. Hubbard, "The acoustic resonator interferometer: II. Ultrasonic velocity and absorption in gases," *Phys. Rev.*, vol. 41, pp. 523–535; August, 1932.

a useful and direct method of investigating the observed phenomena. Accordingly, we chose to follow Dye's method of determining the equivalent series resistance (R), inductance (L), and capacitance (C) of such a combination resonator. The circuit was the usual one with an inductance, thermogalvanometer, and a variable standard condenser in series, with the resonator connected across the variable condenser. L_1, C_1, and R_1 are the electrical constants of this pickup circuit, in which R_1 is the total resistance of the circuit, including that of the thermogalvanometer. When C_1 is adjusted so that the L_1C_1 circuit has the same resonant frequency as that of the mechanical resonator the curve of the current (i) plotted against the frequency, rises as the frequency approaches the resonant frequency of the L_1C_1 circuit, dips sharply to form the crevasse at the resonant frequency of the mechanical resonator, rises again and then falls off more slowly as the frequency increases beyond the neighborhood of resonance for the L_1C_1 circuit. When the two resonant frequencies coincide, the crevasse divides the L_1C_1 resonance curve symmetrically. One usually discusses instead the current (i or i^2) the ratio $\sigma = i/I$, where I is the maximum current at resonance in the L_1C_1 circuit without the mechanical resonator. One needs only to determine the minimum current ratio (i.e., σ_m) which occurs at the bottom of the crevasse, the sharpness of resonance Q of the mechanical resonator ($Q = \phi^{-1}$ where $\phi = \delta/\pi = RC\omega_0$, δ being the usual logarithmic decrement and $\omega_0 = 2\pi f_0$ where f_0 is the resonance frequency of the mechanical resonator), and the resonance frequency in order to complete the analysis of the resonator constants, assuming that the L_1, C_1, and R_1 are already known.

We give the results of Dye's analysis of the symmetrical crevasse.[4] The terms referring to the air gap are eliminated since none was used in the present work.

$$R = \sigma_m[(1 - \sigma_m)\omega_0^2 C_1^2 R_1]^{-1} \tag{1}$$

$$C = \phi(R\omega_0)^{-1} = (CR\omega_0)^{-1} \tag{2}$$

$$L = (C\omega_0^2)^{-1}. \tag{3}$$

The determination of ϕ or Q is the only one that presents difficulty. If great accuracy is desired the best method consists in plotting σ or σ^2 against the frequency in the region of the crevasse. The width (Δf) of the crevasse between two equal values of σ^2 is then taken from the crevasse curve for various measurements of σ^2 not too near the minimum. Then

$$\phi = \frac{\Delta f}{f_0} \sigma_m \sqrt{\frac{1 - \sigma^2}{\sigma^2 - \sigma_m^2}}. \tag{4a}$$

The plot of Δf against $\sqrt{(\sigma^2 - \sigma_m^2)(1 - \sigma^2)^{-1}}$ (abscissa) should be a straight line of slope M which determines ϕ. Thus

$$\phi = M\sigma_m f_0^{-1}. \tag{4b}$$

One may read directly from the crevasse curve the frequency difference (Δf) for $\sigma^2 = (1 + \sigma_m^2)/2$ without plotting the straight line. This is M, since the radical reduces to 1 at this value of σ^2. When making many readings it is convenient to determine first the σ_m^2, and then measure experimentally the frequency change that occurs between the two values of $\sigma^2 = (1 + \sigma_m^2)/2$ on opposite sides of the crevasse. This was the method used in all but a few measurements.

Apparatus

An X-cut quartz plate having its thickness-vibration frequency close to 2.5 megacycles was cemented to a brass disk, over an opening slightly smaller than the plate. The upper face was sputtered with platinum, the film forming a continuous connection to the brass disk. A film was also sputtered on the plate face showing through the opening, but a wax coating about a millimeter wide was placed around the edge of the quartz before sputtering. This was afterwards removed, leaving an insulating space between the edges

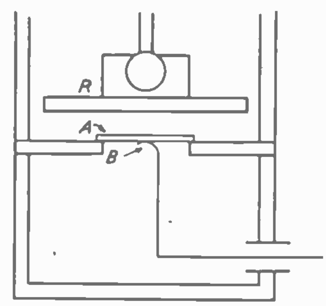

Fig. 1—Combination "quartz-plate and liquid-column" resonator.

of the film on the quartz and the supporting plate. The damping of a quartz plate thus mounted is, of course, greater than for the same plate vibrating freely in air, but this seems to be of no great importance since we were primarily interested in modifications that can be produced by an associated liquid column, and the presence of the liquid would in any case increase the damping over that of an unmounted plate vibrating freely in air. The disk was used as the base of a container for a liquid in which a metal reflector, attached to a micrometer screw, could be moved to adjust the length of the liquid column between the quartz and the reflector shown in Fig. 1. There is a decided advantage in using platinum films as electrodes sputtered directly on the quartz. The air gap is thus eliminated and the analysis yielding the equivalent electrical constants of the quartz plate correspondingly simplified.

The oscillator to which the L_1C_1 circuit is very loosely coupled has been described previously.[5] A master oscillator with electrical band spreading for fine frequency control feeds through a buffer amplifier into

[4] P. Vigoreux, "Quartz Oscillators," His Majesty's Stationary Office, Longon, England, 1939.

[5] F. E. Fox and G. D. Rock, "Ultrasonic absorption in water," *Jour. Acous. Soc. Amer.*, vol. 12, pp. 505–510; April, 1941.

a series of frequency-doubler tubes and then into a power amplifier. A vacuum-tube voltmeter in conjunction with the final stage allows one to monitor the constancy of the output, and insure the constancy of the voltage induced in the inductance of the pickup circuit. The master oscillator is continuously variable from 1 to 2 megacycles. In practice only the band-spread condenser dial was used to vary the frequency in any determination of ϕ. A secondary frequency standard consisting of a quartz oscillator at 1 megacycle is adjusted so that its fifth harmonic is at zero beat with the 5-megacycle carrier of the Bureau of Standards Station WWV. This is used to control by a selection switch any or several of a series of multivibrators at 100, 25, and 10 kilocycles. We could choose a frequency at, for example, 2.5 megacycles and adjust the main dial of the oscillator until zero-beat conditions prevailed with the vernier dial set at half value (both dials were of the National type having 500 large divisions for 180-degree rotation of condenser plates) and then calibrate the vernier dial by using the 10-kilocycle spotting points. This was done for main dial settings giving 2.0, 2.1, . . . 3.0 megacycles. A switch was provided in one lead to the mechanical resonator, so that this could be disconnected when determining the constants of the L_1C_1 circuit. Once the apparatus had been assembled, nothing was changed or manipulated but the switch and the reflector position of the acoustic interferometer, in order to insure the constancy of distributed capacitance in the wiring, etc. It was found useful, when the frequency variation to be measured was small, to employ a frequency bridge to measure the beat frequency between a signal from the secondary standard and the oscillator. A radio receiver in the room was kept tuned to the constant-frequency signal and the Wien frequency bridge inserted in the phone leads.

Experimental

(1) The first measurements were made to determine the constants of the pickup circuit. L_1 was determined from observations of the resonant frequencies obtained with different values of C_1 as read on the precision air condenser. This gives also the distributed capacitance C_0 of the coil and associated apparatus. The sharpness of resonance was measured in the usual manner in which the frequency is varied to cause σ^2 to fall to one half. This yields the total resistance in the L_1C_1 circuit.

(2) The effect of switching the quartz plate and interferometer into the circuit was next noted. The capacitance of the electrodes of the quartz plate and the distributed capacitance caused by the interferometer and its leads, together with any effective series resistance due to leakage across insulating gaps, are purely electrical constants that may be determined at frequencies removed from those at which there is any resonant vibration of the quartz. The constants were determined by noting the capacitance change of the precision condenser needed to restore resonance in the L_1C_1 circuit, and the change (very small) in the current at resonance. Thus the total or effective inductance, capacitance, and resistance for any given experimental conditions are determined if the setting of the precision condenser is known.

(3) The equivalent inductance, capacitance, and resistance of the mounted plate without liquid loading was measured next. Knowing that the resonant frequency of the quartz was near 2.5 megacycles, we adjusted the main dial so that the oscillator was at zero-beat with the 2.5-megacycle frequency from the secondary standard, the vernier dial being at 250. The L_1C_1 circuit was tuned to resonance at 2.5 megacycles. The switch to the resonator was next closed, after which the frequency was adjusted with the vernier dial until the current was a minimum at the bottom of the crevasse. The interferometer was then disconnected and the precision condenser tuned to place the L_1C_1 circuit at resonance. From this and the data of (2) the precision condenser could be reset, after switching the resonator back into the circuit, so that the resonant frequency of the L_1C_1 circuit occurs at the bottom of the crevasse; that is, the resonant frequency of the quartz and the L_1C_1 circuit coincide. The current observed for various settings of the vernier dial enable one to plot σ^2 versus frequency over the crevasse. The curve is given in Fig. 2. From this, the frequency width (Δf) of the crevasse for any value of σ^2 is measured and ϕ determined as indicated above. This, in conjunction with σ_m^2 and the L_1C_1 and R_1 values enable one to calculate L, C, and R for the quartz plate as mounted.

(4) We were now ready to investigate the modifications of L, C, and R that a liquid column could effect. Water was placed in the interferometer and the reflector advanced to make firm contact with the quartz. The oscillator was set at 2.5 megacycles as in (3) and the L_1C_1 circuit adjusted to resonance at this frequency. This is done by first disconnecting the resonator and adjusting to resonance, after which the resonator is again connected and the precision-condenser capacitance diminished as determined in (2). The reflector is then withdrawn until the liquid column plus the quartz plate is again in resonance, as indicated by a sharp dip in the current to a minimum value σ_m^2. This first dip occurs when the liquid column is approximately $\lambda/4$, where λ is the wavelength in water of the sound generated by the vibrating quartz. Successive dips occur at intervals of $\lambda/2$. With the reflector in this "first dip" position, the crystal plus the column is a new combination resonator with its resonating frequency at 2.5 megacycles. As the frequency is varied with the vernier dial, the characteristic crevasse is obtained for σ^2 plotted against f. This is analyzed as in (4) to obtain σ_m^2 and ϕ, and through these the L, R, and C for the combination resonator.

(5) The next step was to shift the frequency of the

combination resonator by tuning the liquid column, as suggested before. The oscillator and L_1C_1 circuit were set at 2.4 megacycles. It was possible to secure a pronounced current dip by adjusting the reflector, although the minimum was much greater than at 2.5

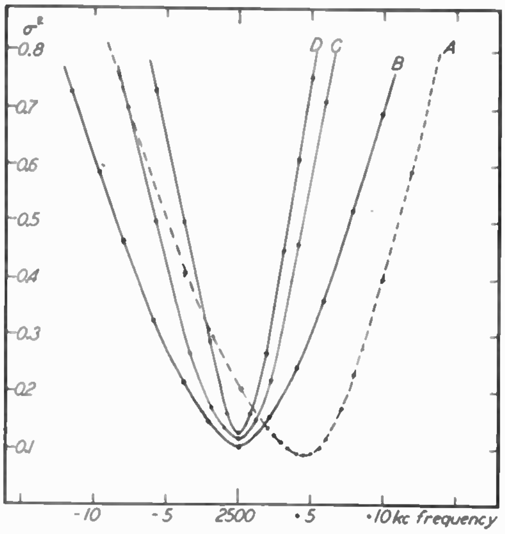

Fig. 2—Crevasse curves for quartz plate, and for combination resonator with various column lengths. *A*, mounted plate; *B*, *C*, and *D* are for the 1st, 10th, and 20th dips, respectively.

megacycles. Having set the reflector for minimum current (σ_m^2) the frequency was again varied and the crevasse plotted. Again σ_m and ϕ were obtained and the *R*, *L*, and *C* determined for the new combination resonator. This was repeated at 2.6 and attempted at 2.3 and 2.7 megacycles but difficulty was found in determining ϕ at these latter frequencies.

(6) A series of determinations of σ_m and ϕ were made at 2.5 megacycles with the reflector set in the 10, 20, . . . 200th dip. It was seen that the sharpness of reso-

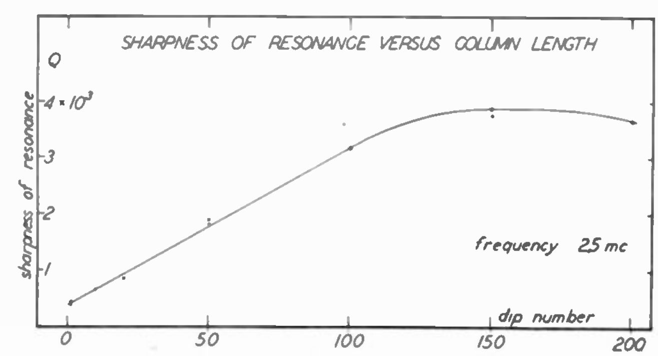

Fig. 3—Sharpness of resonance (*Q*) as a function of the length of the liquid column.

nance (*Q*) is greatly increased by increasing the length of the resonating liquid column, reaching 4000 at the 150th dip as compared with 418 for the mounted quartz plate without the liquid column. Fig. 2 gives the crevasse curves for the mounted plate, and for the "plate-plus-column" resonator with the reflector set at the 1st, 10th, and 20th dip. The values of *Q* are plotted against the dip number in Fig. 3. The decrease in *Q* for the 200th dip is probably due to the absorption of sound waves in the liquid. In Fig. 4 the values of *R*, *L*, and *C* are plotted as a function of the dip number.

(7) Although attempts to secure the characteristic crevasse at 2.3 and 2.7 megacycles had been unsuccessful at the first dip, it was found possible to extend the measurements, when the 100th dip was used, from 2.2

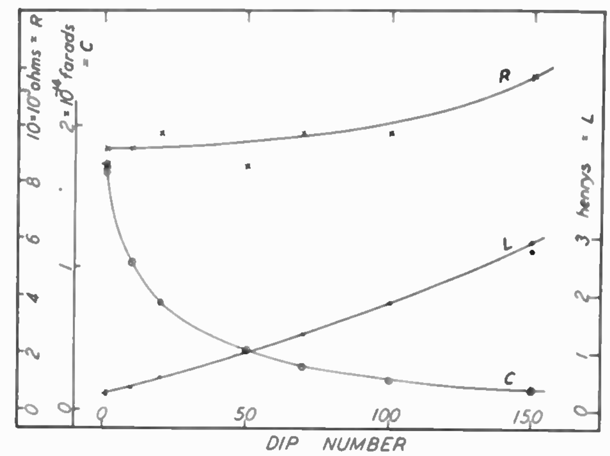

Fig. 4—Variation of inductance (*L*), capacitance (*C*), and resistance (*R*) as a function of the length of the liquid column.

to 3.0 megacycles. Although the variation of *Q* in these measurements as shown by the circles on Fig. 5 is not a simple function of the frequency shift, the equivalent series resistance, capacitance, and inductance of the combination resonator change in quite a regular manner as the combination is adjusted to resonate at

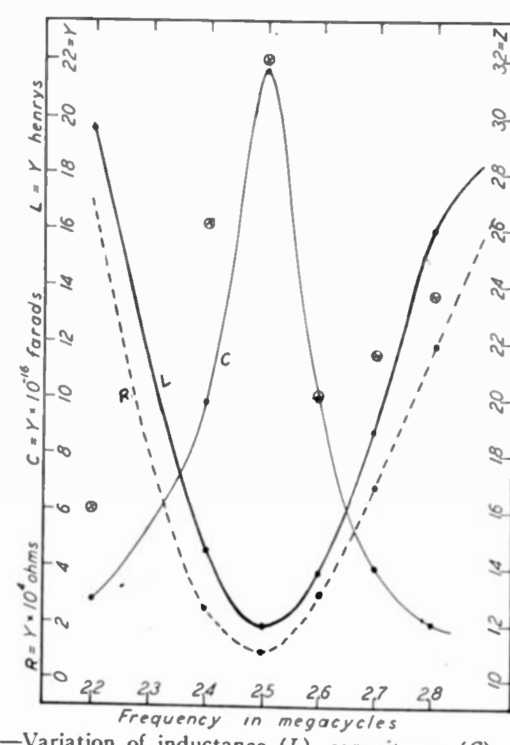

Fig. 5—Variation of inductance (*L*), capacitance (*C*), resistance (*R*), and sharpness of resonance (*Q*) as a function of frequency. Reflector set for the 100th dip.

frequencies removed from the resonant frequency (f_0) for the unloaded resonator. This variation is given in Fig. 5. No attempt was made to measure the sharpness of resonance at frequencies beyond this range since the dip or crevasse was becoming quite small. The dip, however, was still detectable at 2 and 4 megacycles.

It should be pointed out here that there is an assumption involved in the methods of measuring ϕ; namely, that one may treat the combination as a resonating unit determined by a single set of equivalent electrical constants. That this assumption is not strictly true is indicated by an analysis of the shape of the crevasse curve. A plot of Δf against $\sqrt{(\sigma^2-\sigma_m^2)(1-\sigma^2)^{-1}}$ should, if the assumption is entirely valid, be a straight line passing through the origin. Actually the points plotted in this fashion corresponding to current values near the bottom of a crevasse do not always fall on the straight line determined by the origin and current values farther removed from the minimum (i.e., $\sigma^2>(1+\sigma_m^2)/3$). It seems well to point out that the method of determining ϕ by measuring the crevasse width at $\sigma^2=(1+\sigma_m^2)/2$ corresponds to a definition of sharpness of resonance Q of the system under investigation, and is one that coincides most closely with the method of measuring the Q of a simple electrical circuit.

Tables I and II contain the data, observed and calculated, from which curves of Figs. 3, 4, and 5 are

TABLE I

NOTE: $R\times10^3$ = ohms; L = henrys; $C\times10^{-15}$ = farads

Frequency Megacycles	Dip Number	σ_m^2	$\phi\times10^3$	Q	R	C	L
2.5045		0.09	2.39	418	7.55	20.1	0.20
2.5000	1	0.11	2.40	417	9.19	16.6	0.24
2.5000	1	0.10	2.30	435	8.55	17.2	0.24
2.5000	10	0.11	1.49	670	9.19	10.3	0.39
2.5000	20	0.12	1.14	876	9.79	7.4	0.55
2.5000	50	0.10	0.53*	1880	8.55	4.0	1.02
2.5000	50	0.10	0.54	1860	8.55	4.0	1.01
2.5000	70	0.12	0.47	2125	9.79	3.06	1.32
2.5000	100	0.12	0.33	3200	9.79	2.16	1.88
2.5000	150	0.15	0.26	3910	11.70	1.39	2.92
2.5000	150	0.15	0.27	3730	11.70	1.46	2.78
2.5000	200	0.19	0.27	3680	14.30	1.21	3.35

* Measured at values of σ^2 other than $(1+\sigma_m^2)/2$.

drawn, together with that of other measurements made during this investigation. Values of ϕ are given as determined by Δf $\sigma^2=(1+\sigma_m^2)/2$; where other values were used, either as a check or because of the difficulty of measuring crevasse widths when σ_m^2 was itself close to 1, this is indicated by an asterisk.

RESULTS

It is seen that the resonance (or oscillation) frequency of a combination consisting of a quartz plate and a coupled liquid column may be varied over a wide range by properly tuning the liquid column, and that the sharpness of resonance of such a resonator may be increased many times over that of an unloaded quartz plate mounted on a support. The associated liquid column can be utilized to produce a wide variation in the equivalent electrical constants of such a

TABLE II

NOTE: $R\times10^3$ = ohms; L = henrys; $C\times10^{-15}$ = farads

Frequency Megacycles	Dip Number	σ_m^2	$\phi\times10^3$	Q	R	C	L
2.4	1	0.225	3.89	257	15.4	16.75	0.26
2.5	1	0.11	2.40	417	9.19	16.6	0.24
2.5	1	0.10	2.30	435	8.55	17.2	0.24
2.6	1	0.27	8.16*	122	21.7	23.2	0.16
2.4	10	0.31	2.46	406	21.4	7.63	0.58
2.5	10	0.11	1.49	670	9.19	10.3	0.40
2.6	10	0.28	2.55	392	22.6	6.9	0.54
2.2	100	0.86	0.62	1600	169.0	0.27	19.5
2.4	100	0.36	0.38	2630	25.5	0.99	4.45
2.5	100	0.12	0.33	3200	9.79	2.2	1.88
2.6	100	0.36	0.50	2020	30.1	1.0	3.74
2.7	100	0.58	0.46	2160	69.2	0.40	8.8
2.8	100	0.70	0.42	2370	119.0	0.2	16.0
3.0	100	0.80	0.61*	1640	228.0	0.14	19.8
3.0	100	0.80	0.605*	1650	228.0	0.14	20.0

* Measured at values of σ^2 other than $(1+\sigma_m^2)/2$.

resonator, for example, an inductance from 0.24 to 20 henrys. This might be applied to such problems as stabilization of oscillating and resonating circuits of which one can adjust the characteristic frequencies over a wide range. The dependence of the sharpness of resonance upon the length of the resonating column suggests its use in filters capable of regulating a variable-width frequency band, and so on. There is no doubt that by the use of a liquid such as mercury instead of water the sharpness of resonance could be further increased, and that this would permit an even greater shift in the resonant frequency of the combination resonator.

CONCLUSION

A detailed explanation of the observed phenomena has not been attempted in the present article for it is felt that this discussion belongs primarily to the field of acoustics, more exactly to ultrasonic interferometry. The results, however, seem important from an electrical engineering viewpoint and this has determined the presentation given here.

ACKNOWLEDGMENT

The authors wish to express their gratitude to Dr. Karl Herzfeld for aid given often during discussions on the problem.

Common-Channel Interference Between Two Frequency-Modulated Signals*

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Summary—*Two frequency-modulated signals received in the same frequency band produce an output from the receiver which is simply a combination of both signals and a beat note whose frequency is modulated in accordance with both signals. The difference of signal strength required for the reduction of crosstalk from the weaker signal is less in frequency modulation than in amplitude modulation, but is the same for different bandwidths of frequency modulation. The required difference can be further reduced by the use of a limiter in the receiver. The beatnote interference remains as background noise fluctuating with the modulation of both signals. This noise is reduced by wide-band frequency modulation. Simple expressions for the detector output in several cases enable the identification of frequency effects which are unavoidably detected as distinguished from amplitude effects which can be removed by a limiter. Common-channel interference is readily tested by oscilloscope patterns. These show the normal operation with or without a limiter, and also the effects of departure from the normal, such as detuning.*

I. Introduction

COMMON-CHANNEL interference is caused by the reception of an undesired signal in the same frequency channel as the desired signal. Such interference is inherently independent of frequency selectivity in the receiver. In amplitude-modulation systems it is independent also of other properties of the receiver, such as the difference between linear and square-law detectors. In frequency-modulation systems, however, such interference is determined by the properties of the receiver and the bandwidth of frequency modulation.

The receiver has a frequency detector which is balanced against amplitude modulation at the unmodulated-carrier frequency of the desired signal. There may or may not be a carrier-amplitude limiter preceding this detector. In response to only one signal, the output of the frequency detector is proportional to the frequency modulation. This is a property of any one of several types of idealized frequency detectors with linear or square-law slope filters and rectifiers.

The amount of common-channel crosstalk interference depends on whether the frequency detector has a linear or a square-law rectifier, unless a perfect limiter is assumed. It is least with a perfect limiter and greatest with square-law rectifiers. It is unaffected by the characteristics of the frequency-modulation system, such as the bandwidth of modulation.

The beatnote interference has the unusual characteristic of simultaneous amplitude and frequency modulation. Its peak amplitude is dependent on the receiver properties but more significantly is affected by some properties of the frequency-modulation system. It is reduced by increasing the bandwidth of frequency modulation. It is further reduced by pre-emphasis and restoration of the higher frequencies of the modulating signal, which incidentally requires a restoring filter after the detector in the receiver.

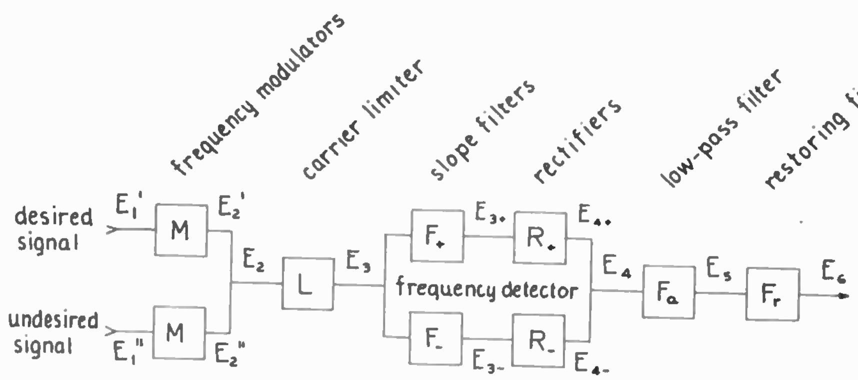

Fig. 1—Frequency-modulation system with interference between two signals in the same frequency band.

The relation between the amplitude and the frequency of the beatnote interference is shown directly by "conical patterns" on the oscilloscope. These are produced by tracing the two-signal output vertically against the difference between the respective modulating signals as the horizontal sweep. This choice of sweep voltage causes the horizontal displacement to be proportional to the beat frequency, so these patterns show directly the effect of any frequency-selective filters following the detector.

All of the relations to be described are based on simple and direct theoretical derivations with the aid of the zero-frequency-carrier concept. There is no assumption as to the waveform of the modulating signals.

The parts of the frequency-modulation system which are essential in the study of common-channel interference are shown in Fig. 1. The audio-frequency modulating voltages are respectively E_1' and E_1'' for the desired and undesired signals. The corresponding

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frequency-modulated-carrier signals are E_2' and E_2''. The composite signal which reaches the receiver is E_2. This may or may not be subjected to the action of a carrier-amplitude limiter before reaching the detector as E_3. The balanced detector comprises a pair of slope filters of opposite slope and a pair of rectifiers. The differential output of the rectifiers is the composite detected signal E_4, including both the desired signal and the interference from the undesired signal.

II. Idealized Balanced Frequency Detectors

An ideal detector for frequency modulation is one which not only reproduces the waveform of frequency modulation but also is unresponsive to amplitude modulation. The latter requirement is partially fulfilled in the balanced frequency detector. This comprises two frequency detectors with a differential output circuit. They convert the frequency modulation into amplitude modulation of opposite polarities. In their differential output appears the signal corresponding to the frequency modulation, but any signal corresponding to original amplitude modulation tends to cancel out. Complete avoidance of response to amplitude modulation would require a perfect limiter preceding the detector.

The elements of the balanced frequency detector are shown in Fig. 1. The slope filters of the two sides have equal response at the center frequency and have opposite slope. The rectifiers are alike but oppositely coupled to the output circuit. These relations assure the nearest approach to cancellation of any output representing incidental amplitude modulation.

The properties of linear slope filters are shown in Fig. 2(a), relative to the unmodulated-carrier frequency in the center. The filter factors are denoted F_+ and F_-. The intercepts at $\pm f_c$ are located arbitrarily for present purposes, but are preferably near the edges of the pass band in practical applications. This location makes the balance least critical and gives sufficient operating range with linear rectifiers.

The use of the linear slope filters with linear rectifiers gives the characteristics of Fig. 2(b).[1] The linear-rectifier properties give rectified voltages equal in magnitude to the envelopes of the voltages from the slope filters. The output of the balanced detector is the difference of the two rectified voltages. In this case, the differential output is proportional to the frequency modulation only between the intercept frequencies $\pm f_c$, as shown in Fig. 2(c).

Square-law rectifiers instead of linear rectifiers modify the frequency characteristics to those of Fig. 2(d). The square-law properties give rectified voltages equal to the square of the magnitudes of the envelopes of the voltages from the slope filters. The differential output, as shown in Fig. 2(e), is proportional to the frequency deviation over an unlimited range. The curvature of the square-law rectifiers cancels out. This type of balanced frequency detector is ideal for theoretical studies, because square-law rectification is most easily formulated in mathematical terms.

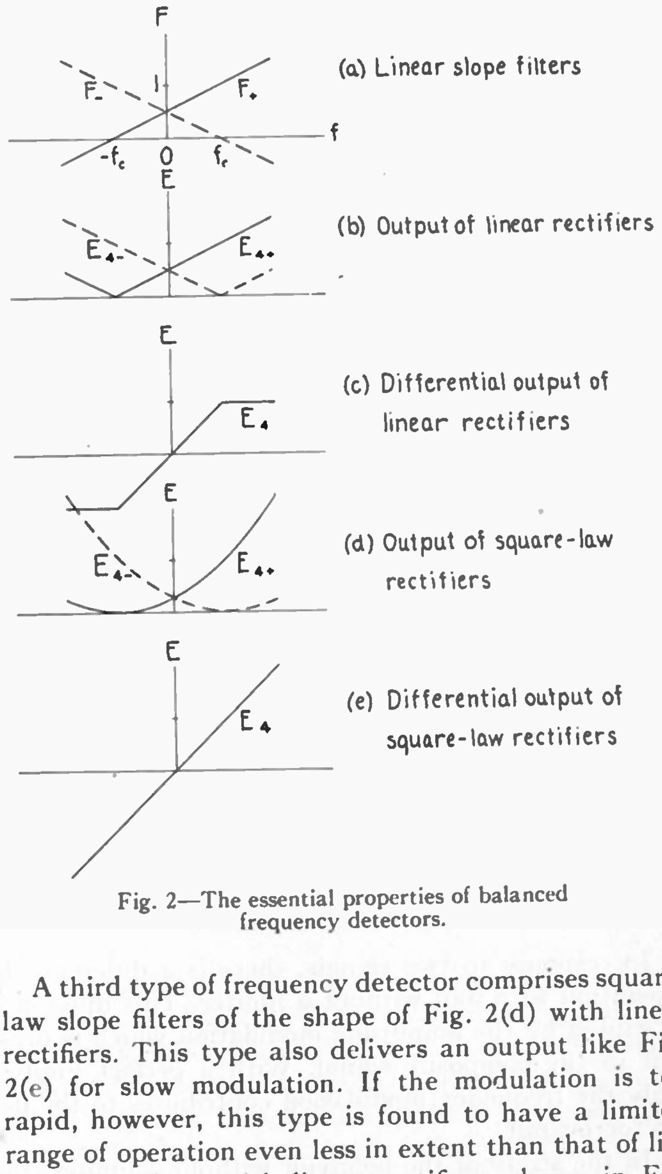

Fig. 2—The essential properties of balanced frequency detectors.

A third type of frequency detector comprises square-law slope filters of the shape of Fig. 2(d) with linear rectifiers. This type also delivers an output like Fig. 2(e) for slow modulation. If the modulation is too rapid, however, this type is found to have a limited range of operation even less in extent than that of linear slope filters with linear rectifiers, shown in Fig. 2(c).

The first type, with linear slope filters and linear rectifiers, is the only one which tolerates a departure from balance without causing distortion of the signal. The other types rely on the balance to cancel the distortion introduced by the square-law characteristics of the individual rectifiers or slope filters.

Linear rectifiers have a practical advantage over square-law rectifiers in that the output signal amplitude varies only half as much with input amplitude. Also the relative response to the amplitude modulation in the composite signal is found to be only half as great with linear rectifiers as with square-law rectifiers.

[1] E. H. Armstrong, "A method of reducing disturbances in radio signaling by a system of frequency modulation," Proc. I.R.E., vol. 24, pp. 689–740; May, 1936. (Balanced frequency detector with linear slope filters and linear rectifiers, his Figs. 5 and 6 compared with Fig. 2 herein.)

From these considerations, it appears that the first type with linear slope filters and linear rectifiers has some advantages over the other types. It is found to operate free of distortion if the frequency modulation is held within the limits of the linear slope in Fig. 2(c), regardless of the waveform of modulation and the marginal sidebands outside of these limits. The most efficient rectifier is the diode peak detector, which best meets the requirement of linearity.

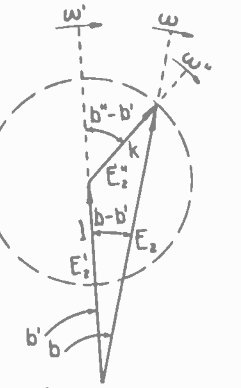

Fig 3—The vector sum of two frequency-modulated signals.

Since the second type with linear slope filters and square-law rectifiers is the simplest for theoretical study, its behavior also is to be described.

If a perfect limiter is assumed preceding the detector, the amplitude effects are removed and it becomes immaterial whether linear or square-law rectifiers are used. Of course, a perfect limiter is not possible in practice, but comparable performance can be obtained with practical limiters of careful design.

III. The Amplitude and Frequency Modulation in the Resultant of Two Signals Superimposed[2]

In response to two signals, there is a difference in operation with and without a limiter. This difference is caused by the amplitude modulation which is present in the composite signal. With a perfect limiter, only the frequency modulation contributes to the detector output.

In the study of the behavior without a limiter, the theoretical derivations do not require the separate expression of the amplitude and frequency modulation. If the limiter is present, however, the frequency modulation does have to be expressed separately.

The superposition of two signals of constant amplitude is shown vectorially in Fig. 3. The unit vector E_2' is the desired signal and the vector E_2'' of length k is the undesired signal. Their frequencies and therefore their phase angles b' and b'' are modulated. The resultant of these two vectors is the composite signal E_2 whose phase angle is b. The average frequency of the composite signal is that of its stronger signal because the relative phase displacement caused by the weaker signal is alternately forward and backward.

Fig. 4 shows the alternating modulations caused by the weaker signal. The fundamental frequency of these modulations is the beat frequency, which is the frequency difference between the two signals. To the extent that the waveforms depart from sine waves, they include also harmonics of the beat frequency.

The relative amount of amplitude modulation is denoted a and its waveform is shown as Fig. 4(a). It is plotted to the scale a/k to show the change of waveform within the same limits and with nearly the same fundamental component. It is noted that this is the shape of the envelope of a carrier accompanied by a single-sideband component.

The alternating phase displacement caused by the weaker signal is shown in Fig. 4(b). It is plotted to the scale b/k, which maintains the same fundamental component. As the amplitude of the weaker signal approaches that of the stronger signal ($k=1$) the phase modulation approaches a saw-tooth waveform. The phase reversal at the instant when the two signals are in opposition appears in the saw-tooth waveform as a phase step of π radians. This case is shown for a value of k approaching unity from the lesser side, so the phase step is negative as required to complete the saw-tooth waveform. This is necessary to assure that the average frequency remains that of the stronger signal.

The corresponding waveforms of frequency modulation are shown in Fig. 4(c), plotted in terms of ω/k

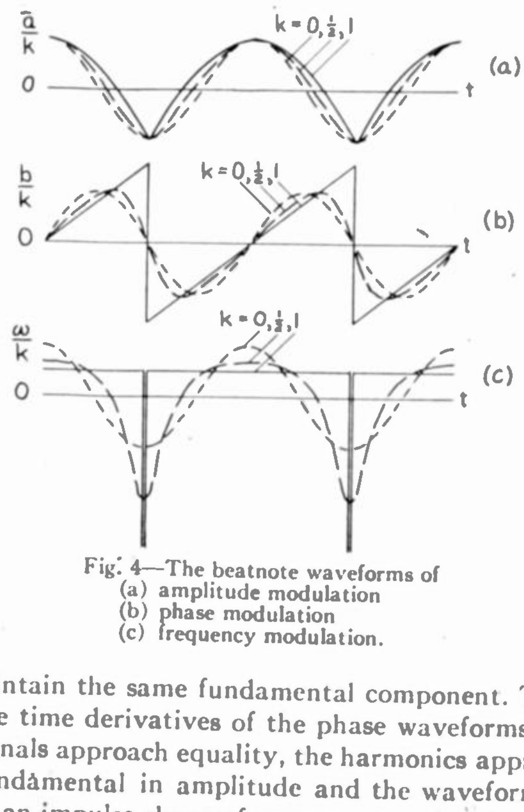

Fig. 4—The beatnote waveforms of
(a) amplitude modulation
(b) phase modulation
(c) frequency modulation.

to maintain the same fundamental component. These are the time derivatives of the phase waveforms. As the signals approach equality, the harmonics approach the fundamental in amplitude and the waveform assumes an impulse shape of very great peak value.

It is these beatnote waveforms of frequency modulation to which the receiver responds if a perfect limiter

[2] Hans Roder, "Noise in frequency modulation," *Electronics*, vol. 10, pp. 22–25, 60, 62, and 64; May, 1937. (The vector resultant of two signals, his Fig. 4 compared with Fig. 3 herein.)

is assumed. Since the average frequency of the composite signal is that of the stronger signal, there is no response proportional to the frequency modulation of the weaker signal. Therefore there is no crosstalk from an undesired signal weaker than the desired signal. Reciprocally, a stronger undesired signal completely masks the desired signal. There is heard only the modulation of the stronger signal, together with the beatnote and its harmonics.

If there is not a perfect limiter assumed, there is some response to the amplitude modulation in the composite signal, Fig. 4(a). The amplitude modulation has not only the beatnote fundamental and harmonic components, but also an increase of its average value caused by the presence of the undesired signal. This is in contrast to the average frequency, which remains unchanged. It is the change of the average amplitude to which is attributed the crosstalk interference from a weaker signal, which is to be described and which occurs only in the absence of a perfect limiter. The beatnote components of the amplitude modulation modify the amplitude and harmonic content of the beatnote output from the frequency detector, in a manner which also remains to be described.

IV. The Response to Two Signals[3,4]

In the derivation of the detector output in response to one or two signals, the conditions in the system are assumed which normally give an output equal to the input modulating signal. Each modulating signal E_1 varies within the limits of ± 1. The resulting frequency modulation of the modulated-carrier signals E_2 is within the limits of $\pm f_c$, the same as the limits on the range of linear operation of the frequency detector with linear slope filters and linear rectifiers, Fig. 2(c). Therefore, the bandwidth of modulation is $2f_c$. The desired signal E_2' has unit carrier amplitude, and the undesired signal has a carrier amplitude k, which is therefore the relative amplitude of the undesired signal.

The simplest expression for the response to two signals is obtained in the case of square-law rectifiers, because it happens that the amplitude and frequency modulation combine to give a beatnote free of harmonics. The output for this case is simply

$$\begin{aligned} E_4 = \; & E_1' && \text{desired signal} \\ & + k^2 E_1'' && \text{crosstalk} \\ & + k(E_1' + E_1'') \cos (b'' - b') && \text{beatnote. (1)} \end{aligned}$$

This output comprises one term which is a replica of the desired-signal modulating voltage E_1' and another which is a replica of the undesired-signal modulating voltage E_1''. The latter is identified as the crosstalk interference, which is recognizable as the undesired signal. The remaining term is the beatnote between the two signals. Its phase is the difference between the progressive phase angles b' of the desired signal and b'' of the undesired signal. Therefore the frequency of the beatnote is modulated in accordance with the difference of the two modulating voltages $(E_1'' - E_1')$. The amplitude of the beatnote is modulated in accordance with the sum of the two modulating voltages. This combination of amplitude and frequency modulation of the beatnote gives it a nondescript character which retains no recognizable qualities of the two signals except their syllabic pulsations.

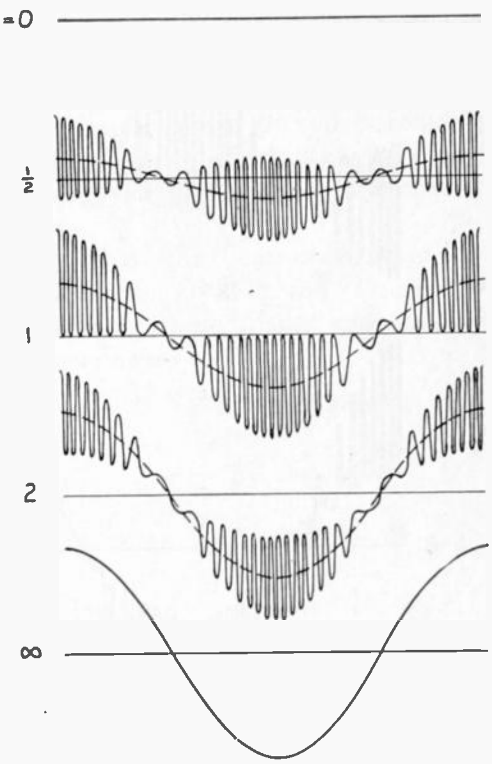

Fig. 5—The output in the case of square-law rectifiers and the desired signal unmodulated showing the crosstalk and beatnote components.

The interference is most noticeable while the desired signal is unmodulated ($E_1' = 0$), leaving only the crosstalk and beatnote terms. Their amplitude would increase indefinitely with the strength of the undesired signal (k), were it not that an automatic volume control is used. It is assumed that this control holds uniform the mean-square voltage of the composite signal. This has the effect of dividing all terms in (1) by the factor $(1 + k^2)$.

On this basis, Fig. 5 shows the interference output in the case of square-law rectifiers, for several different values of the relative strength of the undesired signal. The undesired signal has sinusoidal frequency modulation, the extent of frequency modulation in each direction being $33\pi/2$ times the modulating frequency.

[3] M. G. Crosby, "Frequency modulation propagation characteristics," Proc. I.R.E., vol. 24, pp. 898–913; June, 1936. (Two signals, comprising the same signal received over two different paths.)

[4] M. G. Crosby, "Frequency modulation noise characteristics," Proc. I.R.E., vol. 25, pp. 472–514; April, 1937. (Two signals, comprising a desired signal and noise. The case of a perfect limiter. The separate identification of the desired-signal output and the beatnote interference. The beatnote waveform, his Fig. 4 compared with Fig. 4 herein. The conical pattern, his Fig. 16 compared with Fig. 8 herein.)

Therefore the maximum frequency of the beatnote is $33\pi/2$ times that of the modulation. This large ratio is taken to clarify the difference between the crosstalk of relatively low frequency and the superimposed beatnote of varying frequency. The incommensurate ratio is taken to give an integral number (33) of beatnote cycles during one cycle of modulation. An odd multiple of $\pi/2$ is taken to give a symmetrical waveform of the beatnote during each half cycle of modulation.

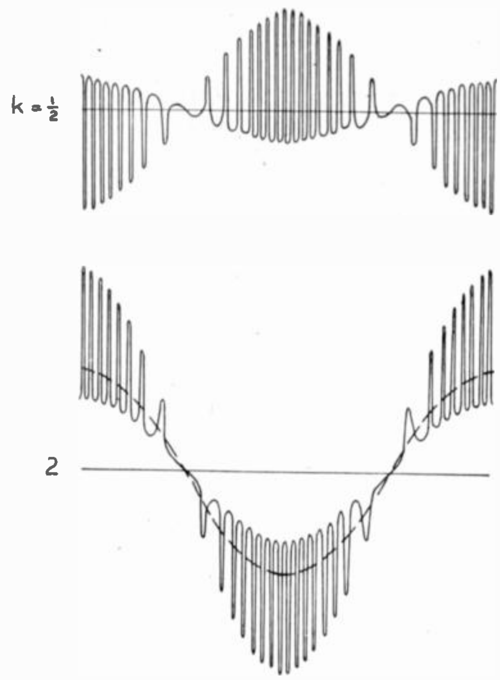

Fig. 6—The output in the case of a perfect limiter showing the unsymmetrical waveform of the beatnote component.

Fig. 5 shows the relative importance of crosstalk and beatnote, depending on the relative signal strength. The beatnote term predominates if the undesired signal is much weaker, and reaches a maximum if the two signals are equal. The crosstalk term increases with the strength of the undesired signal until it exceeds the beatnote term; at the same time, the desired signal is blocked out by the automatic volume control, but this effect is not shown in Fig. 5.

If wide-band frequency modulation is used, the beat frequency during part of the time is so high that it is inaudible and may be filtered out in the audio-frequency amplifier. In this case, the beatnote is diminished during the peaks of modulation, and its wide range of frequency modulation gives it a sizzling or spitting sound.

The corresponding interference with linear rectifiers or a limiter differs in detail but retains the same general characteristics.

In the case of a perfect limiter, it is immaterial which type of rectifier is used in the frequency detector, and also whether the detector is balanced, because the amplitude modulation is removed from the composite signal. If the undesired signal is the weaker ($k<1$), the remaining frequency modulation yields the output

$$\begin{aligned} E_4 = {} & E_1' && \text{desired signal} \\ & + (E_1''-E_1')[k\cos(b''-b') && \text{beatnote} \\ & \quad - k^2\cos 2(b''-b') \\ & \quad + k^3\cos 3(b''-b') - \cdots]. && (2) \end{aligned}$$

There is no crosstalk from the weaker signal, because the average frequency is that of the stronger signal, as noted with reference to Figs. 3 and 4. The beatnote and its harmonics have the waveform of Fig. 4(c). The harmonics are unimportant if there is a substantial difference of signal strength. This case differs from the preceding case in that the beatnote amplitude and frequency are both modulated in step, because both are proportional to the difference of the two modulating voltages $(E''-E')$.

Fig. 6 shows for comparison two of the examples of Fig. 5, but for the case of a perfect limiter. For the undesired signal weaker, only the beatnote appears with its harmonics. The general shape of the waveform is inverted relative to the corresponding example in Fig. 5. With the undesired signal stronger, the crosstalk appears in full strength and the beatnote peaks are inverted. The transition between these two conditions is a critical test for equality of the two signals, if a limiter is used which has nearly ideal properties.

The case of linear rectifiers is the most difficult of analysis because its output terms involve elliptic integrals, as does also its factor of automatic volume control which maintains uniform the average voltage of the composite signal. Also the two signals cause distortion of each other. However, the behavior which is most closely identified with the use of linear rectifiers is easily expressed if one signal is much weaker than the other. It is assumed that the undesired signal is the weaker ($k<1$).

For one approximate expression with linear rectifiers, the only assumption is a difference of signal strength so great that the second and higher powers of k are negligible.

$$\begin{aligned} E_4 = {} & E_1' && \text{desired signal} \\ & + kE_1''\cos(b''-b') && \text{beatnote.} \quad (3) \end{aligned}$$

On these assumptions, the crosstalk from the weaker signal E_1'' is lost, as well as the distortion and masking effects on the desired signal. The harmonics of the beatnote are also lost. There remains only the replica of the desired signal E_1' and the fundamental component of the beatnote interference. It is noted that the amplitude of the beatnote depends on the modulating voltage E_1'' of only the undesired signal, not of both signals as in the preceding cases. The beatnote disappears during interruptions in the modulation of the undesired signal.

For another approximate expression with linear rectifiers, only the third and higher powers of k are neglected, but the desired signal is assumed to be unmodulated ($E_1'=0$) so only the interference remains.

$$E_4 = \frac{k^2}{2} E_1'' \qquad \text{crosstalk}$$

$$+ E_1'' \left[k \cos (b''-b') \qquad \text{beatnote} \right.$$

$$\left. - \frac{k^2}{2} \cos 2(b''-b') + \cdots \right]. \qquad (4)$$

Here it appears that all three terms have coefficients midway between the corresponding terms in (1) for square-law rectifiers and those in (2) for a limiter, on the same assumption that the desired signal is unmodulated. The fundamental beatnote term, on these assumptions, is the same in all cases. The crosstalk term is twice as great with square-law rectifiers but absent with a limiter. The second harmonic of the beatnote is twice as great with a limiter but absent with square-law rectifiers.

Since the composite signal comprising both of the frequency-modulated signals has both amplitude and frequency modulation, and since the three cases studied differ only in their response to amplitude modulation, a comparison of these cases enables the effects of amplitude modulation to be identified separately from those of frequency modulation. The desired-signal output is the same in all cases if the undesired signal is much weaker, so the comparisons are based on the interference terms.

The crosstalk comparison for the cases with a limiter or with linear or square-law rectifiers is based on (1), (2), and (4), assuming the desired signal unmodulated and the undesired signal slightly weaker so the third and higher powers of k are negligible. The crosstalk terms in these cases are summarized as follows:

(limiter)	$0\, E_1''$	(2a)
(linear)	$\frac{k^2}{2} E_1''$	(4a)
(square-law)	$k^2\, E_1''$	(1a)
(amplitude effect)	$\frac{k^2}{2} E_1''$.	(5a)

The amplitude effect is the difference between the limiter and linear cases, or one half the difference between the limiter and square-law cases. Since the crosstalk is absent with a perfect limiter, it appears to be caused by the amplitude modulation in the composite signal. It is twice as great with square-law as with linear rectifiers, because the square-law rectifiers are doubly sensitive to amplitude modulation.

The comparison of the beatnote fundamental component for the three cases is based on (1), (2), and (3), assuming only that the undesired signal is sufficiently weak to make negligible the second and higher powers of k.

(limiter)	$(E_1''-E_1')\, k \cos (b''-b')$	(2b)
(linear)	$E_1''\, k \cos (b''-b')$	(3b)
(square-law)	$(E_1''+E_1')\, k \cos (b''-b')$	(1b)
(amplitude effect)	$E_1'\, k \cos (b''-b')$.	(5b)

Again, the effect of amplitude modulation is merely the difference between the limiter and linear cases, or half the difference between the limiter and square-law cases. The significance of this effect remains to be described further on in this discussion. Its cause is not obvious and seems not to be susceptible of simple explanation.

The comparison of the beatnote second-harmonic component is based on (1), (2), and (4), again assuming the desired signal unmodulated and neglecting the third and higher powers of k.

(limiter)	$-\, k^2 E_1'' \cos 2(b''-b')$	(2c)
(linear)	$-\frac{k^2}{2} E_1'' \cos 2(b''-b')$	(4c)
(square-law)	$-\, 0 E_1'' \cos 2(b''-b')$	(1c)
(amplitude effect)	$+\frac{k^2}{2} E_1'' \cos 2(b''-b')$.	(5c)

Here the effect of amplitude modulation is derived in the same manner. It is interesting but this term is of secondary importance in practice.

These three comparisons show that the limiter and square-law cases are the two extremes, while the linear case is intermediate. The limiter case has no crosstalk, while the square-law case has no beatnote harmonics. The linear case is least simple because it has all terms and because the exact expression of the coefficients involves transcendental factors such as elliptic integrals. Each of the three comparisons is based on approximations chosen to show the intermediate value of the coefficient in the linear case. In general, this represents a tendency rather than an exact rule. For comparison with the frequency effects expressed alone in (2), the approximate amplitude effects are summarized as follows:

$$\frac{k^2}{2} E_1'' \qquad \text{crosstalk}$$

$$+ k\, E_1' \cos\ (b''-b') \qquad \text{beatnote}$$

$$+ \frac{k^2}{2} E_1'' \cos 2(b''-b'). \qquad (5)$$

These terms are present, in addition to the frequency terms (2), with linear rectifiers and no limiter. They are doubled by changing to square-law rectifiers and are removed by the insertion of a perfect limiter preceding the detector.

There is a particular significance to the fundamental beatnote term in (5), the only important one of the amplitude terms if the undesired signal is much the weaker. The amplitude of this beatnote term is proportional to the frequency modulation of the desired signal, which is the stronger signal and therefore determines the average frequency of the composite signal. The unbalancing of the frequency detector by the frequency modulation of the stronger signal is a reasonable explanation of the amplitude detection responsible for this beatnote term.

V: The Conical Pattern

While the desired signal and the crosstalk have obvious significance in the detector output, it is difficult to interpret the beatnote as to its interference effect. This is especially true in wide-band frequency modulation, the beatnote being inaudible part of the time. As an aid in testing and interpreting the beatnote interference, the conical pattern is to be described with its various forms and uses.

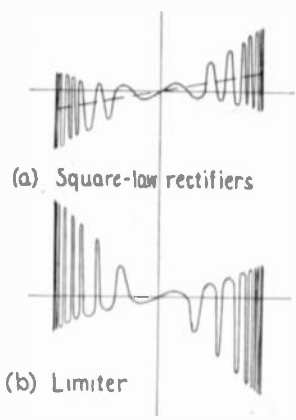

Fig. 7—The output of Figs. 5 and 6 for $k=1/2$ but plotted against the beat frequency to form a conical pattern.

Fig. 7 shows two examples of the conical pattern. They are based on Figs. 5 and 6, but are traced not against time but rather against the modulating voltage of the undesired signal. The desired signal being unmodulated, the beat frequency is proportional to the modulating voltage of the undesired signal. Therefore the output is effectively plotted against the frequency of the beatnote, and this pattern shows the relation between the amplitude and the frequency of the beatnote. Fig. 7 is computed for $k=\frac{1}{2}$, with (a) for square-law rectifiers as in Fig. 5 and (b) for a limiter as in Fig. 6.

Each of the curves in Fig. 7 is called a "conical pattern" because it is the lateral projection of a three-dimensional curve traced on a conical surface by a vector which rotates about the axis of this surface. This vector rotates at the beat frequency, and shifts along the axis at a distance from the center proportional to the beat frequency. In Fig. 7, its amplitude is also proportional to the beat frequency, so its point travels on a conical surface. The conical pattern was originated to show just what appears in Fig. 7, the proportionality between the frequency and amplitude of the beatnote caused by a weak component of noise superimposed on a strong unmodulated carrier of the desired signal.

The conical pattern is most easily shown on an oscilloscope. The output of the receiver and the frequency-modulating voltage of the undesired signal are applied to the respective vertical and horizontal deflecting plates. The modulating frequency and the extent of frequency modulation are usually not critically related, so the pattern appears not as a steady repetitive trace but rather as a trace shifting within the envelope of the pattern. This does not decrease the value of the pattern for showing the relation between frequency and amplitude of the beat note. An oscilloscope having a screen of long persistence is advantageous in observing the envelope rather than the trace itself.

The conical pattern is most distinct if the modulating frequency is much less than the extent of frequency modulation, as in Fig. 7. Otherwise the conical pattern is indistinct, unless there are superimposed a sufficient number of noncoincident traces to determine the envelope. The envelope loses its sharp corners if there is a low-pass filter following the detector, even though

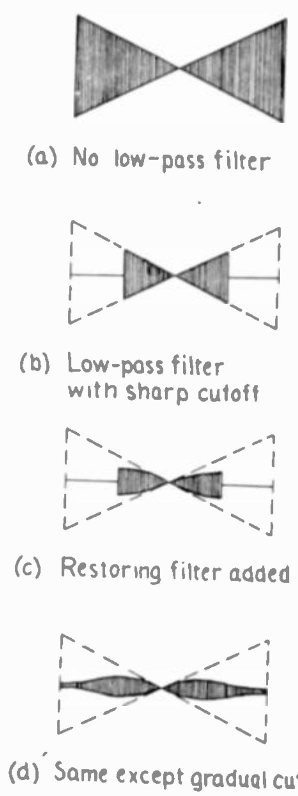

Fig. 8—Conical patterns showing the effect of low-pass filters after the detector.

this filter has a cutoff frequency slightly higher than the highest frequency of the beatnote. This distortion of the envelope is caused by the failure to retain the sidebands outside the width of frequency modulation, which spread out farther during rapid modulation. If the beatnote has strong harmonics, the envelope is distorted also by the loss of these harmonics.

In Fig. 8 are shown some examples of the conical pattern in its practical applications. In each case, the modulating frequency is sufficiently low to give a sharp outline of the pattern. Fig. 8(a) shows merely the conical envelope without distortion. Fig. 8(b) shows the effect of a sharp-cutoff low-pass filter which passes the beatnote only while the beat frequency is less than the cutoff frequency. This filter is intended to pass as much of the audio range as is needed for adequate reproduction. The conical pattern shows clearly the accompanying reduction of the peak amplitude of the beatnote interference.

If the transmitter employs pre-emphasis of the higher frequencies of modulation, this is compensated by a restoring filter in the receiver.[5] The restoring filter merely attenuates the higher frequencies within the audio range. Its effect on the beatnote output is shown in Fig. 8(c), including a further reduction of the peak amplitude. In practice, the low-pass filter is more likely to have a gradual cutoff which, with the restoring filter, gives a conical pattern of the shape of Fig. 8(d).

The interference effect involves the variation of audibility of the beatnote during the modulation of its frequency. Knowing the characteristics of audition and the frequency scale of the conical pattern enables an estimation of the beatnote interference in terms of its audibility. Such an estimate should take into account also the time during which the beatnote is audible.

The conical patterns of Figs. 7 and 8 are observed with the unmodulated desired signal on the mean frequency of the modulated signal, so the patterns are symmetrical. The center point indicates the occurrence of equality between the frequencies of the two signals and the center frequency of the balanced detector. In the case of a limiter in the receiver, formula (2), it is found that the amplitude of the beatnote is proportional to its frequency, regardless of the tuning relative to the frequency detector, so the crossover point on the envelope always represents equality between the signal frequencies, or zero beat.

Relying on this relationship, Fig. 9 shows the effect of detuning the unmodulated signal. The conical pattern is symmetrical (a) while the unmodulated signal is on the mean frequency of the modulated signal, but departs from symmetry as the unmodulated signal is detuned toward the lower limit of frequency modulation, through (b) to (c). The apex or crossover point of the envelope moves to one edge of the pattern (c) as the unmodulated signal is detuned to the limit of frequency modulation. This gives a critical test for comparing the frequency modulation of one signal against the steady frequency of another. It is most reliable if the modulated signal is weaker than the unmodulated signal, so the conical pattern is nearly horizontal, and the signals differ in strength enough so the requirements on the limiter are not too severe. Such a test is needed for checking the performance of a frequency-modulated signal generator, and would be valuable as a monitor in a transmitter.

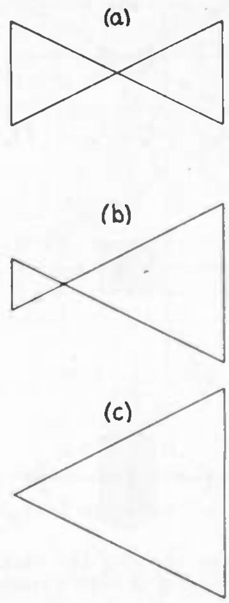

Fig. 9—Conical diagrams showing the effect of detuning an unmodulated signal relative to a frequency-modulated signal.

The detuning of the unmodulated signal during modulation of the other signal gives some indication of what happens to the beatnote amplitude during modulation of both signals. In general, this depends on the limiter and detector properties of the receiver, Fig. 9 being valid only for a receiver with a limiter.

Fig. 10 shows the outer limits of the conical envelope with both signals modulated, in the three cases, of a limiter (a), linear rectifiers (b), or square-law rectifiers (c). These patterns are traced against the difference of the two modulating voltages $(E_1''-E_1')$ so the horizontal displacement is still proportional to the beat frequency. The coefficient of the beatnote term in each of (2), (3), and (1) is used to determine the maximum beatnote amplitude at any beat frequency, both signals being modulated within the same limits (E_1' and E_1'' between -1 and $+1$). The maximum beat frequency occurs with opposite maximum modulation of the signals, so it is double the maximum modulation ($\pm f_c$). It is accompanied by maximum amplitude in the case of a limiter (a) or minimum in the case of square-law rectifiers (c). In the cases without a limiter, it is important that both signals are tuned to the frequency detector. In the case of linear rectifiers (b), the amplitude is modulated only by the undesired signal E_1'', so the peak amplitude of the beat note is the same for all beat frequencies.

The differences among the three cases are present only during simultaneous modulation of both signals,

[5] M. G. Crosby, "The service range of frequency modulation," *RCA Rev.*, vol. 4, pp. 349–371; January, 1940. (The pre-emphasis and restoration of the higher audio frequencies in the modulating signal, his Figs. 4 and 5.)

and are caused by (5b) identified with amplitude detection accompanying the frequency detection. They are associated with the idea that the departure of the stronger signal from the center frequency unbalances the frequency detector, leaving it sensitive to the beatnote amplitude modulation. The greatest amplitude of low-frequency beatnote occurs while both signals are modulated in the same sense and to the maximum extent, so the detector is furthest from balance.

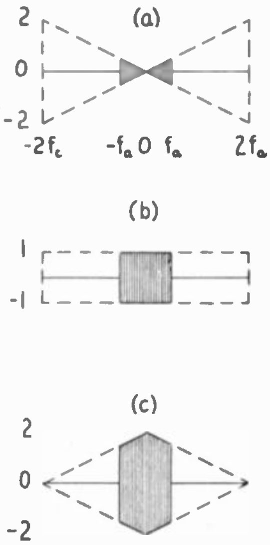

Fig. 10—Conical diagrams showing the maximum amplitude of beatnote during modulation of both signals.
(a) limiter
(b) linear rectifiers
(c) square-law rectifiers.

The shaded areas in Fig. 10 show the region passed by a low-pass filter similar to that used as a basis for Fig. 8(b). The remaining peak amplitude of the beatnote is least with a limiter (a) and greatest with square-law rectifiers (c). This is in contrast to the beatnote amplitude with one signal unmodulated, which is nearly independent of the limiter and rectifier properties in the receiver. This distinction has little practical significance, because the output of the stronger signal usually obscures such beatnote interference as is caused by its modulation.

VI. The Comparison of Crosstalk and Beatnote Interference Under Various Conditions

Having shown the general characteristics of the output in response to two signals, there remains to present graphically the relative importance of the crosstalk and beatnote interference, depending on the relative strength of the two signals and on the properties of the limiter or of the detector. First, the response of all kinds is to be summarized for the limiter and square-law cases, then the crosstalk output is to be compared for the various cases.

In the case of a perfect limiter, Fig. 11 shows the peak amplitude of all terms relative to that of the desired signal alone. Each term is evaluated during modulation of only one of the two signals. The desired signal (*a*) has its normal value, unless the undesired signal is stronger and therefore completely masks the desired signal. The crosstalk (*b*) is absent if the desired signal is the stronger, but otherwise completely displaces the desired signal. The beatnote fundamental component (*c*) has a maximum value for both signals equal. In general, its relative peak value is equal to the voltage ratio of the weaker signal over the stronger. The beatnote harmonics add to the fundamental, their total peak value becoming indefinitely great for equal signals.

A low-pass filter reduces the peak value of the beatnote fundamental in the ratio of its cutoff frequency over the maximum frequency modulation f_a/f_c, as shown in Fig. 8(b). This is exemplified by Fig. 11(d) for a ratio of 1/5; the low-pass filter may have a sharp cutoff at 15 kilocycles, the upper limit of the audio range, with maximum modulation of 75 kilocycles.

If there is also a restoring filter, of nominal cutoff frequency f_r, and if its cutoff frequency is less than that of the low-pass filter, there is a greater reduction of the beatnote approximately in the ratio f_r/f_c, as shown in Fig. 8(c). This further reduction is indicated in Fig. 11(e) for a ratio of 1/50, the restoring filter having a gradual cutoff at 1.5 kilocycles.

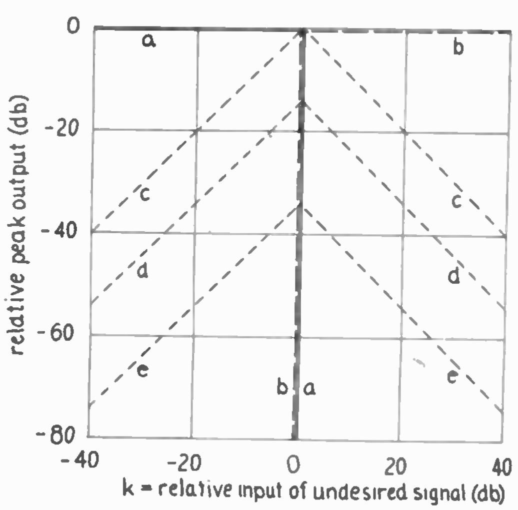

Fig. 11—The relative peak output of the individual components in the case of a perfect limiter.
(a) desired signal
(b) crosstalk
(c) beatnote fundamental component with no low-pass filter
(d) same with added low-pass filter, $f_a = f_c/5$
(e) same with added restoring filter, $f_r = f_c/50$.

The discontinuities in the curves of Fig. 11, at the condition of two equal signals, are caused by the assumption of a perfect limiter. In practice, this assumption fails at this condition, so the curves are rounded off and there is some overlap between the desired signal (a) and the crosstalk (b). These discontinuities are absent in the square-law case and are less severe in the linear case.

In the case of square-law rectifiers, Fig. 12 shows the same relative peak amplitude of the output terms. A perfect automatic amplification control is assumed to hold uniform the mean-square value of the composite signal voltage, to make this case comparable with that of a perfect limiter. The desired signal (*a*) and the crosstalk (*b*) overlap but the stronger signal is favored very much. There is no effect of the stronger signal masking the weaker, as occurs in the limiter, but there is a similar effect of the stronger signal attenuating the weaker through the automatic control. The beatnote curves correspond to those of Fig. 11, (*c*) without any low-pass filter, (*d*) with the low-pass filter, and (*e*) with the restoring filter. In this case, there are no beatnote harmonics.

In each of Figs. 11 and 12, the crosstalk curve (*a*) and the beatnote curve (*b*) add up to unity. This is true as a general rule, on the assumption that there is either a perfect limiter or a perfect automatic volume control using a rectifier of the same type as those in the frequency detector.

The case of linear rectifiers is intermediate in behavior, somewhere between Figs. 11 and 12. All three cases have nearly the same beatnote interference, on the same assumption that only one of the two signals is modulated at a time. The more interesting comparison among the three cases therefore is their crosstalk interference.

Fig. 13 gives a comparison of the crosstalk inter-

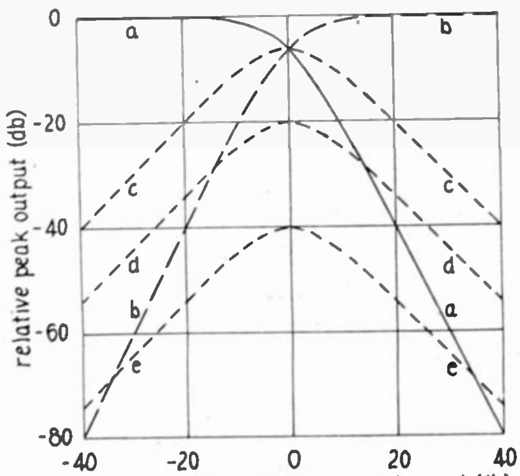

Fig. 12—The relative peak output of the individual components in the case of square-law rectifiers.
(a) desired signal
(b) crosstalk
(c) beatnote with no low-pass filter
(d) same with added low-pass filter, $f_a = f_c/5$
(e) same with added restoring filter, $f_r = f_c/50$.

ference for several cases. With the undesired signal the weaker ($k<1$), the perfect limiter avoids any crosstalk (*a*), and that from linear rectifiers (*b*) is half as great as that from square-law rectifiers (*c*). The curve (*b*) is computed from elliptic integrals and involves the assumption of an automatic control holding uniform the average voltage of the composite signal.

Practical cases are likely to fall between the case of a perfect limiter (*a*) and that of linear rectifiers (*b*), because nearly linear rectifiers are used with an imperfect limiter, a perfect limiter being impossible of realization. The practical limiter is imperfect in two respects, a failure of limiting action below a certain threshold value, and a departure from uniform output above the threshold value. (Other imperfections such as departure from instantaneous action are here neglected.) If the limiter action is level above the

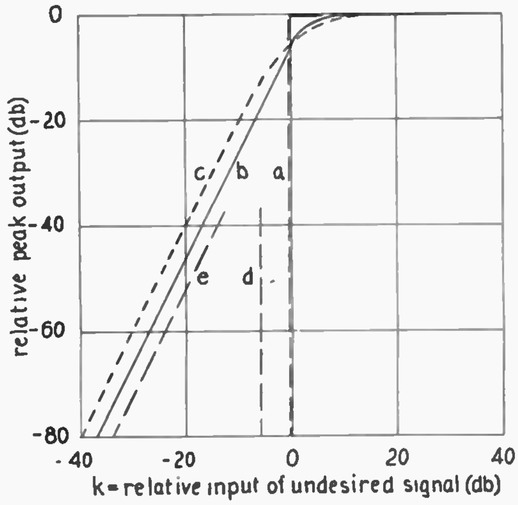

Fig. 13—The relative peak output of the crosstalk component.
(a) perfect limiter
(b) no limiter, linear rectifiers
(c) no limiter, square-law rectifiers
(d) limiter with threshold at 1/2 amplitude of desired signal
(e) square-root limiter and linear rectifiers.

threshold and the desired signal is at least double the threshold voltage, no crosstalk can occur unless the undesired signal exceeds one half the desired-signal voltage. This limit is indicated in Fig. 13(*d*). If instead the limiter has a slope halfway between level and linear, the crosstalk output approaches one half that of the linear case. This is called the "square-root" case, describing its slope of one half, and is indicated in Fig. 13(*e*). Favorable conditions in practice are likely to fall between (*a*) and (*d*).[6]

The peak amplitude of the crosstalk is less than that of the beatnote in all cases, if both are less than $(f_r/f_c)^2$ times the peak amplitude of the desired signal. Under the assumptions of maximum frequency modulation, $f_c = 75$ kilocycles, and the restoring filter, $f_r = 1.5$ kilocycles, the beatnote predominates if both interference terms are at least 78 decibels below the desired signal. This relation is based on Fig. 12 for the square-law case, which has the greatest crosstalk. This ratio becomes $2(f_r/f_c)^2$ or -72 decibels in the case of linear rectifiers.

The relative audible interference from the crosstalk and beatnote depends on the kind of program and the requirements of reproduction. For the same peak value, the crosstalk and beatnote are of comparable audibility, but the spitting sound of the beatnote tends to make it the more detrimental. If it is necessary to

[6] I. R. Weir, "Field tests of frequency and amplitude modulation with ultrahigh-frequency waves," *Gen. Elec. Rev.*, vol. 42, pp. 188–191; May, 1939; and pp. 270–273; June, 1939. (The common-channel interference between two signals, his Figs. 9 and 10.)

hold the interference at least 70 or 80 decibels below the desired signal, the beatnote is the determining factor, and the presence or absence of a limiter is unimportant. If more interference is tolerable, the crosstalk is the determining factor and it can be reduced by a limiter.

Increasing the bandwidth of frequency modulation ($2f_c$) for the same bandwidth of the modulating signal (f_a) has no effect on the crosstalk but does enable the reduction of the beatnote output by a low-pass filter after the detector. Such a filter attenuates the beatnote during the time in which the beat frequency is outside the frequency band occupied by the modulating signal.

VII. The Derivation for One Signal in Various Types of Frequency Detectors[7]

The theoretical derivation is simplified by the use of the zero-frequency carrier. This procedure has been shown to yield the practical solution, even in cases of unsymmetrical sidebands as in frequency modulation.[8] The carrier frequency does not appear in the statement of the problem or in its solution.

In this derivation, no specific form of modulating signal is assumed. This avoids the usual limitation to a sinusoidal modulating signal. This generalization actually simplifies the expression of the solution.

The modulating signal, $E_1(t)$ in Fig. 1, has any variation with the time t, subject to some limitation on its bandwidth. (In sound transmission, it is the audio-frequency input signal.) The frequency modulator M is regarded as operating on a carrier of unit amplitude and zero frequency. Therefore the frequency deviation is the same as the frequency f of the modulated signal $E_2(t)$. This frequency deviation is proportional to the modulating signal

$$f(t) = f_c E_1(t) \tag{6}$$

in which f_c is the deviation per unit voltage of E_1.

The three forms of frequency symbols are used interchangeably, whichever fits into the expression at hand:

$$\omega = 2\pi f; \quad p = i\omega = i2\pi f \tag{7}$$

in which ω is the angular frequency and p corresponds to the differential operator D_t with respect to time.

The modulated carrier has a phase angle b related to the frequency modulation as follows:

$$\frac{db}{dt} = \omega = \omega_c E_1 \tag{8}$$

$$b = b_0 + \int_0^t \omega\, dt = b_0 + \omega_c \int_0^t E_1\, dt \tag{9}$$

in which b_0 is the phase angle at $t=0$. The progressive phase angle b henceforth includes both frequency and time variables. The modulated carrier is represented by the vector

$$E_2 = \exp ib = \cos b + i \sin b. \tag{10}$$

There are several types of balanced frequency detectors to be considered, as described with reference to Fig. 2. They differ in the kind of slope filters and the kind of rectifiers. Without loss of generality, the same f_c is taken as the intercept frequency of the slope filters.

The linear slope filter of Fig. 2(a) has the filter factors

$$F_+(f) = \frac{1}{2} + \frac{f}{2f_c} = \frac{f_c+f}{2f_c}; \quad F_-(f) = \frac{1}{2} - \frac{f}{2f_c} = \frac{f_c-f}{2f_c} \tag{11}$$

in which the + and − subscripts denote the two sides of the balanced detector. Each of these factors changes between zero and unity between the intercepts at $\pm f_c$. The fact that the slope continues over negative and positive values of the filter factor denotes the use of a resonant trap in each filter.

The filter factor on each side is expressed in terms of the differential operator and then applied to the modulated voltage E_2 to secure the differentiated voltages E_{3+} and E_{3-} in Fig. 1:

$$F_\pm = \frac{1}{2} \pm \frac{p}{2p_c} = \frac{1}{2} \pm \frac{1}{2p_c} D_t \tag{12}$$

$$E_{3\pm} = \left(\frac{1}{2} \pm \frac{1}{2p_c} D_t\right) E_2 = \frac{1}{2} \exp ib \pm \frac{i}{2p_c} \cdot \frac{db}{dt} \exp ib$$
$$= \left(\frac{1}{2} \pm \frac{1}{2} E_1\right) \exp ib. \tag{13}$$

This shows the amplitude modulation superimposed on the frequency modulation by the slope filters.

This application of the differential operator p is common ground between Heaviside operational methods and Fourier integral methods.[9,10] In this case, it is the same as the transformation

$$E = (R + j\omega L)I = (R + Lp)I = (R + L \cdot D_t)I = RI + L\frac{dI}{dt} \tag{14}$$

in which E is the voltage, R is the resistance, L is the inductance, and I is the current having any variation with the time t, and D_t means d/dt.

The resulting voltages E_{3+} and E_{3-} are delivered to the rectifiers, which are linear rectifiers in this case. Since the unrectified signal is expressed as a modulated zero-frequency carrier, its magnitude is the signal envelope to which the rectifier responds. The magnitude

[7] J. R. Carson and T. C. Fry, "Variable frequency electric circuit theory with application to the theory of frequency modulation," *Bell Sys. Tech. Jour.*, vol. 16, pp. 513–540; October, 1937.

[8] H. A. Wheeler, "The solution of unsymmetrical-sideband problems with the aid of the zero-frequency carrier," Proc. I.R.E., vol. 29, pp. 446–458; August, 1941.

[9] Vannevar Bush, "Operational Circuit Analysis," 1929. See pp. 17–21.

[10] G. A. Campbell and R. M. Foster, "Fourier integrals for practical applications," Bell Telephone System Monograph B-584, September, 1931. Abridgment, *Bell Sys. Tech. Jour.*, vol. 7, pp. 639–707, October 1928. See Table I, No. 208.

of E_{3+} or E_{3-} carries an amplitude modulation which is proportional to the modulating signal E_1.

The linear rectifiers produce output voltages E_{4+} and E_{4-} equal to the envelope amplitudes of the differentiated signals, in this case the magnitude of the voltages E_{3+} and E_{3-}:

$$E_{4\pm} = |E_{3\pm}| = |\tfrac{1}{2} \pm \tfrac{1}{2}E_1| \tag{15}$$

as shown in Fig. 2(b). The output voltage E_4 from the balanced detector is the differential output of the rectifiers:

$$E_4 = E_{4+} - E_{4-} = |\tfrac{1}{2} + \tfrac{1}{2}E_1| - |\tfrac{1}{2} - \tfrac{1}{2}E_1|. \tag{16}$$

This has a range of linear proportionality limited by the intercept frequencies, in which the output is equal to the modulating voltage:

$$-f_c < f < f_c; \quad -1 < E_1 < 1; \quad E_4 = E_1 \tag{17}$$

as shown in Fig. 2(c).

In this derivation, it is notable that the bandwidth between the intercept frequencies $\pm f_c$ need be only sufficient to include the maximum value of the deviation f. The marginal sidebands outside of this width $2f_c$ do not reqnire a further separation of the intercept frequencies but do require a continuation of the linear slope as far out as the sidebands are appreciable.

The next case is that of the same linear slope filters with square-law rectifiers, as shown in Fig. 2(d) and (e). The rectified voltages are taken equal to the square of the signal magnitudes:

$$E_{4\pm} = |E_{3\pm}|^2 = (\tfrac{1}{2} \pm \tfrac{1}{2}E_1)^2 = \tfrac{1}{4} \pm \tfrac{1}{2}E_1 + \tfrac{1}{4}E_1^2. \tag{18}$$

The differential output is simply

$$E_4 = E_{4+} - E_{4-} = E_1. \tag{19}$$

This is valid over an unlimited range, irrespective of the intercept frequencies of the slope filters. Therefore this type of balanced frequency detector is ideal for theoretical purposes.

The remaining case is that of square-law slope filters and linear rectifiers. The slope filters have the form of Fig. 2(d) and the differential output has the form of Fig. 2(e). The filter factors are those of (11) and (12), squared:

$$F_\pm = \left(\frac{1}{2} \pm \frac{f}{2f_c}\right)^2 = \frac{1}{4} \pm \frac{p}{2p_c} + \frac{p^2}{4p_c^2}$$

$$= \frac{1}{4} \pm \frac{1}{2p_c} D_t + \frac{1}{4p_c^2} D_t^2. \tag{20}$$

This factor requires two resonant traps at the intercept frequency in each of the slope filters. Applying this factor to E_2, the differentiated voltages are found to be

$$E_{3\pm} = \left(\frac{1}{4} \pm \frac{1}{2}E_1 + \frac{1}{4}E_1^2 + \frac{1}{4p_c}\frac{dE_1}{dt}\right) \exp ib. \tag{21}$$

The output of each linear rectifier is the magnitude

$$E_{4\pm} = \left|\frac{1}{4} \pm \frac{1}{2}E_1 + \frac{1}{4}E_1^2 + \frac{1}{4p_c}\frac{dE_1}{dt}\right|$$

$$= \left|\left(\frac{1}{2} \pm \frac{1}{2}E_1\right)^2 + \frac{1}{4p_c}\frac{dE_1}{dt}\right|. \tag{22}$$

If the modulating voltage E_1 has only slow variations, the differential output is simply E_1, as in the preceding case. However, such an assumption is not generally justified, in which case the peak value of E_1 must be held less than one, by an amount sufficient to assure the first of the last two terms exceeding the second. It is concluded that this type, because of its double differentiation, gives a linear output which is reliable over a lesser range of modulation than the first type with linear slope filters and linear rectifiers.

The theory of these three types of balanced frequency detectors indicates advantages for the first and second types. The first type, with linear slope filters and linear rectifiers, has the advantage that an undistorted replica of the modulating voltage appears in each of the rectifiers, so a departure from balance leaves no distortion of the output. This type has the disadvantage of operating over only a limited range of frequency modulation. The second type, with linear slope filters and square-law rectifiers, is ideal for theoretical purposes because it has an unlimited range of operation, and because square-law rectifiers are most susceptible of mathematical treatment. It has the practical disadvantages that square-law rectifiers are less efficient. Also they cause distortion in each rectifier so exact balance is required to secure an undistorted output.

These two types of detectors are denoted simply the linear and square-law types, referring to the rectifiers; both types have linear slope filters.

With only one signal and ideal conditions, the presence or absence of a limiter is immaterial, because the signal amplitude is uniform.

VIII. The Derivation For Two Signals in a Square-law Frequency Detector

Each of the two signals has the form of the one signal in the preceding section. The desired and undesired signals are identified by superscripts (′ and ″) as in Fig. 1.

For the study of common-channel interference, these two signals are to be superimposed. The desired signal E_2' has unit amplitude while the undesired signal E_2'' has a relative amplitude k, constant and usually less than 1. The composite signal is

$$E_2 = E_2' + E_2'' = \exp ib' + k \exp ib''. \tag{23}$$

Referring to equations (12) and (13) for the linear filter factors and the form of the differentiated signal, the composite signal becomes

$$
\begin{aligned}
E_{3\pm} &= (\tfrac{1}{2} \pm \tfrac{1}{2}E_1') \exp ib' + k(\tfrac{1}{2} \pm \tfrac{1}{2}E_1'') \exp ib'' \\
&= [(\tfrac{1}{2} \pm \tfrac{1}{2}E_1') + k(\tfrac{1}{2} \pm \tfrac{1}{2}E_1'') \exp i(b''-b')] \exp ib'.
\end{aligned}
\tag{24}
$$

The square-law rectified voltages are the squared magnitudes of the coefficient in brackets [ ], since exp ib' is a unit vector.

$$
\begin{aligned}
E_{4\pm} &= |E_{3\pm}|^2 = [(\tfrac{1}{2} \pm \tfrac{1}{2}E_1') + k(\tfrac{1}{2} \pm \tfrac{1}{2}E_1'') \cos (b''-b')]^2 \\
&\quad + [k(\tfrac{1}{2} \pm \tfrac{1}{2}E_1'') \sin (b''-b')]^2 \\
&= (\tfrac{1}{4} \pm \tfrac{1}{2}E_1' + \tfrac{1}{4}E_1'^2) + k^2(\tfrac{1}{4} \pm \tfrac{1}{2}E_1'' + \tfrac{1}{4}E_1''^2) \\
&\quad + 2k(\tfrac{1}{4} \pm \tfrac{1}{4}E_1' \pm \tfrac{1}{4}E'' + \tfrac{1}{4}E_1'E_1'') \cos (b''-b').
\end{aligned}
\tag{25}
$$

In the differential output, all except the $\pm$ terms cancel out, leaving simply

$$
\begin{aligned}
E_4 &= E_{4+} - E_{4-} \\
&= E_1' && \text{desired signal} \\
&\quad + k^2E_1'' && \text{crosstalk} \\
&\quad + k(E_1' + E_1'') \cos (b''-b') && \text{beatnote.}
\end{aligned}
\tag{26}
$$

As is characteristic of square-law rectifiers, the desired-signal output is unaffected by the presence of the undesired signal. The crosstalk output is free of distortion and proportional to k^2. The relative phase of the two carriers is noncritical.

The beatnote interference is unusual in its properties. It is a sinusoidal wave whose amplitude and frequency are modulated by both signals. Its frequency is

$$
f'' - f' = \frac{1}{2\pi} \frac{d}{dt} (b''-b') = f_c(E_1'' - E_1'). \tag{27}
$$

This is the instantaneous difference of the two frequency deviations, so it is determined by the frequency modulation of both signals. In speech modulation, the beatnote interference is heard as an irregular harsh rasping noise having only syllabic relation to the modulating signals. In wide-band modulation, the beatnote is inaudible some of the time.

Under the influence of automatic volume control (not the same as a limiter), there is manifested a blocking effect of either signal on the other, as shown in Fig. 12. The total power of the two signals is $1+k^2$, and the square-law rectifier response is proportional to the power. Dividing the output voltage (19) by this factor gives the relative amplitudes when subjected to an action which maintains uniform the average power input to the frequency detector.

$$
\begin{aligned}
\frac{E_4}{1+k^2} &= \frac{1}{1+k^2} E_1' + \frac{k^2}{1+k^2} E_1'' \\
&\quad + \frac{k}{1+k^2} (E_1' + E_1'') \cos (b'-b'').
\end{aligned}
\tag{28}
$$

The curves of Fig. 12 are plotted in terms of peak values of output for E_1' or E_2' having unit peak values. Curve (*a*) is the blocking of the desired modulation by the undesired carrier unmodulated. Curve (*b*) is the reverse effect, the crosstalk from the undesired modulation while the desired carrier is unmodulated.

While the mean or effective value of the beatnote interference is difficult to compare with the desired modulation, its peak value is interesting and can be expressed for some cases. The maximum amplitude of the beatnote occurs when the amplitude coefficient $(E_1' + E_1'')$ in (26) has its maximum value of two, and the frequency coefficient $(E_1'' - E_1')$ in (27) is small; this means when both signals have maximum modulation of the same polarity and nearly but not exactly the same amplitude. The relative peak value of beatnote interference, while both signals are modulated, is therefore independent of the bandwidth of modulation, but the relative mean value and average audibility is reduced by wider modulation. This is shown in Fig. 10(c), in which the peak amplitude of the beatnote is plotted against the beat frequency.

If only one signal is modulated, as for the curves of Fig. 12, the peak value of beatnote interference is reduced by wider modulation which increases the beat frequency beyond the range of audibility. The audibility decreases so rapidly above 5 kilocycles, that a low-pass filter may be assumed following the detector with a cutoff frequency f_a of about 5 kilocycles. Such a filter may be present, or may be approximated by a high-audio-frequency attenuator intended to compensate for emphasis in the transmitter. The maximum modulating voltage E_1 on either signal, which produces a beat frequency f_k not exceeding f_a, is found from (27) to be

$$
E_{1a} = \pm \frac{f_a}{f_c}. \tag{29}
$$

The corresponding peak voltage of the beatnote interference, multiplied by the factor of automatic volume control, is found from (28) to be

$$
\frac{k}{1+k^2} \cdot \frac{f_a}{f_c}. \tag{30}
$$

This peak value is plotted in Fig. 12 as curves (*c*) and (*d*) for the ratio $f_a/f_c = 1$ and $1/10$. This corresponds to a maximum deviation f_c one and ten times the audio frequency f_a.

For laboratory studies, sinusoidal modulation is most convenient. Therefore this case for frequency modulation deserves an explicit solution of the interference waveform. The example chosen is the unmodulated desired signal and sinusoidal modulation of the undesired signal. The modulating wave is

$$
E_1'' = m \cos \omega_m t \tag{31}
$$

in which m is the modulation factor such that the maximum deviation is mf_c. The resulting frequency deviation is, from (6),

$$
\omega'' = m\omega_c \cos \omega_m t. \tag{32}
$$

Assuming the initial phase angle b_0 of each carrier is zero in (9), the modulated phase angle is

$$b'' = \int_0^t \omega'' dt = \frac{m\omega_c}{\omega_m} \sin \omega_m t. \qquad (33)$$

From (28), the output of the balanced detector is

$$\frac{E_4}{1+k^2} = \frac{k^2}{1+k^2} m \cos \omega_m t \qquad \text{crosstalk}$$

$$+ \frac{k}{1+k^2} m \cos \omega_m t \cos b'' \qquad \text{beatnote.} \quad (34)$$

This is plotted in Fig. 5 for several cases. The desired signal is not modulated. The crosstalk output is a replica of the undesired modulating signal. The beatnote output is expressed completely as follows:

$$km \cos \omega_m t \cos \left(\frac{m\omega_c}{\omega_m} \sin \omega_m t \right). \qquad (35)$$

The beat frequency is equal to the frequency deviation given in (32), both its amplitude and its frequency being modulated at the modulation frequency. In wide-band modulation, a low-pass filter would cut off the beatnote during part of the time, while its amplitude and deviation are greatest, and the remaining beatnote interference would be weaker and more irregular. Even without the filter, it would be less audible. Therefore, while the audibility of the beatnote interference increases with increasing weak modulation, it tends to decrease with increasing strong modulation which causes the frequency deviation to go much beyond the audio-frequency range.

IX. The Derivation for Two Signals in a Linear Frequency Detector

The output of each rectifier in the linear frequency detector is the magnitude of $E_{3\pm}$ given in (24). The square of this magnitude, from (25), is

$$(E_{4\pm})^2 = |E_{3\pm}|^2 = \frac{1}{4}(1 \pm E_1')^2 \left[1 + k^2 \left(\frac{1 \pm E_1''}{1 \pm E_1'} \right)^2 \right.$$

$$\left. + 2k \left(\frac{1 \pm E_1''}{1 + E_1'} \right) \cos (b'' - b') \right]. \qquad (36)$$

Since the square root of the bracketed expression is difficult to interpret, it is expanded in a series, including only terms in the first powers of E_1' and E_1'', and up to the second power of k.

$$E_{4\pm} = \tfrac{1}{2}(1 \pm E_1') \left\{ 1 + k \left(\frac{1 \pm E_1''}{1 \pm E_1'} \right) \cos (b'' - b') \right.$$

$$\left. + \frac{k^2}{2} \left(\frac{1 \pm E_1''}{1 \pm E_1'} \right)^2 [1 - \cos^2 (b'' - b')] + \cdots \right\}$$

$$= \frac{1}{2}(1 \pm E_1') + \frac{k}{2} (1 \pm E_1'') \cos (b'' - b')$$

$$+ \frac{k^2}{8} (1 \mp E_1' \pm 2E_1'' \mp 2E_1'E_1'')$$

$$[1 - \cos (2b'' - b')] + \cdots. \qquad (37)$$

The differential output is

$$E_4 = E_{4+} - E_{4-}$$

$$= E_1' \left(1 - \frac{k^2}{4} + \cdots \right) \qquad \text{desired signal}$$

$$+ \frac{k^2}{2} E_1'' \qquad \text{crosstalk}$$

$$- \frac{k^2}{2} E_1'E_1'' \qquad \text{mixed signal}$$

$$+ \frac{k}{2} E_1'' \cos (b'' - b') \qquad \text{beatnote}$$

$$+ \frac{k^2}{4} (E_1' - 2E_1'' + 2E_1'E_1'' - \cdots)$$

$$\cos (2b'' - b') + \cdots. \qquad (38)$$

Two cases of this expansion appear as (3) and (4). It is complicated by the masking factor on the desired signal, and by the distortion (harmonics and intermodulation) of both signals. Each coefficient is part of an infinite series.

The automatic volume control in this case is also assumed to have a linear rectifier, to correspond with the detector. This control is substituted for the limiter L in Fig. 1, and is assumed to maintain the average value of the voltage E_3 at a uniform value of unity. Rewriting (23), the composite signal voltage is

$$E_2 = \exp ib'[1 + k \exp i(b'' - b')] \qquad (39)$$

and its magnitude is

$$|E_2| = \sqrt{1 + k^2 + 2k \cos (b'' - b')}$$

$$= 1 + 2k \cos (b'' - b')$$

$$+ \frac{k^2}{4} [1 - \cos 2(b'' - b')] + \cdots. \qquad (40)$$

The average value of its magnitude is

$$\overline{E_2} = \frac{2}{\pi}[2E(k) - (1 - k^2)K(k)] = 1 + \frac{k^2}{4} + \cdots$$

$$= \frac{4}{\pi} \quad (k = 1)$$

$$= k \quad (k \gg 1) \qquad (41)$$

in which $K(k)$ and $E(k)$ are respectively the complete elliptic integrals of the first and second kinds.[11-13] The

[11] (The real part of K or E is used for $k > 1$.)
[12] Jahnke and Emde, "Tables of Functions," 1933, chapter 15, pp. 127, 145, 150.
[13] D. Bierens de Haan, "New Tables of Definite Integrals," 1867–1939. Table 67, (5) and (7); uses F' and E' instead of K and E.

value of E_3 becomes

$$E_3 = \frac{E_2}{\overline{E_2}} \tag{42}$$

which has an average value of unity.

The desired-signal output is reduced by two factors, the masking factor (in parentheses) (38), and the factor $1/\overline{E_2}$. Both factors cause more reduction with increasing relative amplitude k of the undesired signal.

A precise evaluation of the differential output E_4 also requires elliptic integrals. Instead of expanding into a series, formula (36) may be rewritten,

$$E_{4\pm} = \tfrac{1}{2}(1 \pm E_1')\sqrt{1 + k_\pm^2 + 2k_\pm \cos(b'' - b')} \tag{43}$$

in which

$$k_\pm = k\,\frac{1 \pm E_1''}{1 \pm E_1'}\,. \tag{44}$$

In order to remove the beatnote, $E_{4\pm}$ is averaged over the beatnote cycle of $(b''-b')$, leaving only the modulating voltages as they appear in the output of either rectifier.

$$\overline{E_{4\pm}} = \frac{1}{2}(1 \pm E_1')\,\frac{2}{\pi}\,[2E(k_\pm) - (1-k_\pm^2)K(k_\pm)]. \tag{45}$$

This is obtained by the same elliptic integral as used to obtain (41) from (40) above. The differential output $\overline{E_4}$ is to be expressed only for special cases.

The distortion of the modulating voltages E_1' and E_1'' in the output complicates the solution, but adds little of interest because the more important effects involve the beatnote noise and the signal amplitudes. The latter information is obtained with close approximation by solving for the case of small modulating voltages:

$$E_1' \ll 1,\ E_2' \ll 1,$$

$$k_\pm = k, \qquad \frac{dk_\pm}{dE_1'} = \mp k, \qquad \frac{dk_\pm}{dE_1''} = \pm k. \tag{46}$$

The relation is used,[14]

$$\begin{aligned} \frac{d}{dk}\cdot\frac{2}{\pi}\,[2E(k)-(1-k^2)K(k)] &= \frac{2}{\pi k}\,[E(k)-(1-k^2)K(k)] \\ &= \frac{k}{2} \quad (k \ll 1) \\ &= \frac{2}{\pi} \quad (k = 1) \\ &= 1 \quad (k \ll 1). \end{aligned} \tag{47}$$

On this basis, the differential output of the detector (omitting the beatnote terms) is

[14] Jahnke and Emde, footnote reference 12, pp. 128–129.

$$\begin{aligned} \overline{E_4} = \overline{E_{4+}} - \overline{E_{4-}} &= \left(E_1'\frac{d}{dE_1'} + E_1''\frac{d}{dE_1''}\right)(E_{4+} - E_{4-}) \\ &= E_1'\,\frac{2}{\pi}\,E(k) \quad \text{desired signal} \\ &\quad + E_1''\,\frac{2}{\pi}\,[E(k) - (1-k^2)\,K(k)] \quad \text{crosstalk.} \end{aligned} \tag{48}$$

Applying the automatic-volume-control factor, the resulting output is

$$\begin{aligned} \frac{\overline{E_4}}{\overline{E_2}} &= E_1'\left(1 - \frac{E(k)-(1-k^2)K(k)}{2E(k)-(1-k^2)K(k)}\right) \quad \text{desired signal} \\ &\quad + E_1''\,\frac{E(k) - (1-k^2)K(k)}{2E(k) - (1-k^2)K(k)} \quad \text{crosstalk.} \end{aligned} \tag{49}$$

This formula shows that the reduction of response to the desired signal is equal to the increase of response to the undesired signal. This crosstalk is plotted as Fig. 13(*b*).

In a like manner, the beatnote fundamental and harmonic components can be evaluated in terms of elliptic integrals. For present purposes, however, the series expansion (38) is more useful.

X. The Derivation for Two Signals in a Limiter

The limiter L in Fig. 1 receives the composite signal voltage E_2 of varying amplitude. The action of the limiter on this voltage is such that its output voltage retains the instantaneous frequency of the input voltage but has a uniform amplitude of unity. This action is conceived as a fast-acting control of amplification which holds the signal envelope at a constant amplitude.

The composite input voltage is

$$\begin{aligned} E_2 = E_2' + E_2'' &= \exp ib' + k \exp ib'' \\ &= \exp ib'[1 + k \exp i(b''-b')] \\ &= \exp ib'[1 + k\cos(b''-b') + ik\sin(b''-b')] \\ &= |E_2| \exp ib = |E_2| \exp ib' \exp i(b''-b') \end{aligned} \tag{50}$$

in which b is the progressive phase angle of the composite signal and $|E_2|$ is the magnitude of the envelope. This phase angle is determined by the above expressions:

$$\tan(b - b') = \frac{k\sin(b''-b')}{1 + k\cos(b''-b')} \tag{51}$$

$$b = b' + \text{antitan}\,\frac{k\sin(b''-b')}{1 + k\cos(b''-b')}\,. \tag{52}$$

Therefore the frequency of the composite signal is

$$\begin{aligned} \omega = \frac{db}{dt} &= \frac{db'}{dt} + \frac{k^2 + k\cos(b''-b')}{1+k^2+2k\cos(b''-b')}\left(\frac{db''}{dt} - \frac{db'}{dt}\right) \\ &= \omega' + k(\omega'' - \omega')\,\frac{k + \cos(b''-b')}{1 + k^2 + 2k\cos(b''-b')}\,. \end{aligned} \tag{53}$$

The output of the limiter is

$$E_3 = \frac{E_2}{|E_2|} = \exp ib \tag{54}$$

which is a signal of unit amplitude and of phase b or frequency ω.

While this signal could be subjected to the processes in a frequency detector, this is unnecessary, because the detector in this case would merely deliver an output voltage proportional to the frequency (deviation). Therefore the differential output of the detector is

$$E_4 = \frac{\omega}{\omega_c} \cdot \tag{55}$$

This was proved for a signal of unit amplitude and any frequency modulation in the case of one signal; it is a conclusion from (17) and (19) in that case. In this case, the detector output is

$$E_4 = E_1' \qquad \text{desired signal}$$

$$+k(E''-E')\frac{k+\cos(b''-b')}{1+k^2+2k\cos(b''-b')} \quad \text{beatnote.} \tag{56}$$

If the undesired signal is the weaker, $k<1$, the second term has an average value of zero, and therefore represents only the beatnote (fundamental and harmonics) as indicated. On the other hand, if the undesired signal is the stronger, $k>1$, it amounts to interchanging the two signals and inverting k:

$$E_4 = E_1'' \qquad \text{crosstalk}$$

$$+\frac{1}{k}(E_1'-E_1'')\frac{\frac{1}{k}+\cos(b'-b'')}{1+\frac{1}{k^2}+\frac{2}{k}\cos(b'-b'')} \quad \text{beatnote.} \tag{57}$$

These two solutions show clearly the effect of the limiter in favoring the stronger of the two signals and eliminating the other, as plotted in Fig. 11(a) and (b). The beatnote is a harmonic series of the following form for $k<1$ (and corresponding form[15–17] for $1/k<1$):

$$k\frac{k+\cos(b''-b')}{1+k^2+2k\cos(b''-b')} = k\cos(b''-b')$$

$$-k^2\cos 2(b''-b') + k^3\cos 3(b''-b') - \cdots. \tag{58}$$

This beatnote series is included in (2) and the waveform is plotted in Fig. 4(c) on a scale which leaves the fundamental component the same for various values of k.

[15] Crosby (footnote 4) in equation (7) and Fig. 4 shows the harmonic waveform of the beatnote.

[16] E. P. Adams, "Smithsonian Mathematical Formulae," 1922. On p. 82, no. 13 is the expansion needed to express the beatnote in a Fourier series.

[17] de Haan's table 50, item (5), is the integral form needed to evaluate the coefficients of the beatnote Fourier series.

The beatnote oscillates between the peak values

$$(k<1)\ \frac{k}{1+k} \text{ and } -\frac{k}{1-k};\quad (k>1)\ \frac{1}{k+1} \text{ and } -\frac{1}{k-1}\cdot \tag{59}$$

Its quadratic-mean (root-mean-square) value is

$$=\sqrt{\tfrac{1}{2}(k^2+k^4+k^6+\cdots)} = \frac{k}{\sqrt{2(1-k^2)}} \quad (k<1). \tag{60}$$

An interesting example is that of no modulation of the desired signal and sinusoidal modulation of a weaker undesired signal. Only the beatnote term remains:

$$E_4 = kE_1''\frac{k+\cos b''}{1+k^2+2k\cos b''} \quad (k<1) \tag{61}$$

in which

$$b'' = \frac{m\omega_c}{\omega_m}\sin\omega_m t;\quad \omega'' = \frac{db''}{dt} = m\omega_c\cos\omega_m t. \tag{62}$$

In these formulas, $m\omega_c$ is the maximum deviation, ω_m is the modulating frequency, and ω'' is the beat frequency. This output voltage E_4 is plotted in Fig. 6.

XI. Conclusion

There are several types of balanced frequency detectors capable of reproducing without distortion a frequency-modulated signal. Two of these types employ linear slope filters, one with linear rectifiers and the other with square-law rectifiers.

If a perfect limiter is assumed, it is immaterial which type of frequency detector is used; otherwise the choice depends on which type has the more desirable behavior toward interference. Toward common-channel interference, the linear rectifiers give less crosstalk from the weaker of the two signals. During modulation of the stronger signal, the linear rectifiers give less beatnote interference.

The different kinds of interference are subdivided into frequency and amplitude effects. The frequency effects are inherently associated with the ability to detect the frequency modulation of the signals. The amplitude effects can be avoided by the use of a limiter; they are twice as great with square-law rectifiers as with linear rectifiers.

The beatnote interference is caused by frequency and amplitude modulation in the composite signal. Its interference effect is mainly caused by the frequency modulation, so it cannot be reduced very much by avoiding the amplitude effects. It can be reduced by using a bandwidth of frequency modulation exceeding twice the bandwidth required by the modulating signal.

The crosstalk interference from a weaker signal is caused entirely by the amplitude modulation in the composite signal. It can be minimized by the use of a limiter, and its amplitude from linear rectifiers is only one half as great as from square-law rectifiers.

The conical pattern, especially well adapted for oscilloscope tests, is useful in observing common-channel interference and in comparing the frequency modulation of one signal with the steady frequency of another.

The relations described have been verified by tests. The waveform of the interference output has been found to agree with Figs. 5 and 6, in corresponding cases without and with a limiter. The conical diagrams of interference output, Figs. 7 and 8, have been reproduced on the oscilloscope under the various conditions illustrated. The conical diagram of Fig. 9 has been used in checking the amount of frequency modulation in a signal. The crosstalk component has been separated from the beatnote by a filter to test the relations of Fig. 13, both without a limiter and with limiters of practical design. The beatnote amplitude, as in Figs. 11 and 12, has been checked by oscilloscope observations corresponding to Figs. 5 to 8. The nature of the crosstalk and beatnote interference has been verified in listening tests.

Institute News and Radio Notes

1942 Winter Convention

The technical sessions of the Winter Convention to be held on January 12, 13, and 14 at the Commodore Hotel in New York City will be high-lighted by four addresses of major importance bearing on radio's expanding role in the prosecution of the war.

At the opening session on Monday morning, January 12, Dr. F. B. Jewett. Chairman of the Board of Bell Telephone Laboratories, Vice-President of the American Telephone and Telegraph Company, and now at the head of the communication division of the National Defense Research Committee, will speak on: "The Mobilization of Science with Special Reference to Communication." Facing in combat the German genius for organizing technical forces, the United States is systematically drawing on superior scientific and engineering resources in the country's commercial laboratories and universities. How the talent engaged in radio and the other electronic and communication industries is being focused on the problems of war will be the topic of Dr. Jewett's significant address.

Lieutenant Colonel Rex V. D. Corput, Jr., Director of the United States Signal Corps Laboratories at Fort Monmouth, New Jersey, will outline the functions of his organization at the Wednesday afternoon session. Recent publicity has been given to the development, at those laboratories, of apparatus for intercommunication among squadrons of tanks in action. Since such developments constitute radio's front-line attack on the problems of adapting modern devices to requirements of the Army in action, Colonel Corput's address on the scope of activities at the Signal Corps' leading laboratory will be of special interest.

The principal speaker at the banquet, Tuesday evening, will be Mr. Don Francisco, Director of Communications on the staff of Mr. Nelson A. Rockefeller, Coordinator of Inter-American Affairs. Mr. Francisco is an ace advertising executive, former President of Lord and Thomas, who recently made a whirlwind tour of Latin America to find out—to quote *Time*—"Why the United States usually puts its wrong foot forward." Not only does he control the battery of United States short-wave radio broadcasting stations serving Central and South America and the West Indies, but he also superintends the production of American motion pictures sent to South American countries. In Mr. Francisco we have secured at once a forceful speaker and a man who is sure to bring the Institute a message the importance of which transcends the usual boundaries of engineering thought.

Also addressing the banquet will be Mr. Adolfo T. Cosentino, Director of Radio Communications of the Argentine, who is traveling from South America to attend the convention. Mr. Cosentino will address the convention with particular reference to the South American viewpoint on radio and its influence. He is the 1941 Vice-President of the Institute and since he is both a radio engineer and a public official is in a good position to speak informatively and interestingly on the subject.

Among the novel features of this convention may be mentioned the following:

An unusual group of exhibits, in which the current interest in the availability of regular and substitute materials will be reflected.

The reception at which President and Mrs. Van Dyck will meet the members of the Institute and women guests of the convention on Monday evening.

Dr. Baker Honored at Rochester Fall Meeting

Walter R. G. Baker (A'19-F'29), who was recently named Vice-President in charge of radio and television of the General Electric Company, was presented with a gold escutcheon plaque in recognition of his work as Chairman of the National Television Systems Committee. The award was presented at the banquet during the Rochester Fall Meeting on November 11 by Institute Past-President L. C. F. Horle(at right), assisted by Virgil M. Graham, Chairman of the Rochester Fall Meeting Committee.

Dr. Baker was born in Lockport, New York, in 1890. On his graduation from Union College he joined the General Electric research laboratories in 1917. During the war, he worked on vacuum-tube transmitters and receivers for military service. In 1924, he was placed in charge of the design of all radio products and, two years later, his responsibility was increased to include their development, design, and production.

In 1929, he took charge of the radio engineering activities of the newly formed RCA-Victor Corporation at Camden, New Jersey. Within a year he was placed in charge of production and then became general manager of the plant.

He returned to the General Electric Company in 1935 as managing engineer of the radio receiver activites in Bridgeport, Connecticut. Four years later he was placed in charge of the radio and television department.

The honorary degree of D.S. was conferred on him by Union College in 1935.

Dr. Baker has served as Director of the Engineering Department of the Radio Manufacturers Association for a number of years.

His services to the Institute include membership on the Board of Directors, the holding of offices in both the Philadelphia and Connecticut Valley Sections, and contributions to the Proceedings.

FORTHCOMING MEETINGS

Winter Convention
New York, N. Y.
January 12, 13, and 14, 1942

Summer Convention
Cleveland, Ohio
June 29, 30, and July 1, 1942

The buffet dinner and party on the same evening where an opportunity will be afforded the members to mix and become acquainted.

Columbia's demonstration of color television, over the air, at the Monday evening technical session.

The Dutch-treat luncheons on Monday and Tuesday, for the purpose of giving out-of-town members a definite opportunity to become acquainted with New York members.

The frequent bus trips to beautiful Alpine, New Jersey, on the Palisades of the Hudson, all day Tuesday, to visit Major Armstrong's frequency-modulation broadcasting station.

The opportunity to renew acquaintances with several Past Presidents of the Institute who, in connection with the celebration of our Thirtieth Anniversary, will be present at the banquet and will preside at some of the technical sessions.

The college technical session on Wednesday morning, open to all registrants at the Convention, at which students from near-by colleges will be addressed by various members of the Institute in industry. Among others, Mr. J. V. L. Hogan will speak on "Modern Techniques in Broadcasting"; Mr. B. J. Thompson on "Modern Developments in Electronics"; and Mr. J. H. Hackenberg will demonstrate Western Union's facsimile equipment.

Of chief interest of course are the papers to be presented at the technical sessions. Produced under wartime conditions there is a possibility that adjustments will have to be made at the last minute in this program.

Board of Directors

A regular meeting of the Board of Directors was held on December 3, 1941, and attended by Haraden Pratt, chairman; Alfred N. Goldsmith, editor; Austin Bailey, A. B. Chamberlain, I. S. Coggeshall, W. L. Everitt, guest; H. T. Friis, Virgil M. Graham, O. B. Hanson, R. A. Heising, C. M. Jansky, Jr., F. B. Llewellyn, B. J. Thompson, H. M. Turner, H. A. Wheeler, L. P. Wheeler, and H. P. Westman, secretary.

A report showing a total registration of 520 at the Rochester Fall Meeting held on November 10, 11, and 12 was received.

A report was received on the Institute office operation. Authorization was given for the employment of an office manager who will not only relieve the Secretary of certain routine work but will have sufficient time to make a reasonably accurate accounting of our work activities and requirements.

Removal of the Institute office to another location in the same building at the request of the building management was authorized.

General agreement was reached in the development of a plan whereby the Institute would co-operate more actively with Professor Everitt of the Ohio State University in the holding of the yearly broadcast engineering conferences.

Vladimir K. Zworykin
Recipient of Rumford Award

Vladimir K. Zworykin (M '30-F '38), Associate Director of RCA Laboratories, received on October 8 the Rumford Award of the American Academy of Arts and Sciences. Specifically recognized in the award was the research work in the development of the electron microscope. Reports were presented to the Institute membership on this work both at meetings and through the Proceedings.

Dr. Zworykin was born in Mourom, Russia. At the Petrograd Institute of Technology he did his first experimental work on the transmission of images by radio. After graduating in 1912, he spent two years in X-ray research at the College d'France. He served as an officer in the Signal Corps of the Russian army during the World War.

In 1920, he joined the research staff of the Westinghouse Electric & Manufacturing Company at Pittsburgh. Since 1929 he has been associated with the RCA laboratories and has been engaged in electronic research.

He has contributed regularly to the Institute's meeting and Proceedings. *In 1934, the Morris Liebmann Memorial Prize was presented to him.*

He received the degree of Ph.D. from the University of Pittsburgh in 1926 and that of D.Sc. from the Brooklyn Polytechnic Institute in 1938. The Institution of Electrical Engineers awarded him its Overseas Premium in 1938. He received also the American Pioneer Award in 1940.

Executive Committee

The Executive Committee met on December 1 and those present were Haraden Pratt, vice chairman; Alfred N. Goldsmith, editor; R. A. Heising, B. J. Thompson, A. F. Van Dyck, guest; and H. P. Westman, secretary.

A number of items concerning the Institute office and our solicitation of advertising for the Proceedings were considered.

Approval was granted of 85 applications for election to Associate, 3 for Junior, 68 for Student, and 4 for transfer to Associate grades.

A meeting of the Executive Committee was held on December 3 and was attended by Haraden Pratt, vice-chairman; R. A. Heising, B. J. Thompson, and H. P. Westman, secretary.

A number of items which were considered by the Board of Directors at its meeting on December 3 were examined and in some cases recommendations made to the Board.

A meeting of the Executive Committee was held on December 17 and was attended by Haraden Pratt, vice-chairman; Alfred N. Goldsmith, editor; R. A. Heising, B. J. Thompson, A. F. Van Dyck, and H. P. Westman, secretary.

The major business at this meeting was the preparation of a preliminary budget to be submitted to the Board of Directors and the development of a list of committee personnel, also to be distributed to the Board of Directors in advance of the annual meeting of that body.

High-Frequency Radio Transmission Conditions

In compliance with a request from competent authority, the National Bureau of Standards will no longer make public information on radio transmission conditions and the ionosphere. The Institute will, therefore, until further notice, discontinue publication of the monthly report, "High-Frequency Radio Transmission Conditions."

U.R.S.I.—I.R.E. Meeting

Before 1941 an annual joint meeting of the Institute and the American Section of the International Scientific Radio Union (U.R.S.I.) was held in Washington in April, at which papers on the more fundamental and scientific aspects of radio were presented. The meeting will not be held this year, as was true also last year because of the general absorption of laboratories and engineers in defense activities.

Committee Meetings

Annual Review—December 3
Annual Review, Co-ordinating—November 25
Facsimile, Subcommittee on Fundamentals—November 28
Membership—December 2

Committee and Section Reports

In the answers to the recent questionnaire sent to the membership of the Institute, the members indicated their preference for *more space devoted to papers*. The membership approved reduction in space devoted to other matters, including information on section and committee meetings.

Accordingly, the new policy desired by the membership has been put into immediate force. This issue of the PROCEEDINGS contains one more paper than was originally planned.

Section Meetings

ATLANTA

Moving-picture demonstration of the "RCA Television System and Electron Microscope," by D. A. Reesor, Radio Corporation of America, October 24.

"Frequency Modulation" by L. D. Chesnut, General Electric Company, November 14.

BALTIMORE

"The New WBAL 50-Kilowatt Broadcast Transmitter", by R. N. Harmon, Westinghouse Electric Manufacturing Company, October 17.

"Color Television," by J. N. Dyer, Columbia Broadcasting System, November 28.

BUFFALO-NIAGARA

"Technical Installation of the New Buffalo Broadcasting Stations," by R. J. Kingsley, WEBR-WBEN, and K. B. Hoffman, WKBW-WGR, November 19.

"Great Lakes Radiotelephone," by W. O. Todd, Radiomarine Corporation of America, December 10.

CHICAGO

"Color Television," by P. C. Goldmark, Columbia Broadcasting System, October 17.

CINCINNATI

"Stereoscopic Kodachrome Photography," by Elmer Nuezel, Cincinnati Gas and Electric Company, November 11.

"Lightning," by K. B. McEachron, General Electric Company, November 11.

CLEVELAND

"Inductive Tuning," by B. V. K. French, P. R. Mallory & Company, November 27.

CONNECTICUT VALLEY

"The Present Status of Television," by G. W. Fyler, General Electric Company, October 16.

DALLAS-FORT WORTH

"Frequency Modulation," by R. E. Beam, Southern Methodist University, December 1.

INDIANAPOLIS

"Training of Radio Men For the United States Navy," by Boyd Phelps, Naval Training School, November 28.

"Resistance-Tuned Oscillators," by A. J. Drazy, Purdue University, November 28.

PHILADELPHIA

"Recent Developments in Electromagnetic Relays as Applied to Radio Transmitters," by C. A. Packard, Struthers Dunn, Inc., November 6.

"Alternate Materials in Radio," by R. L. Daggett and M. J. Obert, RCA Manufacturing Company, December 4.

PITTSBURGH

"W47P, Pennsylvania's First Frequency-Modulation Station," by H. R. Kaiser, WWSW and W47P, November 10.

PORTLAND

"Some Aspects of Mercury Vapor," by Marion Morris, Horseheaven Mines, Inc., October 9.

"A Simplified Graphical Method of Designing Impedance-Matching Networks," by R. M. Walker, KOMO-KJR, October 30.

"Industrial Radio-Frequency Power Applications," by W. L. Atwood, Thermal Engineering Company, November 5.

SAN FRANCISCO

"Design and Adjustment of Broadcast Antenna Arrays," by Norman Webster, McClatchy Broadcasting Company, November 14.

SEATTLE

"Development in Radio Aids to Navigation," by P. C. Sandretto, United Air Lines, March 18.

"High-Frequency Communication Over Short Distances in High Northern Latitudes," by R. J. Gleason, Pan American Airways, June 6.

"Recent Advances in the Electron Theory of Photography," by P. M. Higgs, University of Washington, September 19.

TORONTO

"Fluorescent Lamps and Lighting Materials," by Harris Reinhardt, Hygrade-Sylvania Corporation, December 8.

TWIN CITIES

"WCCO's Audio-Frequency Facilities," by A. G. Peck, WCCO, November 19.

"Impedance Measurements from 1 to 100 Megacycles," by R. F. Field, General Radio Company, December 3.

WASHINGTON

"Design of Loop Antennas as Used On Aircraft," by George Levy, United Air Lines Transportation Corporation, December 8.

Membership

The following admissions or transfers (where indicated as such) to Associate grade were approved by the Board of Directors on December 3, 1941.

Adelman, H., 64-58 Alderton St., Forest Hills, L. I., N. Y.

Ahern, W. R., Radio Transmitter Engineering Department, General Electric Co., Schenectady, N. Y.

Alberts, W. S., 6430 McHugh Pl., Cincinnati, Ohio

Angst, D. C., 1933 Illinois, Vallejo, Calif.

Bain, L. S., 140 Cornwall Ave., Town of Mt. Royal, Que., Canada

Barnard, S. B., Hygrade Sylvania Corporation, Emporium, Pa.

Berger, U. S., Bell Telephone Laboratories, Inc., 180 Varick St., New York, N. Y.

Berry, V. W., 4554 Toland Way, Los Angeles, Calif.

Blackman, C. P., 2443 N. Powhatan St., Arlington, Va.

Branscome, E. O., 5501 Cornell Ave., Chicago, Ill.

Brar, S. S., 5540 S. Woodlawn Ave., Chicago, Ill. (Transfer)

Brereton, C. H., 306 Linwood St., St. James, Winnipeg, Manit., Canada

Brown, C. H., 2626 Kirkham St., San Francisco, Calif.

Cameron, F., 113 High St., Nelson, B. C., Canada

Carment, C. A., 6906 Forest Ave., Brooklyn Postal Station, Cleveland, Ohio

Carson, V. S., University of Connecticut, Storrs, Conn. (Transfer)

Chatterjee, S. K., 91 Ballygunge Pl., Calcutta, Bengal, India

Cook, L. V., Heiland Research Corp., 1731 Arapahoe St., Denver, Colo.

Cowperthwait, H. S., Box 1972, Atlanta, Ga. (Transfer)

Cummings, M. M., c/o Library, Naval Training Station, Newport, R. I.

Davidson, R., Coast Artillery School, Fort Monroe, Va.

Demeter, A. A., 34 Classic Ave., Toronto, Ont., Canada

Dietz, H. G., 66-34 Selfridge St., Forest Hills, L. I., N. Y.

Disbral, M., 201 Linden Blvd., Brooklyn, N. Y.

Droz, G., Ste. Croix (Vaud) Suisse, Switzerland

Dunlap, C. R., 1 Woodview Ave., Long Branch, N. J.

Edge, J., 25 Howitt Rd., London, N.W.3, England

Eitelbach, H. L., 562 Winterburn Grove, Grantwood, N. J.

Felch, E. D., 3227 W. 78th St., Los Angeles, Calif.

Ferrell, P., Jr., 107 E. Bayview Ave., Pleasantville, N. J.

Forman, J., 137 W. 5th St., Emporium, Pa.

Funderburg, J., 800 Sunset Blvd., Los Angeles, Calif.

Galetz, R. J., 7830 Karlov Ave., Skokie, Ill.

Gauss, H., 1428 Madison St., N.W., Washington, D. C.
Gran, R. M., 6420—8th St., N.W., Washington, D. C.
Gray, F., c/o Bell Telephone Laboratories, Inc., 463 West St., New York, N. Y.
Grousset, W., 3234 Sinclair, Buenos Aires, Argentina
Haase, W. D., Radio Station KVI, Rust Bldg., Tacoma, Wash.
Hare, A. C., 3110 Esther Pl., Anacostia, Washington, D. C.
Harvey, N. L., 120 E. 4th St., Emporium, Pa.
Hedrick, K. M., 608 Lincoln Apartments, St. Paul, Minn.
Herzog, E. F., Radio Engineering Department, General Electric Co., Schenectady, N. Y.
Hood, J. T., 4816 N. Illinois St., Indianapolis, Ind.
Hopkinson, T. W., National Youth Administration, Manassas, Va.
Hough, J. M., 1845 Indiana St., Vallejo, Calif.
Hudson, P. K., University of Arkansas, Fayetteville, Ark.
Hulst, G. D., Jr., RCA License Laboratory, 711 5th Ave., New York, N. Y.
Hutchinson, R. W., 226 S. Prairie St., Charlotte, Mich.
Johnson, H. D., 396 E. 5th St., Emporium, Pa.
Jurovetzky, M., Ave. Constituyentes 4379, Buenos Aires, Argentina
Kahn, J. J., 1500 N. Halsted St., Chicago, Ill.
Kerstetter, D. R., 324 W. 4th St., Emporium, Pa.
King, L. E., District Communication Office, Naval Operating Base, Norfolk, Va.
Lambert, J., 505 W. 183rd St., New York, N. Y.
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Lonsdale, E. M., Electrical Engineering Annex, Iowa State College, Ames, Iowa
Lundwall, G. F., R.F.D. 1, Hempstead, N. Y.
Maas, W. C., 53 S. Howard Ave., Bellevue, Pa.
Macdonald, F. C., 22 Williams Lane, Chevy Chase, Md.
Mayberry, H. P., Cameron, Pa.
Mayer, H. J., 805 Bluff Ave., Sheboygan, Wis.
Mayer, R. N., Civil Aeronautics Administration, Anchorage, Alaska (Transfer)
McClenon, D., 112 N. Columbus, Arlington, Va.
McKnight, C. B., 3305 Wood Ter., Los Angeles, Calif.
Meadows, L. A., 5910 2nd Pl. N. W., Washington, D. C.
Miller, E., West End, N. C.
Morrison, E. D., Trade Schools, Naval Air Station, Jacksonville, Fla.
Murphy, S. A., Jr., 2909 1st Ave., San Diego, Calif.
Nash, A. W., 719 S. Citrus Ave., Los Angeles, Calif.
Nash, F. M., 2335 Washington St., San Francisco, Calif.
Ober, H. E., 947 Melaview Rd., Cleveland, Ohio
Osmundsen, G. A., 410 Nelson Ave., Grantwood, N. J.
Palevsky, H., Naval Ordnance Laboratory, Navy Yard, Washington, D. C.
Quest, R., c/o J. Bosler, Smith Cove, Waterford, Conn.
Randolph, F. F., 603 Prindle St., Sharon, Pa.
Reitz, N. J., Hygrade Sylvania Corporation, Emporium, Pa.
Rosselot, G. A., Georgia School of Technology, Atlanta, Ga.
Rowley, P., c/o Mrs. W. Rowley, Great Knighton Farms, Knighton Nr. Alcester, Warwickshire, England
Russell, C. C., 19991 Upper Valley Dr., Euclid, Ohio
Ruth, J. A., 9 Leona St., Endicott, N. Y.
Schuler, R. M., 3155 Alta Ave., Fresno, Calif.
Sepmeyer, L. W., 1843 Fox Hills Dr., Los Angeles, Calif.
Sowerby, J. M., Croft House, Salisbury, Wilts., England
Stoops, C. W., 1803 Monroe St., N. W., Washington, D. C.
Sturdy, E. W., 1471 Crest Dr., Los Angeles, Calif.
Van Ryn, K. C., 15 Little Farms Rd., Larchmont, N. Y.
Wassa, G. M., 358 Wadsworth Ave., New York, N. Y.
Wehner, G. L., 57 E. Dixon Ave., Dayton Ohio
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Contributors

Richard L. Campbell

Richard L. Campbell was born on January 12, 1907. He received the B.S. degree in electrical engineering from Iowa State College in 1929. From then until January, 1930, he was employed by the radio department of the General Electric Company. From 1930 to 1933 Mr. Campbell was associated with the television research division of the RCA Victor Company in Camden, New Jersey. In 1933 he went with the research department of the Philco Radio and Television Corporation in Philadelphia and was with them until 1938. Since then, he has been with the Allen B. Du Mont Laboratories, Inc., as a transmission equipment engineer.

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Francis E. Fox was born May 22, 1909, in Wilmington, Delaware. He received the A.B. degree in 1932 at the Catholic University of America, and in 1934 the M.Sc. in physics from the same university. From 1936 he was associated with the physics department at the Catholic University as research assistant, where he received the Ph.D. degree in 1937. Since 1939 Father Fox has been instructor in physics at the Catholic University, and his research has been chiefly in the field of ultrasonics.

Francis E. Fox

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Cecil E. Haller (A'34-M'40) was born at Houston, Ohio, on January 15, 1908. He received the A.B. degree from Ohio

Cecil E. Haller

Wesleyan University in 1930 and the M.S. degree from the University of Pittsburgh in 1934. From 1930 to 1932 he was a member of the graduate student course at the Westinghouse Electric and Manufacturing Company, East Pittsburgh, Pennsylvania, and from 1932 to 1934 he was a research engineer for the same organization. From 1934 to 1936 Mr. Haller was employed as an engineer at the Ken-Rad Corporation, Owensboro, Kentucky. In 1936 he joined the Research and Engineering Department of the RCA Manufacturing Co., Inc., at Harrison, New Jersey.

Robert E. Kessler

Robert E. Kessler was born in East Orange, New Jersey, on November 13, 1914. He attended grade and high school at Glen Ridge, New Jersey, and took radio courses in New York City. In 1932 he obtained an amateur radio license and later a commercial broadcast-station license. In 1936 Mr. Kessler was employed by the Allen B. Du Mont Laboratories, Inc., where he is an operator of television stations W2XVT and W2XWV and is doing development work on transmission equipment.

James B. Knox (A'34) was educated at the University of Manitoba. He was with the China Airways, Shanghai, China, from 1929 to 1930 and the Asia Electric Company, Shanghai, from 1933 to 1935. From 1935–1937 Mr. Knox was employed by Standard Telephones and Cables, Ltd., London, England, and from 1937 to 1941 he was with the Radio Division, Department of Transport, Ottawa, Ontario, Canada. Since 1941 he has been

James B. Knox

with the engineering products division of the RCA-Victor Company, Ltd., Montreal, Canada.

Klaus U. Landsberg studied electrical engineering at the Polytechnical Institutes of Bodenbach, Czechoslovakia, and Berlin, Germany, and received an E.E. degree and a degree in communications and high-frequency engineering. While in Europe he worked as a research and development engineer in Dr. Korn's picture-telegraphy and television laboratories. In 1937 he practiced as a consulting engineer in New York and was associated with Farnsworth Television, Inc., as a television studio development engineer. Directly following this position Mr. Landsberg was employed as a television development engineer and as an operating and maintenance engineer by the National Broadcasting Company, television department, in New York City. He went with Du Mont Laboratories, Passaic, New Jersey, in the spring of 1940 and worked there as transmitter development engineer until his recent appointment as chief engineer of Television Productions, Inc., a subsidiary of Paramount Pictures, Hollywood, California.

Edmund A. Laport (A'25-M'27) was born at Nashua, New Hampshire, on July 2, 1902. He was a commercial radio operator at KDKF, New York City, in 1921, and a receiver service engineer for the Westinghouse Electric and Manufacturing Company in 1922. In 1923, he was a laboratory assistant in the radio

Klaus U. Landsberg

engineering department of the General Electric Company, working on transmitter development. From 1924 to 1932 Mr. Laport was a transmitter engineer with the Westinghouse Electrical and Manufacturing Company. He installed three high-frequency communication stations for the Chinese Ministry of Communications, Peking, China, in 1928, and two 50-kilowatt broadcast stations at Rome and Milan, Italy, 1929–1930 and 1932. He also installed several broadcast stations from 1 to 50 kilowatts in the United States and Canada. During 1933 and 1934 he was associated with Paul F. Godley as consulting engineer. From 1934 to 1936 he was

Edmund A. Laport

employed as a transmission engineer for Wired Radio, Inc., working on variable and suppressed-carrier assymmetric-sideband transmission development. Since 1936 he has been a section engineer, working in high-power transmitters with the RCA Manufacturing Company at Camden, New Jersey. Mr. Laport was transferred to Montreal in December, 1938, as manager of the engineering and development engineering products division of the RCA Victor Company, Ltd.

George D. Rock

George D. Rock was born at Bridgeport, Connecticut on September 21, 1899. He received the B.Sc. degree in electrical engineering in 1921 from the Catholic University of America. From 1921 to 1924 he was an instructor in electrical engineering at the Catholic University, receiving the Ph.D. degree from that University in 1926. Mr. Rock has been associate professor of physics at the Catholic University since 1927, with research mainly in ultrasonics.

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Robert E. Rutherford was born in San Jose, California. After graduation from high school he attended the Pacific Radio School. In 1924 he took a position with the Magnovox Company where he did design and experimental work on radio receivers. In 1927 he joined the Farnsworth Television Company as a research engineer. Mr. Rutherford was sent to England for three months where he conducted tests with Baird Television, Ltd. Later he was sent to Berlin to experiment with television systems for Fernseh A.G. From 1937 to 1939 he was vice president of the Massachusetts Radio Institute. Since 1939 he has been associated with the Allen B. Du Mont Laboratories, Inc., as a television engineer.

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Robert E. Rutherford

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Jesse B. Sherman

Jesse B. Sherman (J'28-A'32) was born on February 8, 1910, at New York City. He received the B.S. degree in electrical engineering from Cooper Union Night School of Engineering in 1933; the B.S. degree in electrical engineering from New York University, Evening Engineering Division, 1935; and the M.E.E. degree from Polytechnic Institute of Brooklyn, 1938. Mr. Sherman was service manager of the Colen-Gruhn Company, Inc., from 1930 to 1935, and in the research and engineering department of the RCA Manufacturing Company, Inc., RCA Radiotron Division from 1935 to 1939. Since 1939 he has been an instructor in electrical engineering at The Cooper Union in New York City.

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M. A. Trainer (A'35) was born on July 25, 1905, at Philadelphia, Pennsylvania. He received the B.S. degree in electrical engineering from Drexel Institute in 1927. From 1927 to 1930 he was an assistant in Alexanderson's radio consulting laboratory at the General Electric Company. Mr. Trainer was in the engineering department of the RCA Victor Company from 1930 to 1935; since 1935 he has been with the RCA Manufacturing Company, Inc.

M. A. Trainer

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Harold Alden Wheeler (A'27, M'28, F'35) was born at St. Paul, Minnesota, on May 10, 1903. He received his B.S. degree in physics at George Washington University in 1925. From 1925 to 1928 he took postgraduate work in the physics department of Johns Hopkins University and was a lecturer there from 1926 to 1927. From 1921 to 1922 he was an assistant in the Radio Section of the National Bureau of Standards. In 1923 Mr. Wheeler became an assistant to Professor Hazeltine and in 1924 he entered the Hazeltine Corporation and the Hazeltine Service Corporation as an engineer. He is a member of Sigma Xi.

H. A. Wheeler

Hail Columbia

We join the nation in saluting Columbia Broadcasting System's new WABC 50KW transmitter, proud of the fact that DAVEN components are included in its construction.

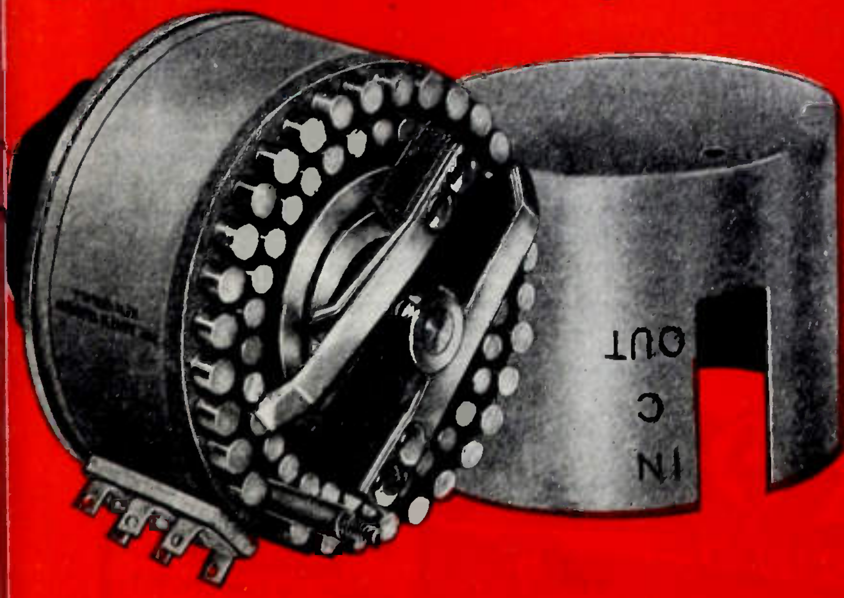
★ **UPPER LEFT** View of antenna rising from the transmitter on Columbia Island.

★ **UPPER RIGHT** A portion of the WABC audio facilities showing a DAVEN VI attenuator (to the right of the standard volume indicator in square instrument case) and DAVEN variable "T" type attenuators (in lower right panel, above jack fields).

★ **BOTTOM** Control console of the new WABC 50KW transmitter on Columbia Island, showing six DAVEN variable "T" type attenuators in the center, lower section of the panel.

Specify DAVEN ATTENUATORS

for high quality program distribution systems and as laboratory standards of attenuation.



The DAVEN catalog lists the most complete line of precision attenuators in the world; "Ladder," "T" type, "Balanced H" and Potentiometer networks—both variable and fixed types—as well as more than 80 models of Laboratory Test Equipment.

Where performance is the primary consideration, special heavy duty type switches for program switching and industrial applications, are available.

Many of your needs can be filled by DAVEN standard components. However, due to the specialized nature of high fidelity audio equipment, a large number of requirements are encountered where stock units may not be suitable. If you have such a problem, write to our engineering department.

THE DAVEN COMPANY

158 SUMMIT STREET • NEWARK, NEW JERSEY


DEFENSE
PLANT
PART OF THE
ARSENAL OF
DEMOCRACY

Science Shoulders Arms

What science in the past has done for peacetime America, needs no recounting.

But science today tackles a grimmer job. Research now must give America at war the fighting weapons she needs.

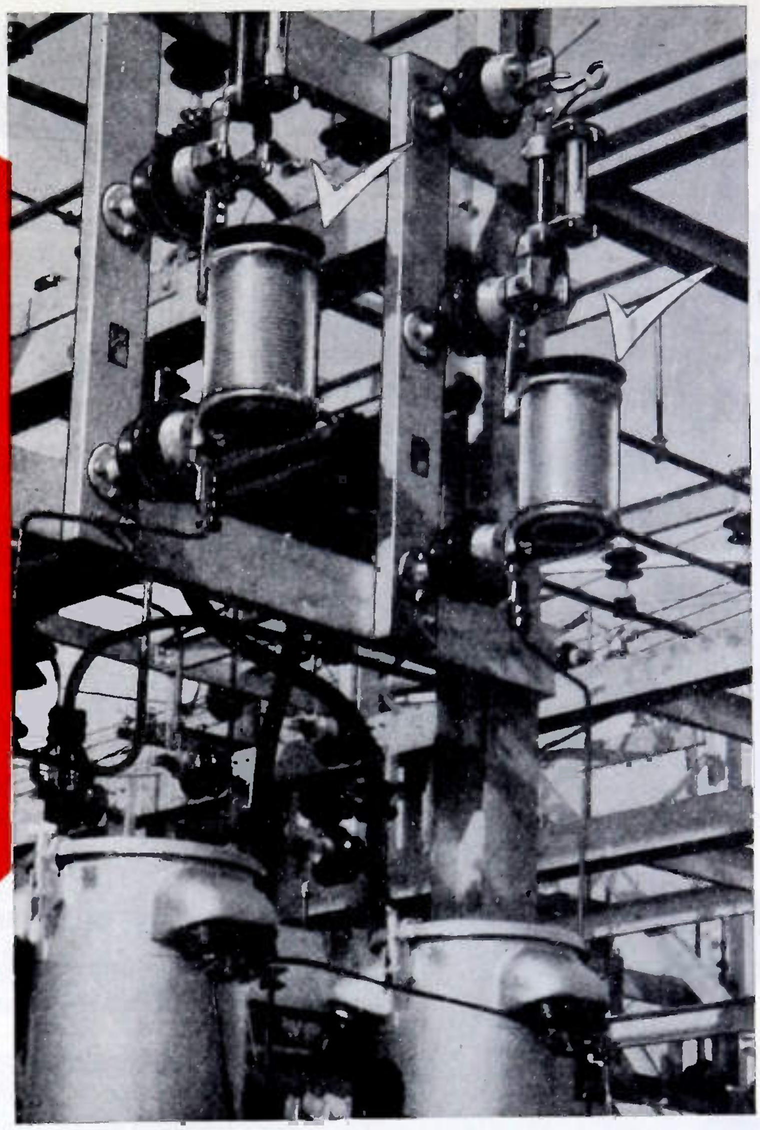
That's why Westinghouse Research Engineers—working in one of the world's leading laboratories—are devoting their full time to the nation's defense requirements. We wish we could tell you about some of the remarkable things they've done already.

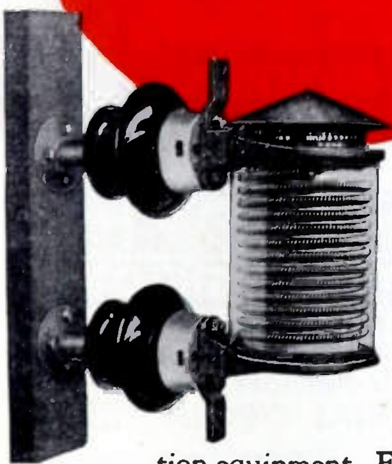
We shall—later.

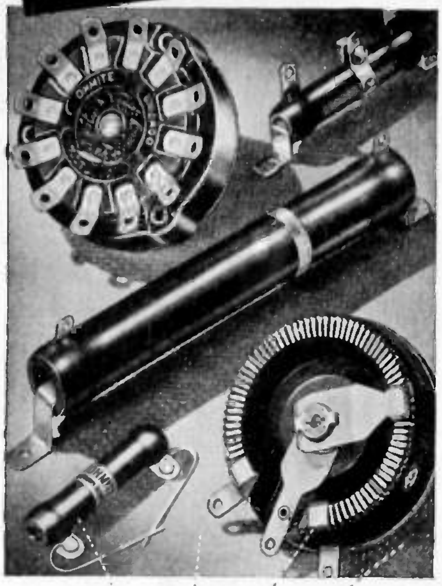
Westinghouse

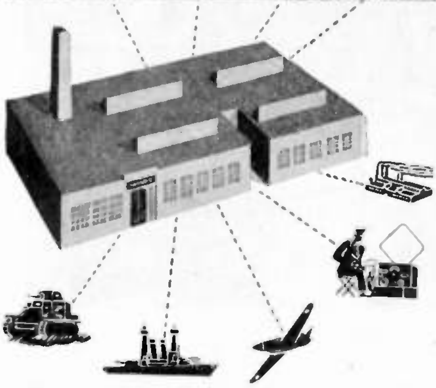
WESTINGHOUSE ELECTRIC & MANUFACTURING CO.

Pittsburgh, Pa.









Current Literature

New books of interest to engineers in radio and allied fields—from the publishers' announcements.
A copy of each book marked with an asterisk (*) has been submitted to the Editors for possible review in a future issue of the Proceedings of the I.R.E.

* THE RADIO AMATEUR'S HANDBOOK (1942), Nineteenth Edition). BY HEADQUARTERS STAFF OF THE AMERICAN RADIO RELAY LEAGUE. West Hartford: American Radio Relay League, Inc., 1941. 450+94 catalog+8 index papers, illustrated, $6\frac{1}{2}\times 9\frac{1}{2}$ inches. Paper: $1.00; buckram: $2.50. Spanish edition, $1.50.

* THEORY OF GASEOUS CONDUCTION AND ELECTRONICS. BY FREDERICK A. MAXFIELD, U. S. Naval Ordnance Laboratory, and R. RALPH BENEDICT, Assistant Professor of Electrical Engineering, University of Wisconsin. New York: McGraw-Hill Book Company, 1941. xiv+465+17 index pages, illustrated, 6×9 inches, cloth. $4.50.

* STANDARD HANDBOOK FOR ELECTRICAL ENGINEERS (Seventh Edition). BY A. E. KNOWLTON, Editor-in-Chief. New York: McGraw-Hill Book Company, August, 1941. xi+2278+25 index pages, illustrated, 6×9 inches, cloth. $8.00.

* SAFETY RULES FOR THE INSTALLATION AND MAINTENANCE OF ELECTRIC SUPPLY AND COMMUNICATION LINES. Washington: Superintendent of Documents, September, 1941. xix+177 pages, illustrated, $5\times 7\frac{1}{2}$ inches, cloth. 65 cents.

* RHOMBIC ANTENNA DESIGN. BY A. E. HARPER, Bell Telephone Laboratories. New York: D. Van Nostrand Company, Inc., 1941. xvi+108+3 index pages, illustrated, $8\frac{1}{2}\times 11$ inches, cloth. $4.00.

* RADIO TROUBLESHOOTER'S HANDBOOK (Second Edition). BY ALFRED A. GHIRARDI, Technical Consultant. New York: Radio & Technical Publishing Company, 1941. x+704+4 index pages, illustrated, $8\frac{1}{4}\times 10\frac{7}{8}$ inches, cloth. $3.50.

* PRINCIPLES OF ELECTRON TUBES. BY HERBERT J. REICH, Professor of Electrical Engineering, University of Illinois. New York: McGraw-Hill Book Company, 1941. xv+379+13 index pages, illustrated, 6×9 inches, cloth. $3.50.

* GASEOUS CONDUCTORS: Theory and Engineering Applications. BY JAMES DILLON COBINE, Assistant Professor of Electrical Engineering, Graduate School of Engineering, Harvard University. New York: McGraw-Hill Book Company, 1941. xix+593+12 index pages, illustrated, 6×9 inches, cloth. $5.50.

* FUNDAMENTALS OF VACUUM TUBES (Second Edition). BY AUSTIN V. EASTMAN, Associate Professor of Electrical Engineering, University of Washington. New York: McGraw-Hill Book Company, 1941. xx+569+13 index pages, illustrated, 6×9 inches, cloth. $4.50.

*ELECTRONICS. BY JACOB MILLMAN and SAMUEL SEELY, Department of Electrical Engineering, College of the City of New York. New York: McGraw-Hill Book Company, 1941. xxi+694+27 index pages, illustrated, 6×9 inches, cloth. $5.00.

* FIRST RADIO BOOK FOR BOYS. BY ALFRED MORGAN. New York: D. Appleton-Century Company, 1941. ix+187+4 index pages, illustrated, $5\times 7\frac{1}{2}$ inches, cloth. $2.00.

POSITIONS OPEN

The following positions of interest to I.R.E. members have been reported as open on December 20. Make your application in writing and address to the company mentioned or to

Box No. . . . .

PROCEEDINGS of the I.R.E., 330 West 42nd Street, New York, N.Y.

Please be sure that the envelope carries your name and address

TRANSMITTING-TUBE ENGINEER

We have a permanent position with a real opportunity for a man with these qualifications: complete knowledge of all phases of advanced engineering—research, design, and methods—on all types of transmitting and electronic radio tubes. All correspondence will be treated confidentially. Taylor Tubes, Inc., 2341 Wabansia Avenue, Chicago, Ill.

OUTSIDE PLANT ENGINEERS

The War Department has openings for civilian radio engineers and telephone engineers with practical experience on outside plant construction for work on fixed-station installation and maintenance. Salaries range from $2000 to $4800 per year, depending on experience and other qualifications. For additional information, apply to the Signal Officer of the Corps Area in which you live.

ELECTRICAL ENGINEER

Electrical engineers for design and development of electrical products such as battery chargers, transformers, electrical instruments, thermostats, electronic devices. Heyer Products Company, Inc., Belleville, N.J.

MONITORING OFFICERS

The U. S. Civil Service Commission will receive applications until January 30 for the positions of Radio Monitoring Officer and Assistant Radio Monitoring Officer, paying $3200 and $2600 per year, respectively. Information and application blanks are obtainable from the Secretary of Board of U. S. Civil Service Examiners at any first- or second-class post office or from the Commission at Washington.

ELECTRICAL ENGINEERS

Several electrical engineers are needed for the physics section of the research division of an eastern manufacturer (non-radio). Candidates must have a thorough grounding in electrical engineering with specialization in communications and electronics. Mathematical training desirable but not necessary. Box 268.

TRANSMITTER COMPONENTS DESIGNERS

Radio engineers wanted for development and design of radio-transmitter components and directional antenna equipment. Transmitter experience desirable although not necessary. Permanent positions. Describe education, experience, aptitudes, and state salary desired. Recent photo requested. E. F. Johnson Company, Waseca, Minn.

VACUUM-TUBE AND RECEIVER ENGINEERS

1) *Vacuum-Tube Development Engineer* having experience on design and manufacture of high-power transmitting vacuum tubes.

2) *Engineer* with experience on ultra high-frequency, especially on transmitter development.

3) *Receiver Design Engineer* familiar with aircraft receivers, preferably including ultra high frequencies. Experience with cathode ray tubes and circuits desirable.

4) *Receiver Engineer* experienced medium-band communication receivers of high gain. Experience with cathode ray tubes and circuits desirable.

Only American citizens need apply. International Telephone and Radio Laboratories, 67 Broad St., New York, N.Y.

RECEIVER ENGINEERS

A midwestern manufacturer has openings for 2 broadcast receiver engineers for permanent work on commercial and military a-m and f-m receivers. Box 259.

TUBE ENGINEERS

We have openings for engineers with experience in radio-tube manufacturing along any of the following lines:
(a) Research, (b) Design and development, (c) Product Engineering (development of factory processes, equipment, and materials, (d) Production engineering.
Apply by letter to Hygrade Sylvania Corporation, 500 Fifth Ave., New York, N.Y., attention: H. C. Richardson.

ELECTRICAL AND MECHANICAL ENGINEERS

A midwestern manufacturer of radio and mechanical equipment needs a number of electrical and mechanical engineers, juniors and seniors, for laboratory development work. Applicants for the former should have education and experience in radio or electronic applications. For the latter, education in machine design is required with experience on typewriter, calculating machine, or radio mechanical design desirable. Good personality and executive ability are prerequisites. Apply in writing to Engineering Department, The Crosley Corporation, Cincinnati, Ohio.

Attention Employers . . .

Announcements for "Positions Open" are accepted without charge from employers offering salaried employment of engineering grade to I.R.E. members. Please supply complete information and indicate which details should be treated as confidential. Address: "POSITIONS OPEN," Institute of Radio Engineers, 330 West 42nd Street, New York, N.Y.

The Institute reserves the right to refuse any announcement without giving a reason for the refusal.

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Engineers and Engineering Executives you are invited

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1942
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NEW YORK
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Morning, afternoon, and evening. The Show remains open until the start of the evening session on all three days.

SPECIAL NOTE—

to the Busy Engineer

This convention is being planned to help you keep in touch with the rest of the industry. Come in for at least one day —stay for all three if you possibly can. You'll go back with new ideas and a fresh perspective on your own job. Make up the time if you must, but come. You'll never regret it.

INSTITUTE OF RADIO ENGINEERS
330 WEST 42nd STREET
NEW YORK, NEW YORK

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2. DID YOU ATTEND COLLEGE? NO ☐ YES ☐ —AND RECEIVED— ☐ NO DEGREE ☐ BACHELOR'S DEGREE ☐ ADVANCED DEGREE
3. DID YOU ATTEND A RADIO SCHOOL? — ☐ YES ☐ NO
4. PLEASE DESCRIBE GENERAL NATURE OF
(A) WORK DONE BY YOUR FIRM (OR DEPARTMENT)
(B) YOUR OWN DUTIES

IF IN A BROADCAST STATION (STD., FM. OR TELEV.) GIVE CALL LETTERS

CONTINUE, IF NECESSARY, ON THE REVERSE SIDE

YEAR OF BIRTH — UNITS — TENS

YEAR OF ADMISSION — UNITS — TENS

PRINTED IN U. S. A.

TEAR ALONG DOTTED LINE

THE membership roster for the 1942 Year Book is now being compiled from the "Year Book Record Cards" that were mailed to all members (except Students) early in the Fall. If you haven't returned yours or if you have changed your address or professional affiliation without informing us of the details, you should fill out and mail the attached coupon promptly.

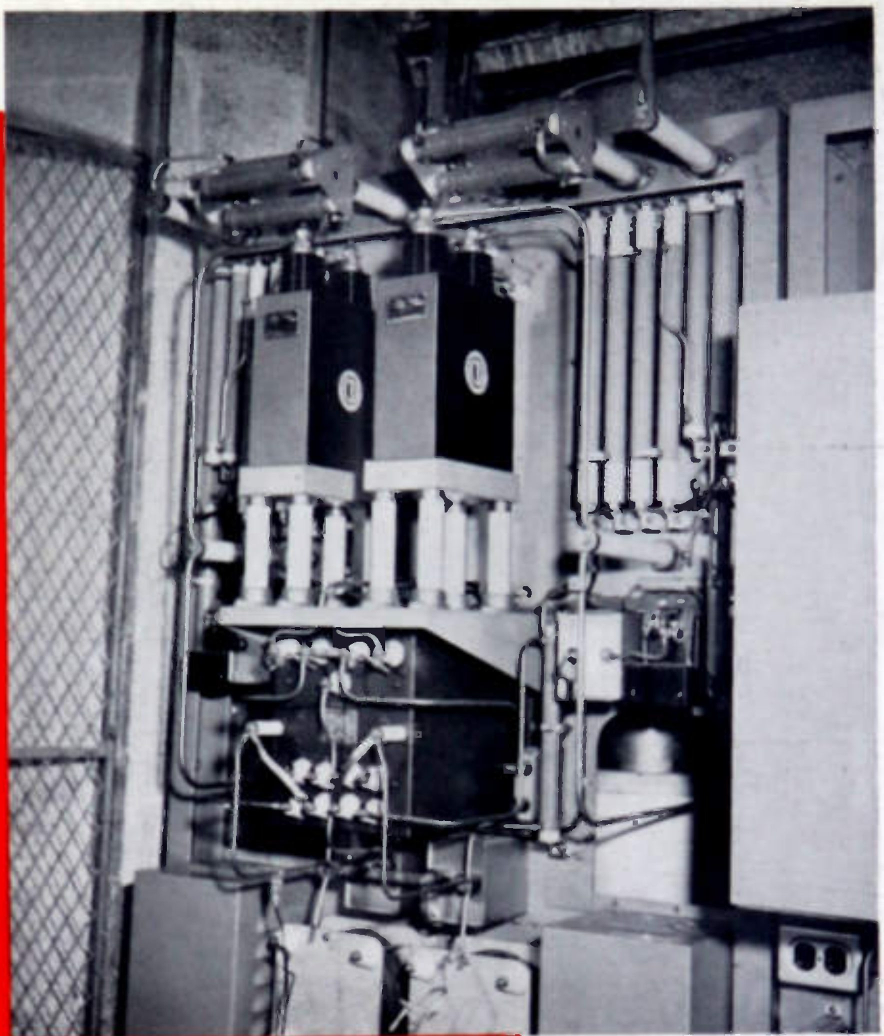
The lists will close on January 20, but corrections will be received on an "if possible" basis for several weeks thereafter.

●

The 1942 Year Book, the first since 1937, will be distributed in April to "paid 1942" members of all grades except Student. It will list the names, addresses, and professional affiliations of all members (except Students) in good standing on January 20.

Address:

INSTITUTE OF RADIO ENGINEERS
330 WEST 42ND STREET
NEW YORK, N.Y.



MAKING CONDENSERS

Fig. 1. The Obsolete Type 222-M Condenser

DIRECT READING

A VARIABLE AIR CONDENSER with semi-circular rotor and stator plates can be made to have remarkable linearity over about 80 per cent of a half turn. When used with calibration curves or charts, the accuracy obtained is so high that for many years manufacturers were discouraged from attempting to make condensers direct reading.

This phase of condenser development is now over. Most new condensers have direct-reading scales calibrated to an accuracy as good as was formerly obtained with calibration curves.

The first step in making a condenser direct reading was taken with the now obsolete Type 222 (Fig. 1). The worm was cut with double threads giving 12½ turns for ½ turn of the rotor. The number of plates were adjusted to make the capacitance increment per turn about 100 μμf. Ten turns (or 80 per cent of the available motion) would then correspond to 1,000 μμf.

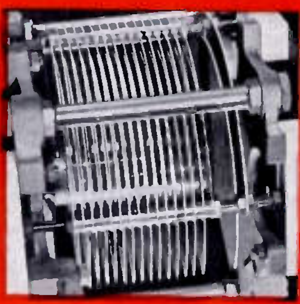


Fig. 2. Adjusting Plates of the Type 222-M

The scale markings were chosen to indicate capacitance taken out of the circuit. Adjusting plates were provided to make the capacitance per turn exactly 100 μμf (Fig. 2). Since the stator plates were supported at three points, the stator adjusting plate could be warped to make up for irregularities in the main stack.

With this construction it was possible to adjust the condenser so that it was direct reading in capacitance difference from the zero mark with an accuracy of 1 μμf or 0.1%, whichever was greater.

Fig. 3. The Improved Type 722 Condenser

The Type 722 Precision Condenser (Fig. 3) was developed as an improvement on the Type 222. Most of the changes . . . ball bearings . . . integral-cut worm . . . cast-aluminum frame . . . worm shaft at right angles to the panel . . . have no immediate bearing on the direct-reading problem.

In the Type 722-D the function of the drum and dial are transposed (Fig. 4). Twenty-five turns of the worm produce a half turn of the rotor plates. The dial is divided into 250 divisions; the usable portion of the condenser then has 5,000 divisions; one μμf covers 5 divisions on the 1,000 μμf condenser.

The stator plate at the right of the stack (Fig. 5) is used to make the capacitance per turn exactly 50 μμf. Since only two stator supports are used, this plate can be tipped to correct for slight irregularities in the main stack. As this plate cannot be warped, a special stator plate, cut out in the middle, is used at the left end of the stack (Fig. 4). This plate increases the capacitance per turn at the ends.

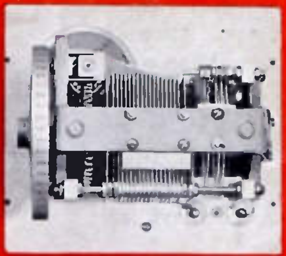
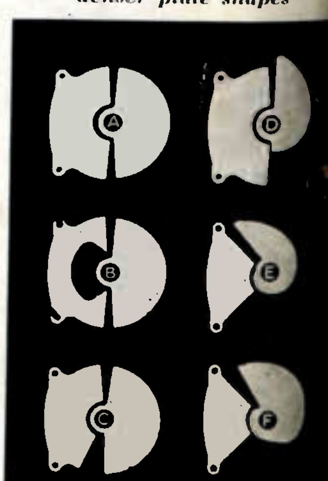


Fig. 4. Dual Sections of the Type 722-D

Zero capacitance is altered by bending the flat plate which extends from the frame. By means of these various adjustments the large section of this condenser is made direct reading in total capacitance to 1 μμf or 0.1% between 100 μμf and 1,000 μμf. A small section is provided also. This has one-tenth the capacitance of the larger. It is adjusted by similar means to be direct reading in total capacitance to 0.2 μμf or 0.1% between 25 and 100 μμf.

By appropriately shaping the rotor and stator plates, these precision condensers can be adapted to use in a large number of direct-reading instruments. Plates A (Fig. 6) are standard semi-circular plates; stator B is the compensating plate used to increase the capacitance per turn at the ends of the calibration; stator C is used in the low capacitance stack to decrease zero capacitance. Pair D give a logarithmic scale over a three-to-one range when used in a tuned oscillator; pair E will give a linear scale for frequency, and a semi-logarithmic scale for capacitance. Plates F are used in a beat-frequency oscillator to give a scale covering three decades.

Fig. 6. Some typical condenser plate shapes



The fundamental mechanical and electrical problems in making condensers direct reading having been solved, it is now possible to design a condenser which can be made direct reading in almost any one of the many related quantities which the condenser may control in a circuit or an instrument.

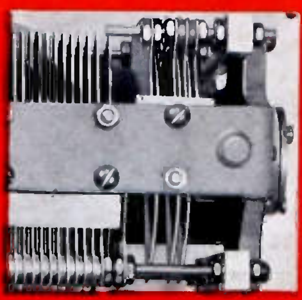


Fig. 5. Special stator plate and low-capacitance section of the Type 722-D Condenser

GENERAL RADIO COMPANY CAMBRIDGE MASSACHUSETTS